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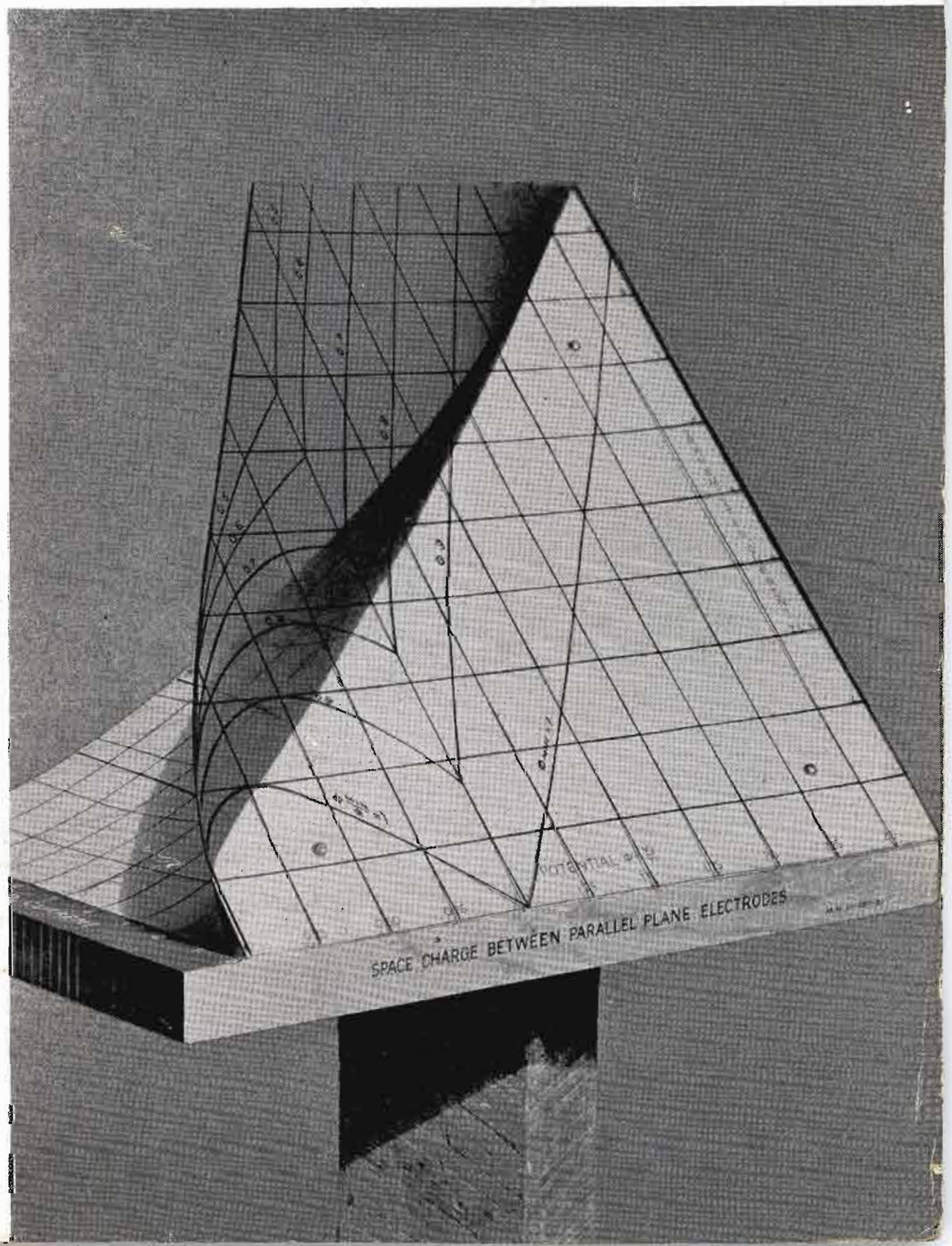

JUNE
1939

WABY	WCOL	WHMA	KWOS
WKBB	WCOV	WGNC	KWBG
WATR	WBLK	WGAU	KTSW
WAPO	WBTH	WHAI	KRBA
WNLC	KFXR	KANS	KVGB
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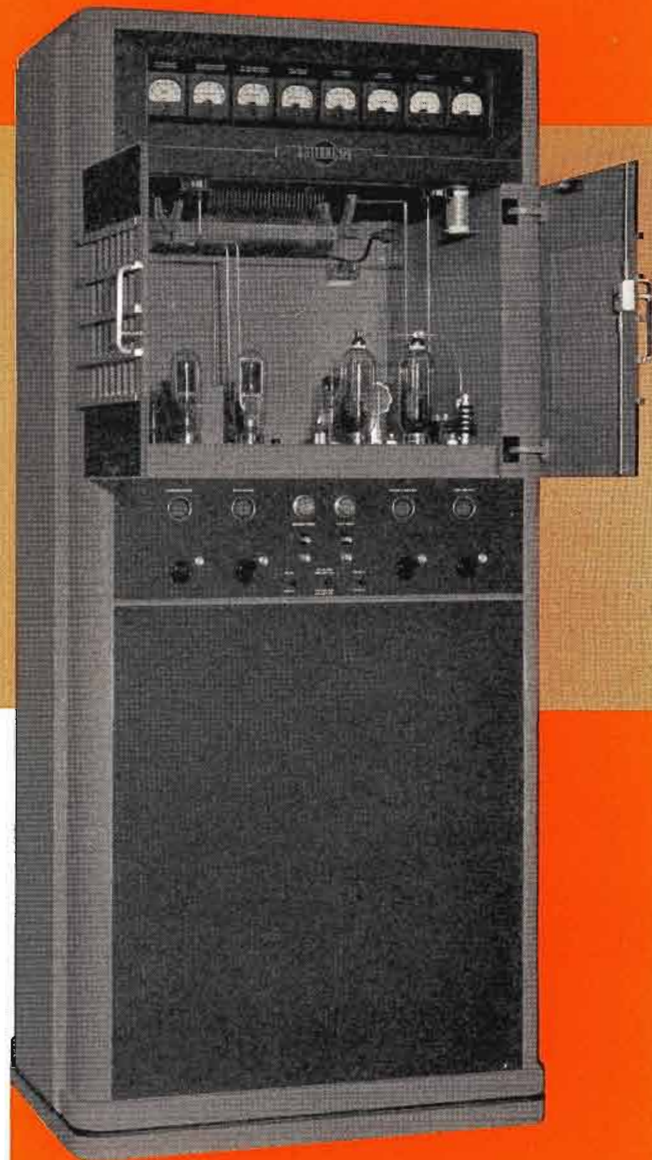
WIBU	KGFW	KOCA	KOVC



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VOLUME 19
NUMBER 6

RAY D. RETTENMEYER
Editor

• Editorial Comment •

BEGINNING this month, conventions seem to be in order. Since these events will cover nearly every phase of radio, we list them briefly here.

From June 13 through 17, the Fifteenth Annual RMA Convention and the National Radio Parts Trade Show will be held at the Stevens Hotel in Chicago. Considerable interest is being evidenced in these gatherings which are likely to set a new record in attendance.

Next in line is the Pacific Coast Annual Meeting of the Institute of Aeronautical Sciences. This meeting is scheduled for June 15, 16, and 17, the place being the California Institute of Technology at Pasadena, California.

Also on the Pacific Coast is the Convention of the Institute of Radio Engineers to be held at San Francisco, California, on June 27, 28, 29 and 30. The program of this meeting will be found elsewhere in this issue.

Last but far from least is the Annual Convention of the National Association of Broadcasters. This gathering takes place at the Hotel Ambassador, Atlantic City, N. J., from July 10 through 13. This meeting promises to be of particular significance. Further details will be published at a later date.

THE FCC's Television Committee has made its first report to the Commission. This lengthy document deals with the standards proposed sometime ago by the RMA.

Briefly, the Committee recommended that the Federal Communications Commission neither approve nor disapprove the RMA standards, since the Committee felt that such action might discourage private enterprise or decrease the incentive for undertaking research. It was also suggested that the proposed standards did not appear suitable for the undeveloped higher frequency channels.

Another suggestion of the Television Committee was that the FCC adopt a policy of cooperation with the industry as a whole and that it immediately arrange a procedure by which it can keep abreast of current developments in television.

• Contents •

COVER ILLUSTRATION

A model devised by engineers of the Bell Telephone Laboratories to illustrate the mathematical relationships between current and potential in the space between grid and plate of a vacuum tube.

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COMMUNICATIONS

FOR JUNE, 1939

IMPEDANCE MEASUREMENTS

On Broadcast Antennas

Part I

By **DONALD B. SINCLAIR**
GENERAL RADIO CO.

This paper was presented before the Second Annual Broadcast Engineering Conference which was held at the Ohio State University, Columbus, Ohio, from February 6 through 17, 1939.

1. INTRODUCTION

THE fundamental problem in which the broadcast engineer is interested is the production of maximum useful power output with minimum power input, a problem which involves not only over-all station efficiency but also antenna characteristics.

The antenna must, first, accept the available transmitter power; it must, second, radiate the power; and it must, third, radiate it where it is wanted. The first characteristic requires correct tuning of the antenna and associated feeding and matching circuits. The second and third characteristics require proper antenna design.

The ultimate test of a broadcast antenna comes from the field strength survey, which measures the final product, namely the received signal. A complete three-dimensional field-strength survey, which includes not only measurements of coverage but also of vertical directivity, theoretically yields both total radiated power and spatial distribution. The labor of obtaining the total radiated power by this method, however, is very great, and the cost of the necessary measurements is prohibitive.

In most well-designed broadcast antennas the difference between the power supplied to the antenna and that actually radiated is not great. It is therefore common practice to use field-strength surveys mainly to determine coverage and to rely on power measurements at antenna to indicate radiated power.

The directivity of a simple antenna is largely a function of its geometry and of the neighboring terrain. That of an array depends, in addition, upon the relative magnitudes and phases of the currents in the array elements. Provided the currents are properly maintained, the directive pattern changes only slightly with time. Once the proper directional characteristics of an array have been established by means of field-strength measurements, the radiating system can be maintained in proper adjustment by maintaining correct power input to the antenna and correct phase relations between the currents. Measurements on the antenna, rather than measurements on the field, are therefore as useful for arrays as for simple antennas and are even more desirable because of the speed, convenience and accuracy with which they can be made. For the same reasons they are of great importance in the lining up and testing of new radiating systems.

The major emphasis in measurements on antennas is laid upon power measurements. Since no entirely satisfactory method of measuring power directly at radio frequencies has yet been devised, power is ordinarily deduced from measurements or resistance and of current. The quantities that must be measured in order to adjust and maintain a radiating system are, therefore: (1) Impedance: (a) Resistance, (b) Reactance; (2) Current: (a) Magnitude, (b) Phase.

The problem of impedance measurement has received considerable attention, and many specific solutions have been evolved. It is the purpose of this article to attempt to collect the various methods in one place and to correlate them to provide a general solution.

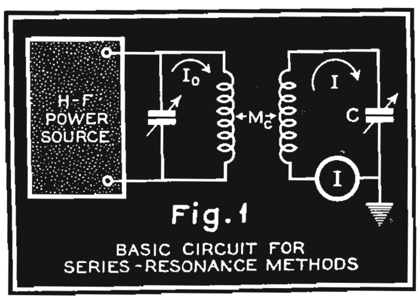


Fig. 1
BASIC CIRCUIT FOR SERIES-RESONANCE METHODS

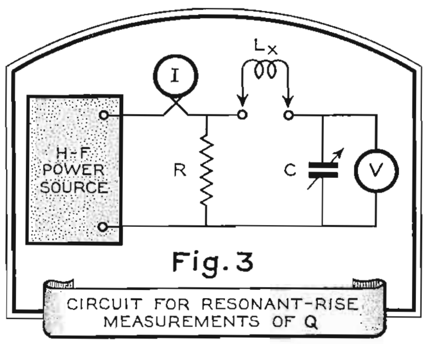


Fig. 3
CIRCUIT FOR RESONANT-RISE MEASUREMENTS OF Q

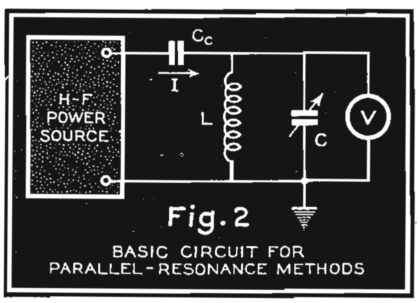


Fig. 2
BASIC CIRCUIT FOR PARALLEL-RESONANCE METHODS

II. METHODS OF MEASURING IMPEDANCE AT RADIO FREQUENCIES

Classification

Two general classes of measurement methods are commonly used at radio frequencies, namely resonance methods and null methods. These methods are similar, insofar as they measure an unknown impedance by comparison with a known impedance, but they differ fundamentally in the means of indication. Resonance methods in general depend upon tuning resonant circuits to a voltage or current maximum while null methods, as the name implies, depend upon balancing to a voltage or current null.

Various specific types of measuring methods have been developed under the two general classifications. The major point of difference between these methods is found in the manner of determining the resistive component. Some of the most common methods are collected in the following tabulation:

I. Resonance Methods:

(1) Series-Resonance Methods: (a) Substitution, (b) Resistance-Variation, (c) Reactance-Variation.

(2) Parallel-Resonance Methods: (a) Substitution, (b) Conductance-Variation, (c) Susceptance-Variation.

(3) Voltmeter-Ammeter Methods: (a) Resonant-Rise Method.

II. Null Methods:

(2) Bridge Methods: (a) Impedance Bridge, (b) Schering Bridge.

(2) Double-T Methods.

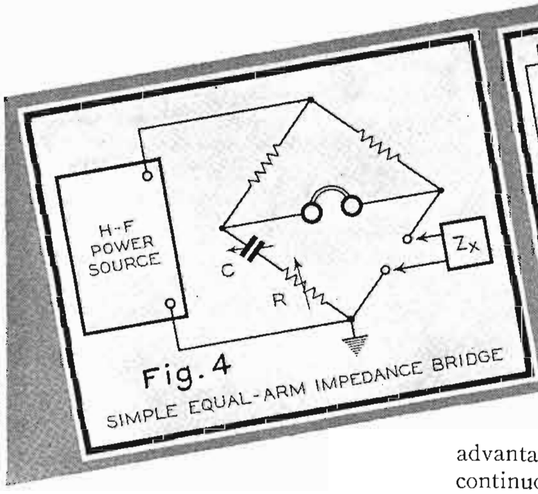


Fig. 4 SIMPLE EQUAL-ARM IMPEDANCE BRIDGE

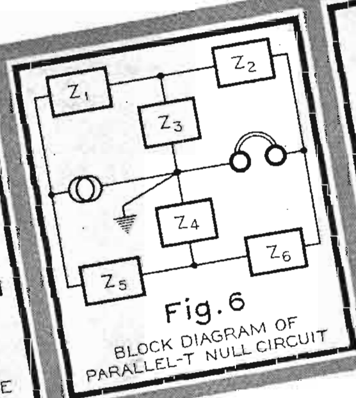


Fig. 6 BLOCK DIAGRAM OF PARALLEL-T NULL CIRCUIT

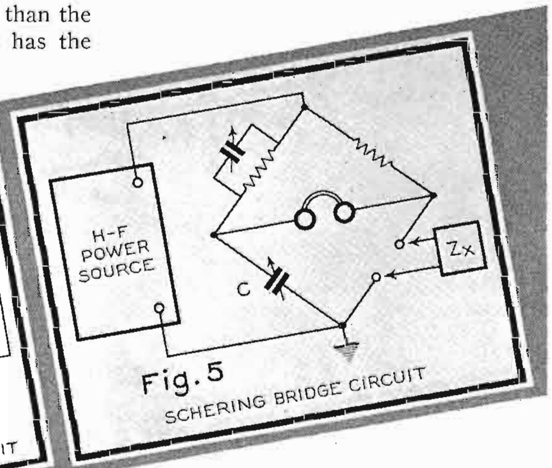


Fig. 5 SCHERING BRIDGE CIRCUIT

Series-Resonance Methods

General Features

The series-resonance methods depend upon the measurement of currents in series-resonant tuned circuits. They are best adapted for the measurement of low impedances because they are primarily methods for determining the series resistances of series-resonant tuned circuits. The basic circuit used is shown in Fig. 1.

Substitution Method

Of the three common series-resonance methods, the most straight-forward is the substitution method. The elementary procedure is simple. First, the unknown impedance is connected in series with the series-tuned circuit and the circuit tuned to resonance. Next, the unknown impedance is removed from circuit and replaced by a continuously adjustable non-reactive resistor. Finally, the circuit is re-tuned to resonance and the resistor varied until the original value of current is restored. The reactance of the unknown is equal to the change in the condenser reactance between the two measurements, and the resistance of the unknown is equal to the setting of the standard resistor.

Resistance-Variation Method

The resistance-variation, or added-resistance, method is less direct than the simple substitution method but has the advantage that it does not require a continuously adjustable standard resistor. For this method the mutual inductance coupling, M_c, is made so small that tuning the measuring circuit causes no appreciable reaction on the frequency or amplitude of the current, I_0, in the high-frequency power source tuned circuit. The voltage, $j\omega M_c I_0$, induced in the measuring circuit, is therefore practically constant, independent of changes in the series impedance of the measuring circuit, and the resonant current, I, in the measuring circuit is inversely proportional to the series resistance.

The procedure when measuring an unknown impedance is as follows. First, the output of the high-frequency power source is adjusted until the resonant current in the measuring circuit is nearly full-scale on the meter. Next, a non-reactive resistor is placed in series with the measuring circuit and the corresponding resonant current noted. From the inverse relationship between current and resistance the circuit resistance can be deduced.

The same procedure is now repeated with the unknown impedance included as part of the series tuned circuit. The difference between the two values of circuit resistance, the one for the circuit alone and the other for the circuit plus the unknown, is equal to the resistance of the unknown, while the difference between the two values of condenser reactance necessary to maintain resonance is equal to the reactance of the unknown.

Reactance-Variation Method

The reactance-variation method does not require a resistance standard at all. The circuit resistance is deduced, instead, from the breadth of the resonance curve.

The procedure, when measuring an unknown impedance, is similar to that followed with the resistance-variation method. The coupling mutual inductance, M_c, is first made so small that variations of the measuring circuit impedance do not appreciably affect the power source. A resonance curve of current as a function of capacitance is then taken. With the unknown impedance connected in series, a second resonance curve is taken. From each resonance curve the corresponding series resistance of the tuned circuit can be deduced. As before, the difference between the two measured resistances is necessarily equal to the resistance of the unknown and the difference between the two values of resonant condenser reactance is equal to the reactance of the unknown.

Parallel-Resonance Methods

General Features

The parallel-resonance methods depend upon the measurement of voltages across parallel-resonant tuned circuits. They are best adapted for the measurement of high impedances because they are primarily methods for determining the parallel conductances of parallel-resonant circuits. The basic circuit used is shown in Fig. 2.

Substitution Method

As in the case of the series-resonance methods, the most straightforward of the three common parallel-resonance methods is the substitution method. The elementary procedure is essentially the same. First, the unknown impedance is connected in parallel with the parallel-tuned circuit and the circuit tuned to resonance. Next, the unknown impedance is removed from circuit and replaced by a continuously adjustable non-reactive resistor. Finally, the cirsuit is retuned to resonance and the resistor is varied until the original value of voltage is restored. The susceptance of the unknown is equal to the change in the condenser susceptance between the two measurements, and the conductance of the unknown is equal to the reciprocal of the setting of the standard resistor.

Conductance-Variation Method

The conductance-variation method is less direct than the substitution method but has the same advantage as has the resistance-variation method, namely that it does not require a continuously adjustable standard resistor. For this method, the capacitive coupling, C_c, is made so small that tuning the measuring circuit causes no appreciable reaction on the frequency or amplitude of the current, I. The resonant voltage, V, developed across the measuring circuit is then inversely proportional to the effective parallel conductance.

The procedure when measuring an unknown impedance by means of the conductance-variation method is as follows. First, the output of the high-frequency power source is adjusted until the resonant voltage developed accross the measuring circuit is nearly full-scale on the meter. Next, a non-reactive resistor is placed in parallel with the measuring circuit and the corresponding resonant voltage noted. From the inverse relationship between voltage and conductance the circuit conductance can be deduced.

The same procedure is repeated with the unknown impedance included as part of the parallel tuned circuit. The difference between the two values of circuit conductance, the one for the circuit alone and the other for the circuit plus the unknown, is equal to the susceptance of the unknown.

Susceptance-Variation Method

The susceptance-variation method, like the reactance-variation method, does not require a resistance standard at all. The circuit conductance is deduced, instead, from the breadth of the resonance curve.

The procedure, when measuring an unknown impedance, is similar to that followed with the conductance-variation method. The coupling capacitance, C_c, is first made so small that variations of the measuring circuit admittance do not appreciably affect the power source. A resonance curve of voltage as a function of capacitance is then taken. With the unknown impedance connected in parallel, a second resonance curve is taken. From each resonance curve the corresponding parallel conductance of the circuit can be deduced. As before, the difference between the two measured conductances is necessarily equal to the conductance of the unknown, and the difference between the two values of resonant condenser susceptance is equal to the susceptance of the unknown.

Voltmeter-Ammeter Methods

Resonant-Rise Method

The simple resonance methods that have been discussed so far involve the measurement of current or voltage in a single branch of a circuit. Another class of resonance methods involves the measurement of currents or voltages in more than one branch. One particular circuit that falls into this category has been exploited commercially under the name Q-Meter. The elementary circuit is shown in Fig. 3.

The resistance, R, in this circuit is made so small that variations in the tuning of the measuring circuit do not cause appreciable variations in the voltage developed across it by the source current, I. When the measuring circuit is tuned to series resonance with an unknown coil, L_x, connected as shown, the ratio of the voltage, V, across the condenser to the input voltage, IR, is equal to the ratio of the condenser reactance to the effective series resistance around the measuring circuit. If the losses in the condenser, C, and the voltmeter, V, are made negligibly small compared with those in the unknown coil, L_x, the voltage ratio becomes equal to the storage factor, Q, of the unknown coil. Provided the source current, I, is maintained at a fixed predetermined value, the voltmeter, V, can be calibrated to read values of Q directly. The reactance of the unknown coil, L_x, is equal and opposite to that of the condenser, C, and can be easily found from the dial reading.

In order to measure arbitrary unknown impedances with this circuit, two measurements must be made, as with the other resonance methods, one with the unknown in circuit and one with it out of circuit.

Bridge Methods

Impedance Bridge

Probably the simplest form of null method that can be used is the equal-arm impedance bridge. The elementary circuit is shown in Fig. 4.

Balance, in this bridge, necessarily occurs when the impedances in the lower left-hand and right-hand arms are equal. At balance the resistance of the unknown impedance is equal to that

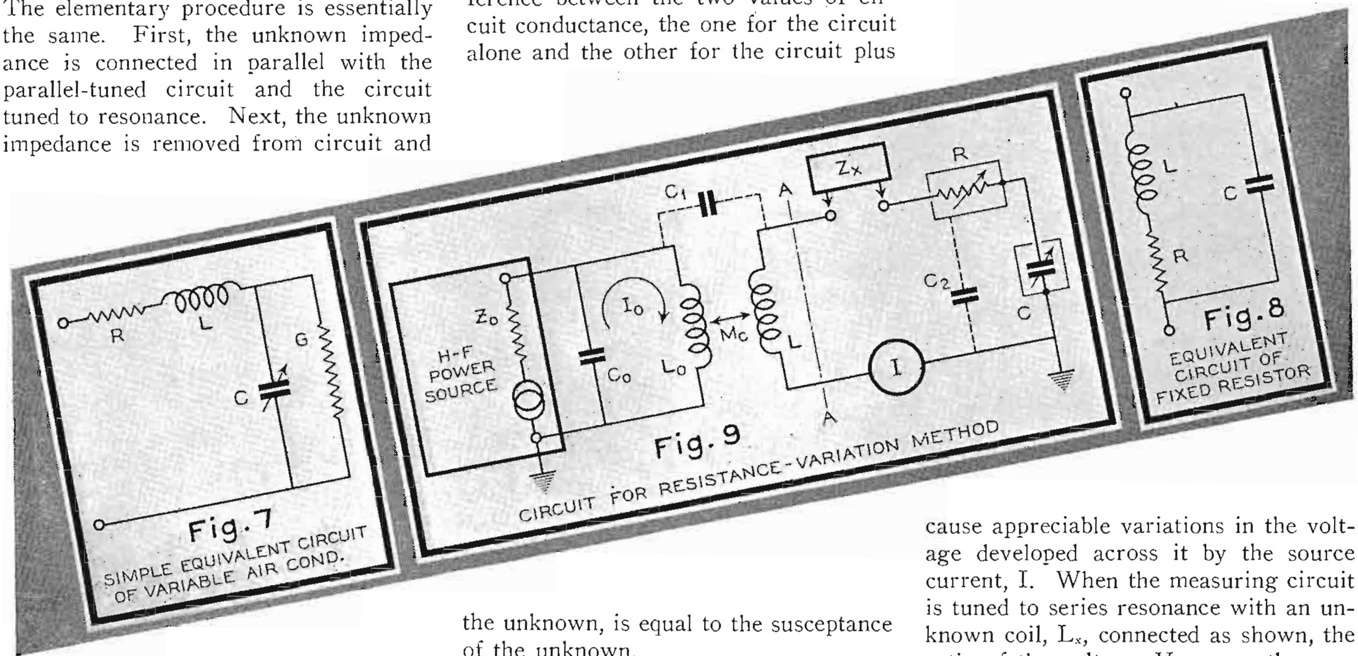


Fig. 7 SIMPLE EQUIVALENT CIRCUIT OF VARIABLE AIR COND.

Fig. 9 CIRCUIT FOR RESISTANCE-VARIATION METHOD

Fig. 8 EQUIVALENT CIRCUIT OF FIXED RESISTOR of the continuously adjustable resistor, R, while the reactance of the unknown impedance is equal to that of the variable condenser, C.

This bridge can easily be arranged for substitution measurements. A suitable impedance is first connected in the lower right-hand bridge arm and the bridge is balanced. The unknown impedance is then placed in series with the lower left-hand bridge arm, and the bridge rebalanced with the resistor, R, and condenser, C. The changes in settings of these standards are measures of the resistive and reactive components of the unknown impedance.

Schering Bridge

The Schering bridge differs from the ordinary impedance bridge in the manner of determining the resistive component. The basic circuit is shown in Fig. 5.

For this bridge, balance occurs when the reactance of the standard condenser, C, is equal to the unknown reactance and when the storage factor, Q, of the left-hand upper ratio arm is equal to the dissipation factor, D_x, of the unknown impedance. The Schering bridge has the advantage that it does not require a continuously adjustable resistor but has a compensating disadvantage for resistance measurements in the fact that it reads dissipation factor, rather than resistance.

Double-T Methods

Double-T null circuits have not yet received as much attention as the better-known bridge circuits for measuring purposes but will probably be more widely exploited in the future. The basic circuit is illustrated in Fig. 6.

The balance conditions for this circuit were discussed by Dr. W. N. Tuttle of the General Radio Company in a paper delivered at the Annual Meeting of the Institute of Radio Engineers in June, 1938. They can be satisfied with independent adjustments of resistive and reactive components, and circuits can consequently be devised which can be made direct-reading in these quantities.

III. SOURCES OF ERROR IN IMPEDANCE MEASUREMENTS

Standard Impedances

General

In the preceding discussion of methods of measurement is was tacitly assumed that the various component circuit elements were perfect and that the wiring impedances were negligible. Unfortunately, neither of these assumptions is true. In fact the magnitudes of unwanted residual parameters in the standards and in the wiring usually determine the accuracy and range of measurements that can be made with any given circuit.

Residual parameters in the impedance standards used are particularly important. They arise because it is physically impossible to fabricate a circuit element that has but one of the three circuit parameters—inductance, resistance and capacitance. Reactance standards inevitably contain residual resistance components, and resistance standards contain residual reactance components.

Not only do unwanted residual parameters occur, but the magnitudes of both the main and residual parameters change with frequency because of current redistribution, or skin-effect. The choice of a suitable standard for use at high frequencies must take into account all these factors.

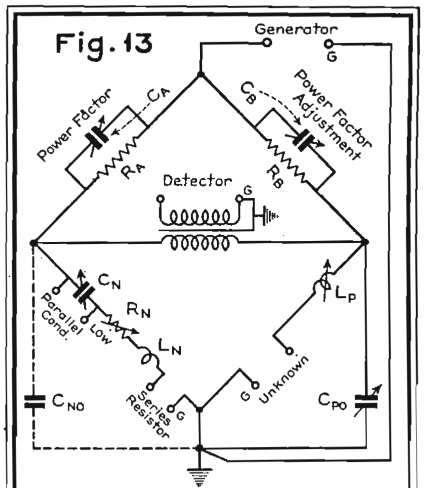
Variable Condenser

The variable condenser has been found, in general, the best reactance standard at radio frequencies. The residual inductance and resistance are very small compared with the capacitance, and skin-effect has practically no effect upon the capacitance.

An approximate equivalent circuit for the variable condenser, which shows the residual parameters, is given in Fig. 7.

The conductance, G, represents loss in the dielectric statorstack supports; the resistance, R, represents loss in the metal structure; and the inductance, L, corresponds to magnetic flux set up by conduction currents in the metal structure.

For the General Radio Type 722-N



RADIO-FREQUENCY BRIDGE CIRCUIT

precision condenser, which has been especially designed for high-frequency service, the orders of magnitude of these parameters at a frequency of 1,000 kc are as follows:

L = 0.024 microhenry
R = 0.008 ohm
G = 0.3 micromho 3 (megohms)

These residuals are so small that at broadcast frequencies they do not cause appreciable errors.

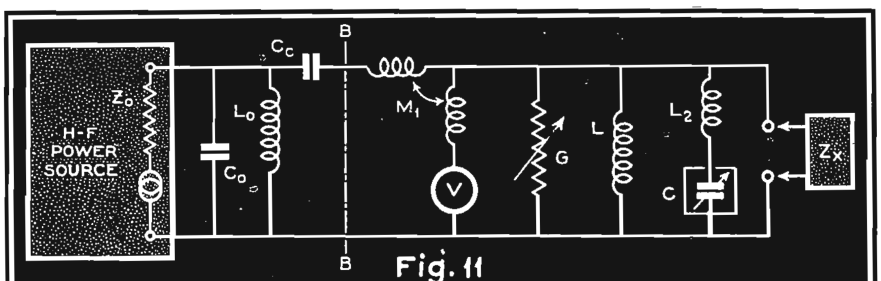
Resistors

The problem of obtaining a resistance standard at radio frequencies is not as easily solved as is the problem of obtaining a reactance standard. Fixed resistors can be made that have reasonably small residual parameters and skin-effect, and that can be used over a wide frequency range. When these fixed resistors are connected into a switching assembly, in order to form a decade resistor, however, the residuals become important, and the upper limit of the frequency range is greatly reduced.

The effects of the residual parameters depend upon the frequency and the value of the resistance. A simple equivalent circuit for a resistor having residual inductance and capacitance is shown in Fig. 8.

The residual parameters not only cause a reactive component to appear in the impedance but also cause the resistive component to vary. For instance, if the resistance, R, is high, the residual capacitance, C, causes the resistive component to decrease with frequency because of the shunting action. If the resistance, R, is low, the residual capacitance, C, and inductance, L, resonate and the resistive component rises as the resonant frequency is approached.

For the range of resistance values normally used in broadcast antenna measurements, namely 0—100 ohms, the variation in the resistive component caused by residual parameters is ordi-



CIRCUIT FOR CONDUCTANCE-VARIATION METHOD narily small, and the more important item is the variation in inductive reactance. As was previously mentioned, the residual inductance can not be made negligible. The next best thing is to make it constant.

The General Radio Type 670 compensated decade resistor was designed to satisfy this requirement. The method used is to switch into circuit compensating copper coils when resistance cards are switched out and vice versa. By careful matching of inductances the over-all inductance of the box is maintained constant within 0.05 microhenrys.

A small variation of resistance with frequency occurs in the broadcast band because of this construction. At the zero setting of the box, the entire resistance is in the copper compensating coils and wiring. At d-c this resistance is very small, of the order of 0.05 ohms. It rises relatively rapidly with frequency, however, because of skin-effect. The resistance of the box at maximum setting occurs almost entirely in the resistance cards since the compensating coils are cut out of circuit. Skin-effect in these cards is very small, and the maximum resistance consequently remains nearly constant. As the frequency rises, the zero resistance therefore tends to approach the maximum resistance. The resultant decrease in incremental resistance gives rise to an apparent negative skin-effect for changes of resistance as read from the dials, which amounts to approximately —0.8% for the units decade and —0.6% for the tens decade at 1 mc.

Circuits

General

Errors occur not only in the standard impedances with which the unknown impedances are compared, but also in the circuits used to effect the comparisons. These errors can arise from many causes, such as residual parameters in the wiring, deviations in practical setups from the theoretical assumptions, and lack of precision in making readings. While similar in general, the errors that arise in the several types of measuring methods are markedly different in particular and are best treated individually in connection with each specific method.

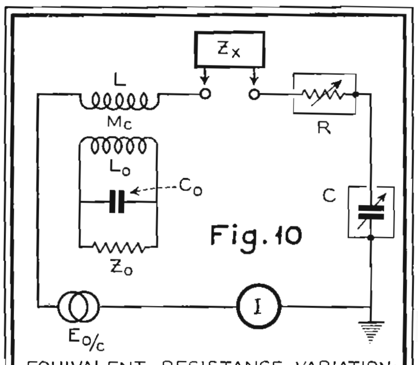


Fig. 10

EQUIVALENT RESISTANCE-VARIATION CIRCUIT SHOWING EFFECT OF CLOSE COUPLING

Series-Resonance Methods

The series-resonance methods have two important sources of error.

1. Residual parameters in wiring,
2. Close coupling to the high-frequency power source.

The first of these sources arises principally because of undesired residual capacitances. The second arises because of a discrepancy between practical setups and the usual simple theory. To illustrate the nature of the errors, the resistance-variation method serves as an excellent example.

A commonly used circuit is given in Fig. 9.

Two residual parameters are shown, namely the capacitances C_1 and C_2. These are by no means all, nor necessarily the most important ones that occur, but they are typical of those found in series-resonant methods.

The capacitance, C_1, feeds energy from the power source into the measuring circuit in a manner not considered in the simple theory. The current through C_1 splits in the measuring circuit, unequal parts returning to ground through the coil and ammeter and through the resistor and condenser. The current in the measuring circuit is consequently not everywhere the same, and error occurs if the simple theory is applied. A grounded electrostatic shield is often interposed between the power source and the measuring circuit to avoid errors arising from this cause.

The capacitance, C_2, shunts the standard condenser, C, and causes error in the determination of reactance. A concrete example shows the nature of this error.

Suppose that $C_2 = 10$ mmfd and that the two settings of the standard condenser with the unknown impedance, Z_x, in and out of circuit are 300 mmfd and 250 mmfd. The reactance difference, as read from the condenser dial, is

$$\Delta X = \frac{10^{12}}{\omega}\left(\frac{1}{250} - \frac{1}{300}\right)$$

$$= \frac{10^{12}}{\omega} \times \frac{50}{75{,}000} \text{ ohms.}$$

The true reactance difference is

$$\Delta X = \frac{10^{12}}{\omega}\left(\frac{1}{260} - \frac{1}{310}\right)$$

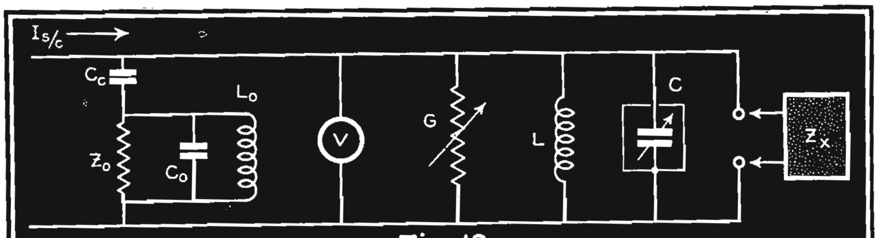


Fig. 12

EQUIVALENT CONDUCTANCE-VARIATION CIRCUIT SHOWING EFFECT OF CLOSE COUPLING

$$= \frac{10^{12}}{\omega} \times \frac{50}{80{,}600} \text{ ohms.}$$

and the error caused by neglecting C_2 is about 7%.

The effect of close coupling can be easily analyzed if the coupling does not affect the source frequency appreciably and if the source output impedance is essentially linear. By Thévenin's theorem, that part of the circuit which lies to the left of the section A-A in Fig. 9 can be replaced by an equivalent circuit which consists of a constant voltage, $E_o/_c$, equal to the open-circuit voltage at the section A-A, in series with a constant impedance equal to that seen looking toward the high-frequency power source from the section A-A with all active voltages zero.

Fig. 9 can therefore be redrawn as shown in Fig. 10.

If the coupling is infinitely loose, as assumed in the simple theory, the circuit resistance that is measured is simply the series resistance of the measuring circuit. If the coupling is finite, a certain amount of resistance is coupled into the measuring circuit from the power source and the circuit resistance measured includes a component that can be ascribed to close coupling. Provided, however, that the assumptions of constant frequency and linear output impedance are satisfied, and that the coupling is held constant, no error need arise from close coupling when an unknown impedance is measured. If determinations of circuit resistance are made with the unknown impedance, Z_x, in and out of circuit, the difference in circuit resistances will be equal to the resistance of the unknown. The difference between the two resonant values of condenser reactance will be equal to the reactance of the unknown.

Parallel-Resonance Methods

The parallel-resonance methods are subject to the same types of error as those that occur in the series-resonance methods. The residual parameters

(*Continued on page* 26)

Automatic Threshold Control

for radio telegraph and telephone receivers

By LEE HOLLINGSWORTH

THERE are many variations of quiet tuning and noise squelching circuits, all of which seem to operate to produce the same general effect, that of rendering the circuit comparatively free from noise impulses when the carrier is removed. The method of quiet tuning described herein offers perhaps the widest application and is believed to produce the most satisfactory results, both for telephone and telegraph reception, of any system of quiet tuning yet brought forth. Advantages other than quiet tuning are also explained.

APPLICATION TO RADIO TELEGRAPH RECEPTION

Muting in the radio and intermediate amplifiers, is generally conceded to be the most effective and practical for both telephone and telegraph reception, to prevent momentary overloading, which distorts modulation components or telegraph characters. At high telegraph keying speeds this overloading ahead of the final detector sometimes reaches serious proportions. For facsimile keying speeds this trouble becomes almost intolerable.

Because of ethereal limitations of short-wave circuits, it is considered best to use only the upper portion of the keyed impulse for operating a tone keyer or a recording mechanism, because echoes and multipath components can be greatly discriminated against since they are generally lower in level than the desired character, especially when modern antennae are used. Hence the term, "riding the tops of the characters for best results."

In order to ride the tops of the characters properly, the speed of the operation of the automatic gain control must be rather fast, on the order of from .1 to .01 second. However, the strength of signal and the automatic fading range of the receiver, and the rate of keying speed limits the rate at which the AGC can be satisfactorily operated. The speed of the AGC action, above a certain point approaches noise elimination characteristics, if the signal strength is sufficient to strongly depress the gain of the receiver, immediately after the start of the impulse.

The operation of this method of automatic gain control is explained with reference to Fig. 1, which shows a workable schematic circuit. Numeral 1 is the final diode detector, and the diode output load is the potentiometer 2. Vacuum tube 3 blocks and unblocks a tone keyer, that is not shown, and tube 4 is a squelch or quiet tuning bias supply unit. Tube 4 may receive its plate supply from battery 5, or from the a-c supply at 6, in which case filter 7 is necessary. This bias supply appears across resistance 8. Both tubes 3 and 4 are biased slightly positive with respect to their cathodes, and have reduced plate and screen voltages to give a desirable plate-current grid-voltage operating characteristic.

With no carrier signal impulses incoming, tube 3 is conductive and blocks a tone keyer, as heretofore mentioned; tube 4 is also conductive and its current through resistance 8 furnishes a negative bias supply to the AGC through resistance 10 by way of condensers 11 or 12 depending on the time constant value desired. During signal impulses tubes 3 and 4 are cut off because of the negative signal voltage across resistance 2; the tone keyer is unblocked and relays a constant tone to the line or recording position, and the bias supply across resistance 8 is reduced to zero. The AGC voltage during impulses is the rectified signal voltage built up across resistance 2 and is fed in the AGC circuit through resistance 9 by way of condensers 11 and 12. The incoming signal impulse affects a changeover from fixed bias supply to a varying bias supply taken from the signal voltage. If the signal voltage across resistance 2 is exactly equal to that tapped from resistance 8, there will be no change in the gain of the receiver. If there is less voltage across resistance 2 during a signal impulse, the gain of the receiver will automatically rise, to produce an output voltage value across resistance 2 that will automatically check the upswing of gain. At another instant, if the signal impulse builds up a greater voltage across resistance 2 than that which is tapped from across resistance 8, then the gain of the receiver is reduced during each impulse.

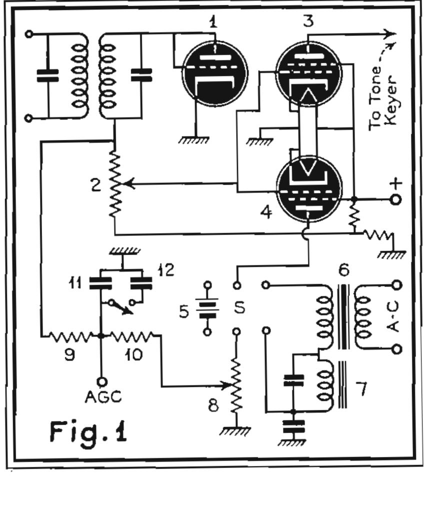


Fig. 1

AUTOMATIC SHIFT OF AGC TIME CONSTANT

(Telegraph reception only)

Manual adjustment of the time constant values cannot take care of sudden changes in receiving conditions, where the rate and depth of fading reaches serious proportions. Neither can a receiver operate normally over a wide gain range with extremely fast automatic gain control (unless the AGC is sliding or re-setting in nature), because the signal will split itself if the input value is high. The gain of the receiver is of course a factor in determining the signal input value necessary to cause splitting. A receiver with a sensitivity of one microvolt rating usually has sufficient gain to start splitting or denting the characters at ordinary telegraphic speeds, when the input is approximately 50 microvolts, and when a time constant on the order of .01 second or less is employed.

For diversity reception, deep and rapid fading usually does not warrant extremely fast operating AGC until the signal falls below approximately 35 microvolts input, during disturbed ethereal conditions, except for facsimile keying speeds as heretofore explained. Below this signal value (35 mv input) it is often advantageous to use a very fast time constant in the AGC circuit, especially where audio fading exists, or is being approached.

Fig. 2 shows schematically a circuit that is capable of shifting the time constant values of the AGC circuit automatically as a function of the signal strength, allowing for automatic coverage of a very wide range of ethereal conditions. *Numerals* indicate parts, the combined operation of which produces this desired result, wherein 1 is the final diode detector whose output load is the double arm potentiometer 2. 3 is a battery that supplies a positive potential to the grid of tube 4 that is used as a bias changer for the tone keyer, which furnishes also a squelch voltage for the AGC circuit between signal impulses. Battery 5 and potentiometer 6 complete the squelch-voltage supply. The squelch or quiet tuning action here is the same as described in Fig. 1. Tube 9, battery 10 and potentiometer 11 furnish a bias supply to keep tubes 12 and 13 blocked until the rectified signal voltage rises above a pre-determined level across potentiometer 2. By blocking tubes 12 and 13 when the signals are comparatively weak, condenser 14 is not connected to ground, consequently of no physical value in the AGC circuit. The AGC time constant under this condition is

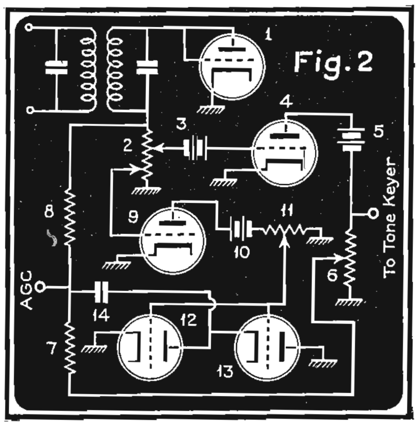


Fig. 2

made up of resistances 7 and 8 and the bypass condensers in the grid circuits of the tubes to which the AGC voltage is applied, therefore very fast AGC action is had. As the rectified signal voltage across resistance 2 rises sufficiently to cutoff tube 9, tubes 12 and 13 are conductive by virtue of the cathode emission and effectively zero grid potential, thereby connecting condenser 14 into active use in the AGC circuit. This charge leaks off rather quickly if the signal suddenly fades weak again, because of the parallel discharge path through resistances 7 and 8.

One of the outstanding characteristics of this circuit is the partial absorption of heavy crashes when condenser 14, which is comparatively large, is suddenly connected to the AGC circuit. This condenser suddenly absorbs the electrons within the AGC circuit, and partially absorbs the sudden high-voltage charge across resistance 2, through resistance 8. Therefore, instead of heavy crashes, when signals are weak, charging the AGC circuit resulting in harmful signal drop out action, the gain goes slightly upward, and holds over for an instant, and at the same time the full force of the crash does not reach the tone keyer.

Normally when the AGC circuits are attempted to be operated with such small time constant values, there is a tendency to produce howls, because of the path from the final diode or biased detector, whichever is in use, back to the input of the preceding tubes. The squelch tube here serves another useful purpose; that of balancing out the negative-resistance component between the output and input, through the AGC circuit, preventing audio oscillations within the receiver. This is quite evident when one considers the action of the first half-cycle of a voltage surge that would normally start an oscillation. The shock voltage is rectified, applied to the AGC, reduces its gain and reduces or cuts off the threshold voltage. At the half-cycle ends the threshold voltage comes on to check any further increase in gain, thus the surge is ended, being unable to sustain itself through the receiver channel. This non-oscillating feature might be useful in television reception to provide rapid automatic gain control for the video-channel amplifiers, since the rate of compensation probably could be rather high before causing distortion on the lowest modulation frequency. Its use in facsimile reception allows fast compensation that produced a marked improvement in general quality of received pictures.

AUTOMATIC THRESHOLD APPLICATION TO RADIO TELEPHONE RECEPTION

Overcoming fading entirely is indeed a big problem in radio-telephone reception. The overcoming of straight amplitude fading of the carrier is perhaps not entirely desirable, especially in short-wave reception, because levelling off the carrier output oftentimes over-amplifies the remaining sidebands that are already beating together due to loss of the carrier, and causes serious distortion.

Perhaps the best known method of reducing selective fading is by the carrier supply method, which feeds local carrier energy into the receiver automatically as the incoming carrier fades, preventing the side bands beating together to produce this undesirable distortion. This type of control is confined entirely to costly commercial telephone installations.

Spaced diversity reception overcomes selective fading very well, by virtue of the AGC selecting the stronger carrier, which automatically forces the weaker carriers from the combining circuit. This method of reception is also confined very largely to commercial telephone and program exchange services.

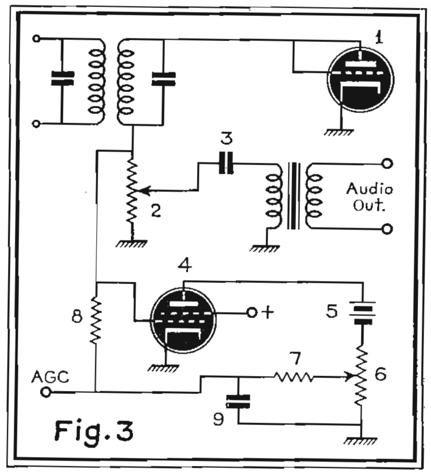


Fig. 3

For general broadcast reception on all waves, the AGC system shown in Fig. 3 serves rather well to produce quiet tuning between stations and to a degree overcome the effects of selective fading. In principle this circuit is not unlike that of Fig. 1.

Referring to Fig. 3, numeral 1 shows the final diode detector, feeding conventionally into an audio circuit, from the diode load resistance 2, through coupling condenser 3. Tube 4 is the squelch or threshold tube, excited by plate battery 5. The plate current from tube 4 passes through resistance 6 building up a negative bias supply. When no carrier is being received, tube 4 is conductive, and the voltage across 6 is supplied through resistance 7 by way of condenser 9, thence to the grids of the radio and intermediate amplifying tubes, that have their gain automatically controlled. When a carrier is received, the plate current of tube 4 is reduced, reduced proportionately to the strength of the carrier, or cut off completely. The AGC voltage is now supplied directly or in part from the rectified signal voltage across resistance 2. If the receiver is tuned to a very weak carrier signal, tube 4 may be only reduced, but this reduction lessens the QAGC supply voltage, increasing the gain of the receiver, which in turn repeats the operation until the receiver reaches full gain or checks the upward swing of gain automatically, by the increased rectified carrier voltage across resistance 2.

During periods of selective fading, the effects are somewhat reduced if the upswing of the receiver gain is delayed, or actually reduced slightly during such periods. During such fading the beating of the side bands oftentimes does not become perceptible, if the gain of the receiver is held within limits for short duration fades. For slow, long duration fades the gain build-up features heretofore explained gives excellent output level control. Adjustment

(*Continued on page* 14)

Frequency-Response Control Networks

By BERNARD EPHRAIM

BASS AND TREBLE control networks are used to increase or diminish the low and high audio-frequency components in tonal qualities by varying the amplifier response curve.

Actual accentuation of the bass and treble registers produces a major effect which is not classed as being psychological as that ordinarily obtained with common tone controls; the latter, mutates the higher and middle-range frequencies and gives pseudo volume to the lower register without increasing the low-frequency gain.

In control systems that are frequency responsive, the gain can be increased to as much as 20 or more decibels, plus or minus, at either end of the auditory spectrum. In this manner, accentuation of the reproduction is at the discretion of the operator or listener who can modify the flat response curve of an amplifier by simple controls for correcting room acoustics, enhancing the qualities of voice frequencies in transmission, or that of suiting ones' personal tone appreciation.

The proper circuit placement for a frequency responsive network is at a point where the audio-frequency gain is very low, preferably after the detector stage. A separate tube is employed to secure the necessary gain in response. The network impedances used in the system load the tube by reflection. The amount of gain required over the normal flat frequency response at each end of the register should be approximately 20 decibels. The normal gain over the complete register, excluding the network system, should be about one-tenth the maximum value; that is, one-tenth of 20 decibels. The interpolation of the extra tube raises the output from the detector a few decibels. This is advantageous in that sufficient gain is procured from the tube to permit current pick-ups being connected directly into the tube grid circuit.

For specific instruction, a 6N7 tube is used; the characteristics being, S, 1.3 ma/V; μ, 32.5; R_p, 12,500 ohms. The tube is worked to produce an output peak voltage of 1.96 for driving a 6J7 phase inverter into a push-pull 6N7 amplifier. The input from the detector to the control stage is .5 volts rms or .707 peak which gives the input stage a gain of 1.96/.707 or 2.78. *This is the normal gain for flat frequency response without the correction network.*

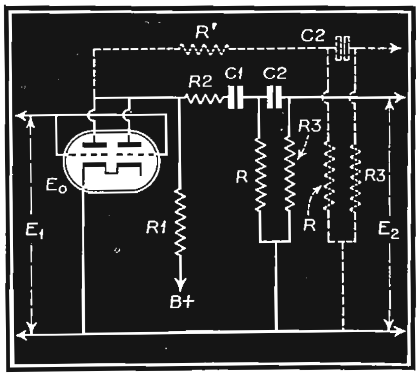


Fig. 4. Circuit for decreasing the bass response.

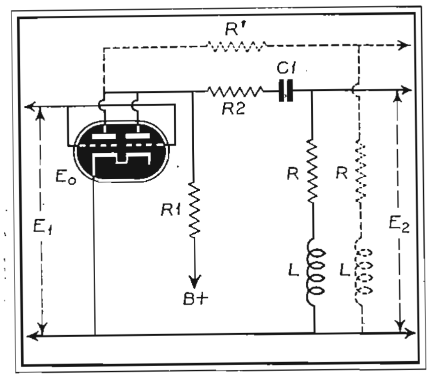
The amount of shading in the response depends upon the discretion of the designer. In general, a two step increase above and below normal flat frequency response is sufficient for all practical purposes. For example, increments can be ranged from 7 to 10 db, and from 15 to 20 db above and below normal as shown by the response levels in Fig. 1.

The amount of gain greatly depends upon the characteristics of the tube employed. It is essential that the tube function only on the straight-line part of its operating curve. The plate circuit must be loaded to twice the normal plate resistance of the tube for eliminating amplitude distortion when the upper register is accentuated.

PLATE LOAD

The minimum resistive load, referring to Fig. 2, is equal to $R_1 \times (R_2 + R)/$

Fig. 2. Circuit for bass emphasis. Dotted lines show electrical equivalent.



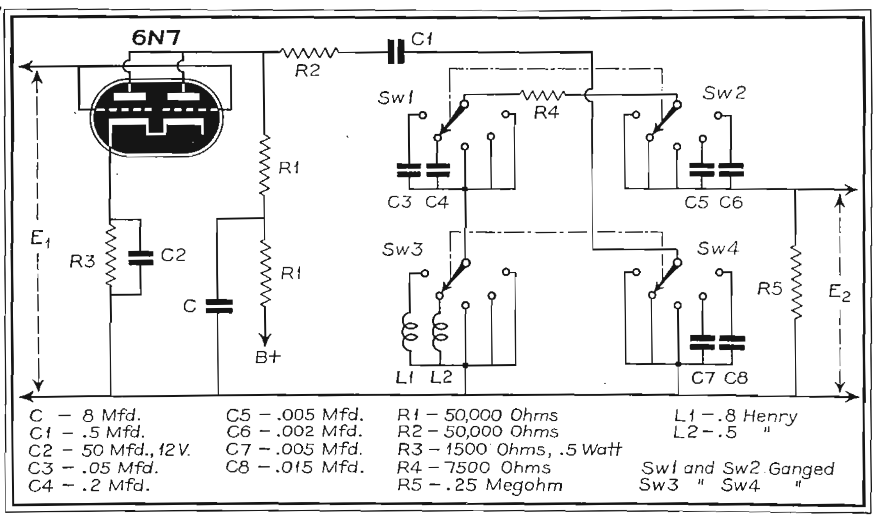
$(R_1 + R_2 + R)$. Ordinarily, R is of a smaller magnitude than R_2 and the actual load is represented by a parallel approximation of R_1 and R_2. Linear behavior will be obtained by combining values R_1 and R_2 so that each are equal. In this manner, calculation is facilitated and the selection of the electrical values are in keeping with available standard parts.

In computing the various resistive quantities in the network, it is only necessary to compute one general value, denoted as R', hence:

$$R' = \frac{R_p \times R_1}{R_p + R_1} + R_2 \quad \ldots\ldots\ldots\ldots (1)$$

Where R_p = 12,500 ohms; R_1 (also R_2) = 50,000 ohms. Note the parallel connection of the tube plates in Fig. 2.

Fig. 6. Complete circuit for treble and bass frequency control.



Substituting values for letters in (1):

$$R' = \frac{12,500 \times 50,000}{12,500 + 50,000} + 50,000$$
$$= 60,000 \text{ ohms}$$

The output voltage E_o developed across load resistor R_1 is found from the expression:

$$E_o = \frac{E_1 \times \mu \times R_1}{R_p + R_1} \quad \ldots\ldots\ldots\ldots.(2)$$

where E_1 is the input voltage to the grid circuit, in this case, 1.96 peak.

Evaluating:

$$E_o = \frac{1.96 \times 32.5 \times 50,000}{50,000 - 12,500} = 50.96$$

The maximum gain from the stage is found from the ratio E_2/E_1 or 50.96/1.96 = 26 for the μ of the stage.

For flat over-all response, less the effect of the definitive networks, resistors R′ and R are only effective, hence

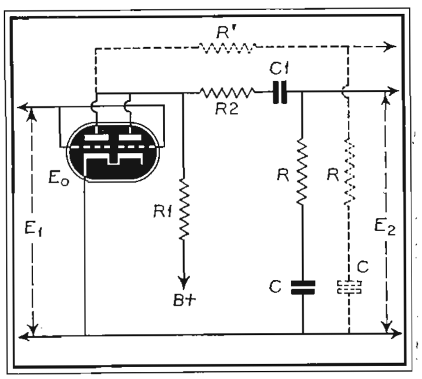


Fig. 3. Circuit for treble emphasis.

$$E_2 = \frac{E_o \times R}{R + R'} \quad \ldots\ldots\ldots\ldots\ldots.(3)$$

where

$$R = \frac{\mu \times R'}{26 - \mu}$$

μ being the normal gain, assuming 3 as a whole number for flat response instead of 2.78;

hence,

$$R = \frac{3 \times 60,000}{26 - 3} = 7820 \text{ ohms}$$

finally

$$E_2 = \frac{50.96 \times 7820}{7820 - 60,000} = 5.8$$

and the gain is found by equating the ratio

$$E_2/E_1 = 5.8/1.96 = 3 \quad \ldots\ldots\ldots.(4)$$

all of which demonstrates that the load resistances are of the optimum value and that the gain is within specified limits.

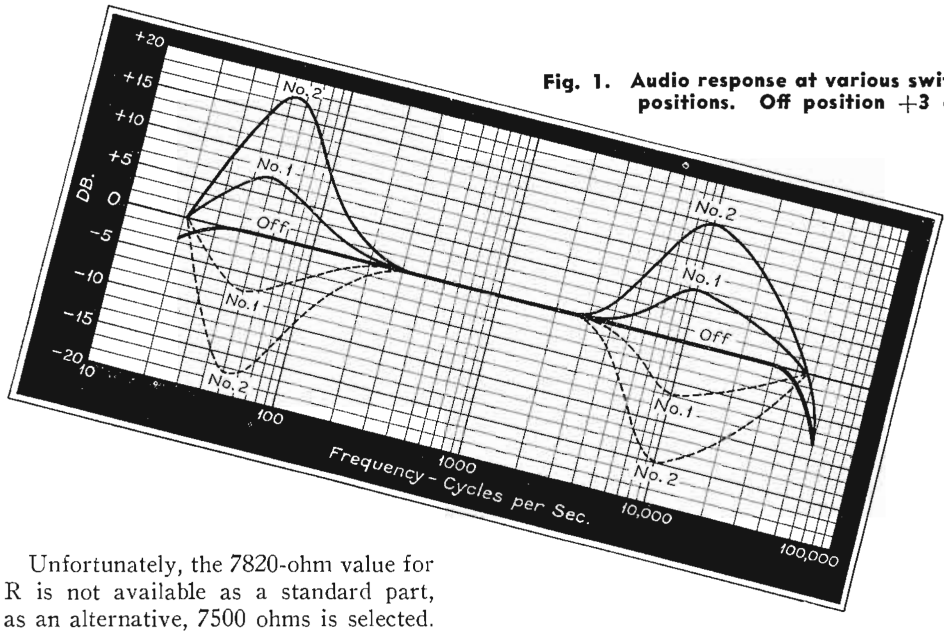


Fig. 1. Audio response at various switch positions. Off position +3 db.

Unfortunately, the 7820-ohm value for R is not available as a standard part, as an alternative, 7500 ohms is selected. This value decreases the voltage amplification obtained in expressions (3) and (4) so that now the amplification is

$$\mu = \frac{26 \times R}{R + R'} \quad \ldots\ldots\ldots\ldots\ldots\ldots.(5)$$

and solving

$$\mu = \frac{26 \times 7500}{7500 + 60,000} = 2.89$$

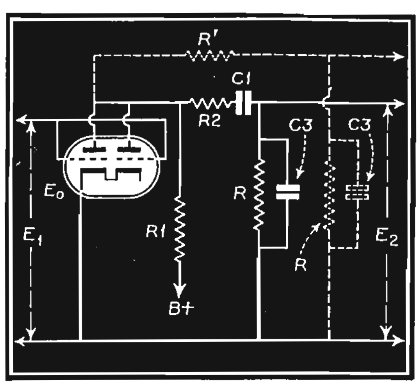
Note that the gain is still within the calculated value which was originally 2.78, and estimated to be 3 to simplify calculation.

A review of the preceding computations reveals that the minimum gain for flat response is 2.89; maximum 26, and that the relative voltage amplification is 26/2.89 or 9 which is approximately 18 decibels. In practice, an odd expression to use here as this is a practical discussion, it will be found that the maximum gain calculations are deceitful and that the maximum voltage gain obtainable is less than that figured by approximately 6; a conservative estimate for gain in the illustration is 20.

INCREASING BASS RESPONSE

The design configuration of the network used for low-frequency emphasis

Fig. 5. Circuit for decreasing treble response.



amplification ratio E_2/E_o at any selected low frequency is determined from the expression

$$E_2/E_o = \frac{\sqrt{1 - f^2 \times C^2 \times R^2}}{\sqrt{1 - f^2 \times C^2 \times (R - R')^2}} \quad \ldots\ldots\ldots\ldots.(6)$$

and in the design specifications, the minimum gain is equated from

$$\mu = E_2/E_o \times 26 \quad \ldots\ldots.(7)$$
$$= 5.8/50.96 \times 26 = 2.86$$

The value of condenser C for maximum bass emphasis can be quickly determined by arbitrarily selecting the lowest frequency which definition is to be given; taking 50 cycles as the lowest bass response and the gain at 20, as previously explained, the capacity is found where

$$C = \frac{\sqrt{1 - \frac{\mu^2}{670}}}{f \times \sqrt{\frac{((R + R')^2 \times \mu^2) - R^2}{670}}} \quad (8)$$

Substituting quantities for letters:

$$\frac{\sqrt{1 - \frac{20^2}{670}}}{50 \times \sqrt{((7500 + 60,000)^2 \times 20^2) - 7500^2}} = .0392 \text{ mfd}$$

$$= .0392 \text{ mfd.}$$

Since .0392 mfd is not available as a standard component, an alternative value that will suffice is .05 mfd.

The amplification ratio equated from expression (6) will give 18.2, and in gain is equal to 18.2/2.89 or 6.3, in decibels this is approximately 16 db.

For minimum bass definition, a gain of only 8 db is required or a voltage

(*Continued on page 32*)

IRE CONVENTION PROGRAM

THE Pacific Coast Convention of Institute of Radio Engineers is being held from June 27-30, 1939, at the Hotel Mark Hopkins, San Francisco, California. An interesting program of technical papers has been prepared. In addition there will be a number of inspection trips to various points of interest. The program follows:

TUESDAY, JUNE 27

9:00 A. M.

Registration.

10:00 A. M.—12:00 NOON

Opening address by R. A. Heising, President of the Institute.

(1) "Communications Engineering in Geophysical Exploration," by Herbert Hoover, Jr., United Geophysical Corporation.

(2) "Federal Communications Commission Engineering Regulations and Standards of Good Engineering Practice for Broadcast Stations," by S. L. Bailey, Jansky and Bailey, Washington, D. C.

(3) "Columbia's West Coast Operations," by L. H. Bowman, Columbia Broadcasting System, Hollywood, Calif.

2:00 P. M.—4:30 P. M.

(4) "Recent Developments in Aerial Navigation," by H. H. Willis, Sperry Gyroscope Company, Brooklyn, N. Y.

(5) "Aircraft Instrument Landing Research at the Massachusetts Institute of Technology," by E. L. Bowles, Massachusetts Institute of Technology, Cambridge, Mass.

(6) "Study of the Effects of Mountains in Radiogoniometry and on the Combined Use of Radio Beacons and Radio Compasses for Aerial Navigation," by Andre Busignies, Le Material Telephonique, Paris, France.

(7) "Acoustic Models of Radio Antennas," by E. C. Jordan and W. L. Everitt, Ohio State University. Columbus, Ohio.

(8) "Recent Advances in Receiving Equipment for Transoceanic Telephony," by F. A. Polkinghorn, Bell Telephone Laboratories, New York, N. Y.

WEDNESDAY, JUNE 28

9:00 A. M.—11:00 A. M.

(9) "Electron Optics," by V. K. Zworykin, RCA Manufacturing Company, Harrison, N. J.

(10) "Current Division in Plane-Electrode Triodes," by Karl Spangenberg, Stanford University, Stanford University, Calif.

(11) "Functions of Electron Bombardment in Television," by I. G. Maloff, RCA Manufacturing Company, Camden, N. J.

(12) "Surface - Controlled Mercury - Pool Rectifier," by T. M. Libby, Pacific Telephone and Telegraph Company, Seattle, Wash.

11:30 P. M.—2:00 P. M.

Trip to Pan American Airways Terminal and inspection of Boeing Clipper.

2:00 P. M. — 4:30 P. M.

(13) "Direct-Current and Audio-Frequency Amplifier," by L. J. Black and H. J. Scott, University of California, Berkeley, Calif.

(14) "Golden Gate International Exposition Radio and Sound Distributing Systems," by C. A. Lahar and L. Hewitt, RCA Manufacturing Company, Camden, N. J.

(15) "Radio-Frequency Spark-Over in Air," by P. A. Ekstrand, Heintz and Kaufman, South San Francisco, Calif.

(16) "Solar Cycle and the F_2 Region of the Ionosphere," by W. M. Goodall, Bell Telephone Laboratories, Deal, N. J.

(17) "Atmospherics and Radio Tansmission Phenomena in Puerto Rico," by G. W. Kenrick and P. T. Sammon, University of Puerto Rico, Rio Piedras, P. R.

(18) "Transmission on 41 Megacycles," by S. S. MacKeown, B. M. Oliver, and A. C. Tregidga, California Institute of Technology, Pasadena, Calif.

THURSDAY, JUNE 29

9:00 A. M.—11:30 P. M.

(Joint Session with American Institute of Electrical Engineers)

(19) "The Klystron as a Generator of Very Short Waves," by W. W. Hansen, R. H. Varian, S. F. Varian, D. L. Webster, and J. R. Woodyard, Stanford University, Stanford University, Calif.

(20) "Instruments and Methods of Measuring Radio Noise," by C. V. Aggers, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Penna.; D. E. Foster, RCA License Laboratory, New York, N. Y.; and C. S. Young, Pennsylvania Power and Light Company, Allentown, Penna.

(21) "Methods of Controlling Radio Interference," by C. V. Aggers, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Penna.

(22) "Technical Framework of our Television," by E. W. Engstrom, RCA Manufacturing Company, Camden, N. J.

12:30 P. M.

Trip to Stanford University jointly with American Institute of Electrical Engineers.

2:00 P. M.—4:30 P. M.

(23) "Electronic-Wave Theory of Velocity-Modulation Tubes," by Simon Ramo, General Electric Company, Schenectady, N. Y.

(24) "Recent Ultra-High-Frequency Developments," by B. J. Thompson, RCA Manufacturing Company, Harrison, N. J.

(25) "Simple Television Antennas," by P. S. Carter, RCA Communications, Rocky Point, L. I., N. Y.

(26) "Continuous-Wave Interference with Television Reception," by C. N. Smyth, Kolster-Brandes, Ltd., Sidcup, Kent, England.

FRIDAY, JUNE 30

10:00 A. M.—12:00 NOON

Trip to the tube manufacturing plants of Eitel and McCullough, San Bruno, and Heintz and Kaufman, South San Francisco.

2:00 P. M.—5:00 P. M.

Trip to University of California.

AUTOMATIC THRESHOLD CONTROL

(*Continued from page* 11)

of the threshold voltage will give these advantages automatically.

Therefore, in receiving broadcast programs, amateur calls, police calls, etc., on a single unit popular priced receiver, all that appears worthwhile at present is an attempt at overcoming fading in its various forms. The greatest improvement possible is perhaps in the hands of the short-wave transmitting stations—that of keeping the modulation percentages down appreciably, to reduce to a minimum the effects of selective fading. This policy might be desirable for night-time operation on the broadcast band, because a rural listener seldom can get consistently good reception beyond a distance of from fifty to seventy-five miles from the stations, and it is very interesting to note that this distortion did not exist to any appreciable degree until high-level modulation was used. Certain types of "monkey chatter" it seems might be another offspring of high-level modulation interwoven with selective fading.

Solving all of the problems of fading, outside of the BX Station's primary area by various methods of AGC that are practical in home receivers, does not appear to be in the immediate offing.

MOLDED CABINETS

for radio receivers

A modern Zenith receiver.

By F. D. SWANSON
Chief Tool Designer
CHICAGO MOLDED PROD. CORP.

WHEN you drive your automobile, write with your fountain pen, switch on your electric lights, or perform any one of a dozen other every day operations, you see and feel synthetic plastics. When you buy any of these various articles of every day use your choice is based almost entirely upon its utilitarian value, but when you purchase a radio you are influenced in your selection just as much by its appearance as by its excellence of reception.

Because good design is such an important factor, it is usually well for the manufacturer of radios to call on an aesthetic designer for the external features of the case. Most of these designers have had some experience with plastics and know how to design in relation to what the product engineer requires of the outside of the case, such as speaker openings, vents for air circulation and openings for controls.

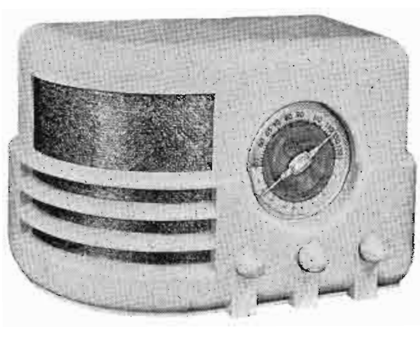
The mechanical designer is apt to violate the laws of art in his efforts to design for economical production. Internally, the case is still a mechanical design to receive, support and anchor the chassis and its component parts. Here, the engineer in charge of mechanical design is concerned chiefly in providing the necessary holes, depressions, projections—all molded in uniform locations and of sizes to a tolerance to take care of the mechanical features that would otherwise require additional manufacturing equipment, parts and assembly.

The more generally used types of plastics in connection with radio cabinets are the phenolic and urea materials. Radio cases of brown, walnut, black, are usually molded of the phenolic materials. But when the gay pastel colors are specified, the urea type of material is best suited.

The urea molded materials are somewhat higher in cost than the phenolic and are justified chiefly because of their color. Molding equipment, if properly designed, will produce cases from either type of material.

At the present time radio case dimensions have generally not exceeded 8″ × 12″ face × 8″ depth from front to back. With these normal dimensions in mind, wall thicknesses should be approximately ⅛″ throughout the case. In the smaller size or "midget" cases 3/32″ wall thickness is proper. Heavy sections or thick irregularities projecting from the walls of the case should be cored out to the general thickness to avoid finish blemishes adjacent to them on external surfaces. This is especially true in the case of the urea materials. Avoid all sharp corners wherever possible on the outside as well as the inside of the case; substitute a 1/32″ fillet or radius. This strengthens the case, aids the flow of material and prevents chipping. Since the case is hard and brittle, some molding strains are set up in the molding procedure. These strains will tend to alleviate themselves at the weakest part of the case, which is usually at the sharp inside corners—round corners absorb these strains.

Both the designer and the mechanical engineer must visualize the manner in which the case is going to be molded, so that undercuts or projections will not be incorporated into the design of the case, for an undercut will not permit withdrawal of the punch or male member of the mold, nor will it permit withdrawal of the finished product from the female member of the mold. Undercuts and side projection require split sections in the mold in order to withdraw the molded part—this slows up production and of course adds to the cost.



A midget Majestic model.

Another point to receive careful attention is design that will meet with approval by the Underwriters' Laboratories. Electrical and fire hazards can be eliminated by proper engineering design.

Since the molding of radio cases is usually a matter of mass production, if there is any element of doubt regarding consumer acceptance or design, it is sometimes advisable to have only a single impression mold made to prove out actual practice before going into mass production equipment.

The molding operation requires pressure of approximately 2½ tons to the square inch of projected molding area and so the number of mold impressions or cavities is limited in relation to the maximum pressure capacity of the molding press to be used.

The largest molding press generally in use among the commercial molders is about 500 tons maximum pressure capacity. It is obvious, therefore, that if the case is a very large one its production and molding equipment is going to be limited to a one impression mold per molding press, though when it comes to the smaller or "midget" cabinets it is possible to use a six impression mold in the larger molding presses, with the result that every time the press closes and opens six radio cases will be ejected from the mold. Normally, the larger presses operate only about 22 times per hour, giving a production of 132 pieces per hour from a six cavity mold. Since time is money, the financial advantage ultimately resulting from the number of impressions in the mold should be considered. So must also the cost of manufacturing equipment needed to mold cases be considered. The most important part of the equipment is the hardened and polished steel mold, the cost of which, per impression, varies considerably in relation to the complexities of the deisgn of the case.

Doubling the number of cavities in a mold does not double the mold cost nor halve the piece part cost, though it nearly doubles the daily production. Therefore, to reach a decision regarding the size mold (number of cavities) best suited to economically take care of a manufacturer's requirements, all three factors—mold cost, piece cost and daily production requirements—must be taken into consideration. No general figures covering these three factors can be given, as each case presents a very

(*Continued on page* 30)

TELEVISION ENGINEERING

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SOUND MOTION PICTURE FILMS IN TELEVISION

Part II

By JOHN A. MAURER
THE BERNDT-MAURER CORP.

IN the first article of this series the writer set forth his reasons for believing that the sound-motion-picture film is destined to play a much more important role in the development of television than its analogue, the transcription disc, has played in sound broadcasting. His thesis may be summed up in this: that the film is more than an economical medium for the distribution of visual program material; it is a medium for the creation of superior television programs with the expenditure of less time and effort than is required for high-quality "live" broadcasts.

If this view of the significance of the film technique in television is even partly correct, it is highly desirable to establish a satisfactory set of standards for television films and film-scanning equipment before the setting up of any large number of transmitting stations. The present article and the one to follow it are concerned with the determination of the proper film dimensions for such a set of standards in order to insure satisfactory picture and sound quality.

The writer feels that the first requirement of any set of television film standards should be demonstrable ability to satisfy not only present requirements as to quality but all requirements that may arise in the future under the RMA standards for television transmission. This means that the film, when properly handled, must be capable of providing all the picture detail that can theoretically be transmitted over a 441-line system, and the associated sound record must be capable of development to a degree of fidelity which will not suffer by direct comparison with direct transmission over a 15,000-cycle channel.

Note that in neither of these respects is the specified degree of quality needed at the present time. Therefore it is not necessary to limit ourselves to presently available techniques in determining the dimensions needed to secure the future availability of high enough quality. What is needed is to estimate carefully and conservatively the probable effect of improvements in technique and materials that are now in the laboratory stages, and on this basis concede as much as seems safe to the present need for economy in film consumption.

If quality of performance were the only consideration we might at once decide in favor of the theatrical standard, or 35-mm film, for television. Even when carelessly handled it provides more than enough detail in the picture, while the sound track, if recorded by the best available present methods, could give a flat frequency response to at least 15,000 cycles, with greater volume range than can be heard enjoyably under living-room conditions.

16-mm films, however, have so much to offer in reduced first cost, lower weight in shipping, reduced space requirements for storage, freedom from fire hazard, and general convenience and mobility of camera and projector equipment, that it is well worth while to study the requirements of picture and sound with the greatest care in order to determine how successfully the 16-mm film can meet them.

It may be asked at this point why some standard of film width other than 16-mm or 35-mm may not be the logical one for television. It is true that the requirements of television program production will probably require the development of special types of motion-picture cameras, while broadcasting definitely requires special projection, or scanning, equipment. But the production of films according to a given dimensional standard requires the existence of equipment of other types (film slitting and perforating machines, printing machines, and developing machines) in the hands of film manufacturing companies and film processing laboratories. Economy obviously dictates that television use one or the other of the two established sound film widths (35 mm or 16 mm) for which this equipment exists and is in regular operation.

The amounts of picture detail made available by television systems, films, and lenses, are variously stated in terms of scanning lines, resolving powers, and diameters of circles of confusion. It becomes necessary to define and interpret these terms if we are to make an accurate comparison of the performance standards that are implicit in the RMA television standards with the possibilities of 16-mm motion-picture films.

The finest line pattern that can appear on a television screen in the vertical direction is the one obtained when every other scanning line is bright, with the intervening ones dark. In the horizontal direction the spot size and the limits of the frequency range transmitted by the system usually make the finest obtainable line pattern the same as the one in the vertical direction. Putting this in numerical terms, the finest line pattern that can be transmitted on a television system according to the RMA standards will have 220 separate lines in the height of the picture.

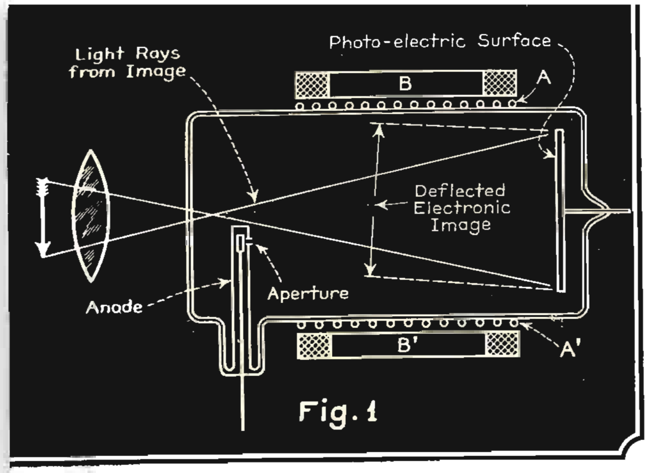
The ability of a photographic film to show fine detail is commonly stated in terms of the number of lines per millimeter it can resolve, or show separately. But in making tests of resolving power, the test object that is photographed con-

(Continued on page 27)

The Fundamentals Of

Part III

Television Cameras



Schematic diagram of Image Dissector tube.

IT makes little difference where one might go throughout the world examining electronically-operated television pick-up devices, he will probably find only variations from the two fundamental patents issued originally in this country, one to V. K. Zworykin about 1928[1] and the other to P. T. Farnsworth about 1931[2]. These two devices pretty much dominate the international television picture at the present time. In foreign countries, the television cameras may appear under unfamiliar names, but a closer scrutiny will probably reveal the basic principles of operation of one of the two cameras to be described in this installment. For instance, in England the Emitron camera of the Marconi-E. M. I. Company resembles Zworykin's *Iconoscope,* and the Baird Electron Camera is similar to Farnsworth's *Image Dissector.* Because of this fact, a study of the two pick-up systems used so extensively in the United States today will give us an up-to-date working knowledge of the television pick-up systems of the world.

In Part I of this series, the necessity for scanning and for the translation of the average light level of each incremental area of the picture into an electric current of corresponding intensity was pointed out. In both types of television camera tubes widely used today both of these processes, i.e., the scanning and the optico-electro translation, occur within the same device. In addition to this, several models also include means for amplification of the feeble signals so that they have a fighting chance against the ever present circuit noises.

The operation of the Image Dissector is made clear by Fig. 1. The image to be scanned is focused by a conventional system of lenses onto the cathode surface which has been treated uniformly for photo-electric emission. It is evident that the bright areas will cause many electrons to be emitted and that the darker regions of the image will cause fewer electrons to be emitted from this photo-cathode surface. The anode in the opposite end of the tube is held at a positive potential with respect to the cathode so that all of the photo-electrons emitted will be accelerated toward the anode. Leaving the photo-electric cathode, then, is a beam of electrons about the size of the image whose electronic density along its cross-section will vary in a manner similar to the light variations over the image as it falls upon the cathode. In other words, if one could take an imaginary slice from this electron bundle leaving the photo-cathode he would find that in the regions of the slice corresponding to the light parts of the image, there would be found many electrons and the areas corresponding to the dark parts would be represented by only a relatively few electrons. This arises from the fact that the photo-electric emission from the cathode is a function of the intensity of the impinging light.

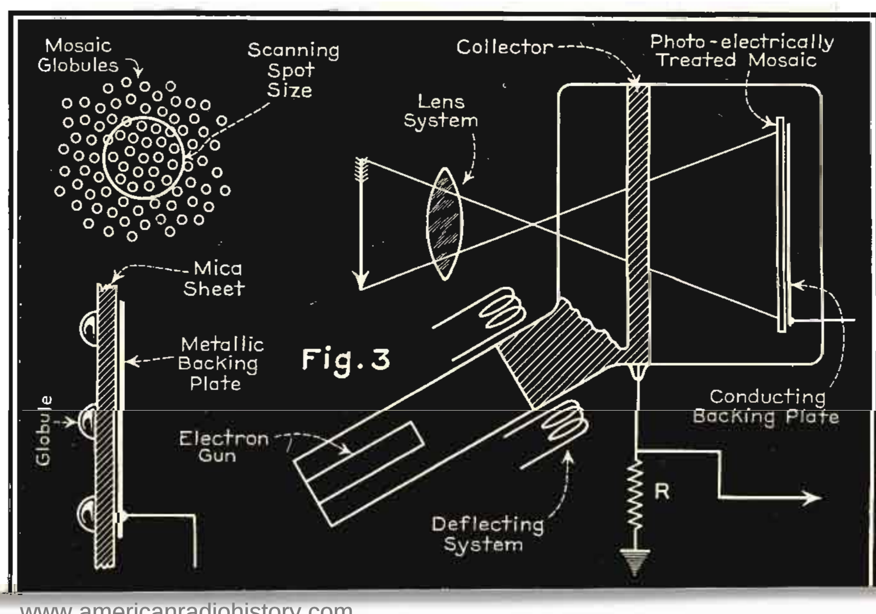
Fig. 2. The basic Iconoscope. Photo courtesy RCA Review.

Because all of the electrons in a beam possess a negative charge, there will be mutual repulsion between the various electrons comprising the electronic image. This effect is further augmented by slight initial variations in velocity and direction as the photo-electrons are emitted. These effects tend to make the electronic image bundle to spread apart, but this spreading is minimized by focusing coils A, A′ mounted coaxially with the tube. Its magnetic field is parallel to the direction of travel of the electrons, and any electron attempting to travel diagonally to this magnetic field has a force acting upon it tending to bring it back into line.

Above and below the evacuated cylinder is a pair of coils (B and B′) connected in series, so situated that their axis is perpendicular to that of the tube. The magnetic field resulting from current flowing in these two coils will result in the electronic image bundle being deflected upward or downward in the plane of the paper in Fig. 1. Another pair of coils is arranged one on the side of the observer and one on the back side so that its magnetic field is perpendicular both to the axis of the tube and the plane of the paper. By means of a current in these coils, the electronic image bundle may be deflected toward or away from the observer. These two sets of coils make it possible to deflect the electronic image at will within the tube by the simple expedient of sending currents of suitable wave form through the coils outside the tube.

As the electronic image approaches the anode pillar, a few electrons will go through the aperture, hit the tiny target inside the anode pillar structure, and this constitutes the signal current. The two sets of deflecting coils are so energized that each picture element is scanned in the proper sequence. For RMA standards of operation, the image beam would be deflected horizontally 441 times per second and vertically 60 times per second to give a 441-line, 30 frames per second, interlaced image. The entire electronic image is moved across the aperture to accomplish the scanning process in this Farnsworth Image Dis-

Schematic drawing of an RCA Iconoscope.



Television Engineering

By F. ALTON EVEREST
Dept. of Elect. Eng.
OREGON STATE COLLEGE

Illustrating the improved Farnsworth tube.

sector, while the scanning point is movable over a stationary image in most other systems.

The photo-electric current representing the light intensity of one *elemental area* is very feeble. A brief calculation will confirm this. Let us assume the use of RMA standards of 441 lines per image, 30 complete frames per second, and aspect ratio of $\frac{4}{3}$. The number of elements per frame is $(441)^2 \frac{4}{3} = 259{,}000$. As there are 30 of these frames per second, the time that one single elemental area will be in front of the aperture will be $\frac{1}{(259{,}000)\,(30)}$ seconds = 0.129 microsecond. Now, let us assume the use of an F-4.5 lens in front of our Image Dissector throwing a brightly illuminated outdoor scene upon the photo-electric cathode. Under these conditions, the total light falling upon the cathode will be in the order of 0.1 lumen. Let us also assume that the photo-electric surface has a sensitivity of 75 microamperes per lumen, an extremely sensitive surface which has been obtained by much research work. The photoelectric current representing a single elemental picture area is $\frac{(75 \times 10^{-6})\,(0.1)}{259{,}000} = 28.9 \times 10^{-12}$ amperes or 28.9 micromicroamperes per element. This current flowing for the 0.129 microsecond is equivalent to 3.74×10^{-18} coulombs which is equal to 23.5 electrons. In an extremely generous mood, we will call it an even 24 electrons, which, one must still admit, is not much of an electric current. This signal current would undoubtedly be lost in the noise associated with ordinary thermionic amplifiers and because of this inherently feeble signal from the Farnsworth Dissector, electron multipliers are used. In one of the later models, this multiplier is built into the anode pedestal.

An early type of RCA Iconoscope (Greek: "image observer") television camera tube is shown in the photograph of Fig. 2. A schematic drawing of a commercial model (Type 1849 and 1850) recently put upon the market is shown in Fig. 3. The type 1849 Iconoscope is designed for motion-picture pick-up, while the type 1850 is much more sensitive and is intended for direct pick-up at low levels of scene illumination.

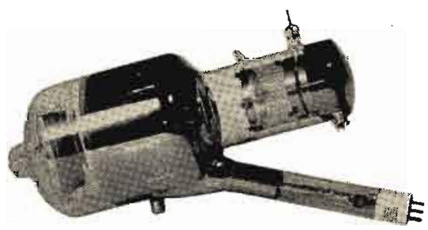
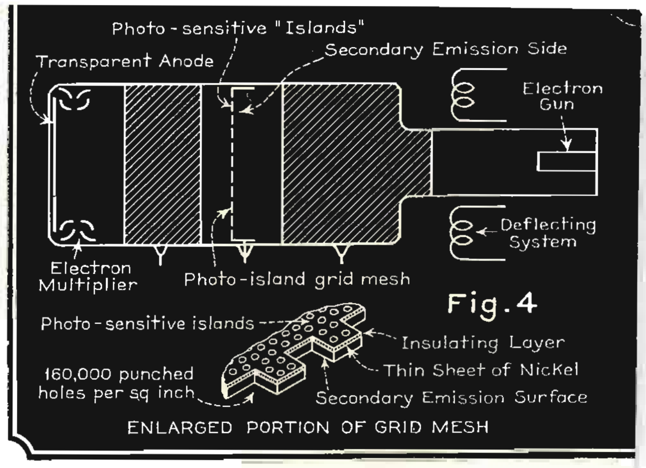


Fig. 5. The new Image Iconoscope. RCA Photo.

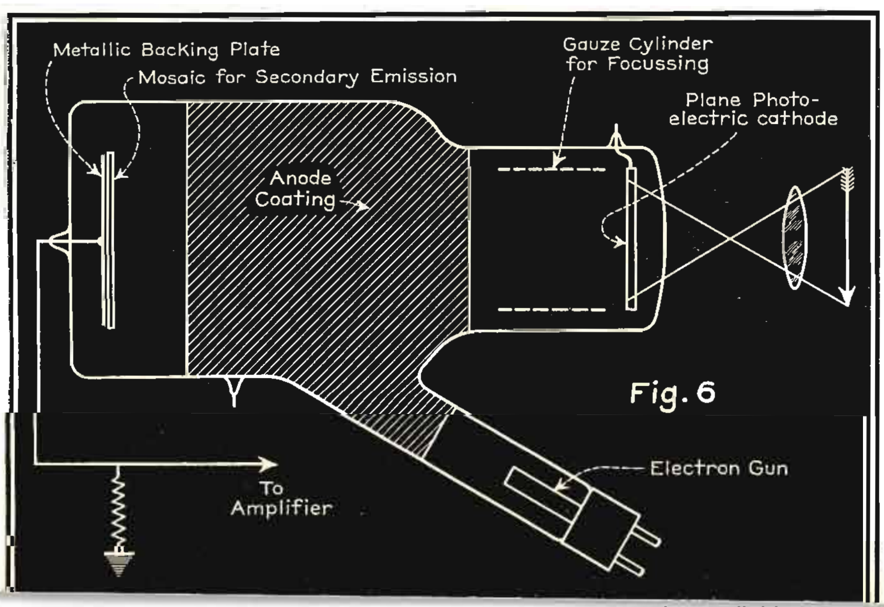
The heart of the Iconoscope is the mosaic electrode which has been especially treated for high photo-electric emission. The mosaic may be formed by the deposit of a multitude of tiny silver globules upon an insulating sheet such as a thin sheet of mica. These globules are then photo-sensitized by caesium and each globule, which is insulated from all its neighbors, becomes a minute photo-electric cell. These globules are so small that there may be dozens of them in one elemental area of the mosaic or the area of the scanning spot. In general, about 30% to 40% of the area of the mosaic is covered by the globules.



An electron gun and associated beam deflecting system are mounted in the neck of the Iconoscope. This gun is very similar to that found in the usual cathode-ray tube and consists of a thermionic cathode for emission of the electrons, means for accelerating the electrons, and means for focusing them into a very fine beam. By means of an electromagnetic or electrostatic system (the Iconoscope uses the former), the beam may be deflected to any spot on the mosaic electrode. To meet the RMA Standards, this beam would be swept horizontally across the mosaic 441 times per frame, and the beam would also be deflected slowly in a vertical direction so that each line would fall adjacent to the preceding one, the 441 lines scanning all parts of the mosaic surface every $\frac{1}{30}$th second.

Let us examine the mechanism by which the signal currents are generated. The image is focused upon the mosaic by means of a suitable external lens system. The light falling upon the mosaic causes photo-electrons to be emitted from each element of the mosaic. The sensitized silver globules lying in a part of the image which is light will give off more electrons than the dark portions. The electrons given off from each mosaic element photo-electrically are attracted to the silver coating on the inner side of the tube which constitutes the anode and which is held at a positive potential with respect to the mosaic. It is obvious that the leaving of the electrons from the mosaic element will leave a deficiency of charges upon it and, by virtue of the capacitance existing to the metallic backing plate on the opposite side of the mica sheet, this will actually result in a charging of this tiny condenser. The magnitude of the charge will depend upon the intensity of the light falling upon it for a given

Schematic drawing of tube shown in Fig. 5.



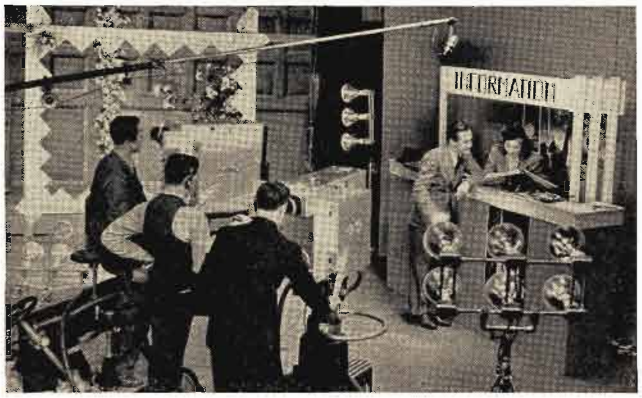


Fig. 9. A typical scene during the broadcasting of a television play. Note the trucks upon which the cameras are mounted and the powerful lights required. RCA photo.

length of time. Because each of these mosaic elements is highly insulated from every other element, it is seen that a scene focused upon the mosaic will immediately give rise to a potential distribution over the face of the mosaic which varies electrically as the light and shade of the scene itself varies optically.

The function of the electron beam is to discharge these tiny charged condensers in a certain order. The sweeping of the electron beam across a charged element will mean the equalization of the charge, or the discharge, of that condenser element. The charging current which flowed to perform this equalization is proportional to the amount of charge on the element, which is in turn porportional to the intensity of the light falling upon that element. The current which flows through resistor R of Fig. 3 produces a voltage which varies as the light variations along that particular scanning line, and this constitutes a feeble signal voltage which can be amplified and utilized.

As mentioned before, the area covered by the scanning beam contains many of these mosaic elements. Because of this, the signal output of one elemental area of the mosaic will be proportional to the average charge attained by all the globules in that elemental area.

The sensitivity of the Iconoscope is much greater than the fundamental Dissector. This results from the storage effect that takes place by the more or less continuous process of charging the minute condensers. While the signal from a single elementary area of the Farnsworth Dissector tube might be in the order of 24 electrons, the signal from a single elemental area of the storage type would be much greater because its charging process has been progressing while all the rest of the approximately 259,000 elemental areas were being scanned in turn. In other words, the Dissector tube has only the time required to scan a single element for the photo-electric emission of its signal current (about 0.13 microsecond (while the Iconoscope merely releases during this same time a charge that has been accumulated for about $\frac{1}{30}$ second. The theoretical gain of the Iconoscope over the Dissector would be about 259,000, but an advantage of only a few thousand has actually been realized.

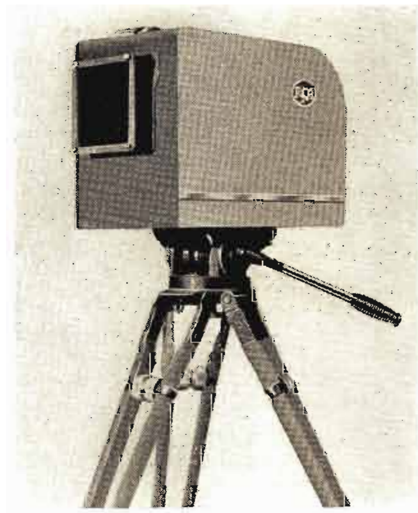


Fig. 8. A new RCA Iconoscope television camera.

The Improved Farnsworth Pick-up Tube

An interesting thing about the improved types of Farnsworth and RCA tubes is that the new Farnsworth tube utilizes a photo-mosaic and the new RCA tube uses electronic images.

The improved Farnsworth tube is shown schematically in Fig. 4. The image is focused upon the special "photo-island" grid after passing through the transparent anode on the end of the tube. This grid has about 160,000 holes punched per square inch in a thin nickel plate. One side is coated with a dielectric material which has deposited upon it many photo-sensitive "islands" which are so-called because they are insulated from each other as are the globules on the Iconoscope mosaic. The image focused upon this "island" surface causes an electrical potential image to be set up over its face. The beam of electrons from the gun hit the special surface on the nickel plate liberating copious quantities of secondary electrons. This cloud of secondary electrons acts as a rapidly moving virtual cathode source, and these electrons are drawn through the tiny holes of the mesh to a degree depending upon the amount of positive charge built up on the "photo-islands" on the other side. In other words, the "photo-islands" act as the control grid of a triode in that they control the number of electrons which shall go to the electron multiplier to represent that particular area. The intensity of illumination determines the amount of positive charge on the islands, and this positive charge determines the number of electrons allowed to go to the electron multiplier which constitutes the signal current.

The main advantage of this tube is that its sensitivity is increased to about ten times that of the conventional Iconoscope. Another advantage is that a peculiar shading signal common to the Iconoscope and evidently a result of roving areas of spurious charges over the mosaic does not appear. A difficulty at the present time is constructing the photo-island mosaic so that its charge leaks off in about $\frac{1}{30}$ second so that no residual charge remains when the next frame starts.

Improved RCA Iconoscope

The Image Iconoscope recently described has resulted in greatly superi-

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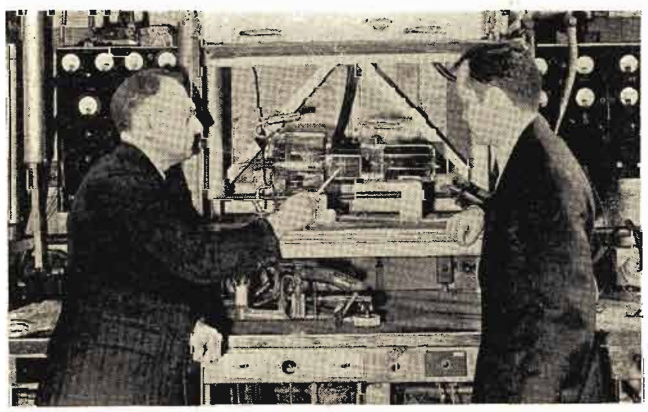
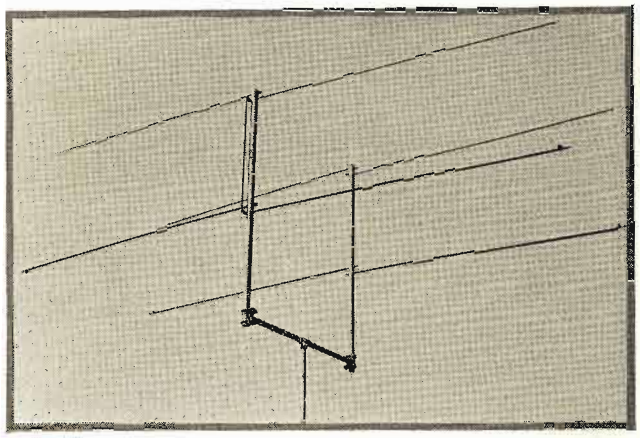


Fig. 7. Two well-known television research engineers, Dr. V. K. Zworykin (left) and E. W. Engstrom, examine a special electronic tube. Dr. Zworykin is the inventor of the Iconoscope. RCA photo.

Television Economics

Part V



A television receiving antenna. RCA Victor photo.

By

Dr. ALFRED N. GOLDSMITH

Consulting Industrial Engineer

THERE is one matter of outstanding importance which greatly affects the broadcasters and the lookers alike. This matter is the proposed discontinuance of present broadcast station procedure whereby the public is admitted to the studio or permitted to view the actual broadcast. It cannot be too strongly urged that the public and non-official observers shall be completely excluded from television studios from the very beginning of that art. In the first place, it is bad showmanship to show the public the various detailed and sometimes disillusioning methods which are used in production. In the second place, studio personnel should be interested in the home audience and their specific and immediate production job only—and not be compelled consciously or unconsciously to pose for a gaping gallery. In the third place, the cost of providing adequate viewing facilities for the public would be unwarranted, particularly in view of the difficulty of placing the observers in such positions that they are not blinded by the lights.

It has been repeatedly stated by the writer that the entertainment industry should be conducted by "sellers of illusion and vendors of glamor." Through receiver costs and its own time and attention, the public seeking radio entertainment pays for entrance into a new world created by the entertainer; and it is both unwise and unfair to destroy the remoteness, entertainment appeal, and psychological novelty of this created world by showing the public "how the wheels grind and turn." Broadcasting now has the rare and unique opportunity to rid itself of an unnecessary and purely extraneous element which should be foreign to the entertainment industry, namely the intrusion of the public into the routine of entertainment production. If the advertising sponsor and his agency will forget their personal preferences and remember that the major function of broadcasting is to entertain the audience of prospective purchasers *in the home,* they will wisely refrain from requesting the trivial gratification of inviting any portion of the public or their own affiliates to witness the production in the studio. It will be a disappointment to the public, a liability to the advertiser, an annoyance to the broadcaster, and a financial drain on all concerned if this note of caution is not immediately heeded.

G. RECEIVING ANTENNA SYSTEMS

G-1 Antennas

Continuing the study of television methods and their economics, the receiving end of the system will next be considered. Ultra-high-frequency transmission is accomplished by direct (line-of-sight) rays, by reflected rays (mirrored by the ground, lateral obstacles, or more remote structures), by diffracted or refracted rays, and by sky waves (which occasionally are reflected back from overhead to the ground over extended paths even on these very high frequencies). Desired reception is usually by means of the direct ray; and only the direct or reflected rays are of practical importance in successful reception.

Differences of opinion exist as to the more desirable plane of polarization of the wave for reception with lowest noise level. European opinion tends toward vertical polarization, but American practice has favored horizontal polarization. Accordingly, in America antennas of the dipole or doublet type are placed with their length horizontally.

As a broad guide, receiving antennas should be placed as high as possible, in the line of sight toward the transmitting station and with directivity toward that station, and as far back on the roof of a building as may be conveniently possible and thus away and partly shielded from streets carrying heavy traffic, in order to minimize the reception of electrical disturbances originating in automobile ignition systems.

One novel factor in television reception is the practical importance of avoiding video echoes. If a television signal reaches the receiver over two paths having different lengths from the transmitter, there will be produced two images separated to an extent dependent upon the path difference. The modulation frequencies in television are so high that the echo effect can readily become noticeable. It can be shown that the horizontal displacement between the original picture and the echo picture, expressed as the corresponding number of picture-element widths, is closely equal to the path difference in feet between the two received wave trains divided by 125. The picture produced on a 12″ kinescope (assumed to be 7.5″ by 10″) has a picture-element width of about 0.017″. It is clear that path differences of the order of several hundred feet will give noticeable doubling of the image, and that even smaller differences may detract from picture definition. Accordingly, it is important in television installations that the antenna shall not respond to powerful reflected waves (or else *only* to them), and also that the antenna and its transmission line shall be so terminated as to avoid repeated reflections in that part of the receiving system.

When trouble is experienced from television echoes in a particular installation, a directional receiving antenna can sometimes be rotated so as to eliminate either the reflected or the direct signal, selecting whichever may be most practical in that specific case. It is not unusual to find a reflected signal which is stronger than the direct signal and which should therefore be selected for reception, if possible. In determining the most suitable receiving arrangements, a convenient test means is to mount a light doublet in portable fashion on the end of a long pole, and thus to determine experimentally the preferred location and orientation of the antenna. Sometimes improvements result if a second unloaded dipole is placed near the receiving dipole in order to reduce the effect of reflecting surfaces or to increase the reception of the direct wave.

Reception is generally fairly simple at distances well within the service range, but becomes somewhat more of a problem toward the boundaries of the service area where electrical noises may cause trouble. Where such difficulties are experienced, an increase in the generated antenna signal voltage and greater directivity of the antenna system are desirable. One method of securing these results is to use long-wire antennas. There are practical limitations to this method. Thus, an antenna which is 8 times as long as the simple dipole (that is, which is 4 wave-lengths long), produces only about twice the signal obtained from the dipole, and further, it produces this full signal only from transmitters lying in four particular directions. When using such long antennas, it is sometimes requisite to damp the antenna either by a terminating resistor or otherwise in order to secure wide-band response and also to prevent excessive delay before the antenna currents reach a steady-state condition.

The so-called "vee beam" antenna consists of two straight arms, each of one wave length, which are mounted at an angle to each other. The center of the open angle should point toward the transmitting station, reception being best in that direction and least at right angles thereto. The receiving characteristic is bi-directional, and the directivity is superior to that of a simple dipole.

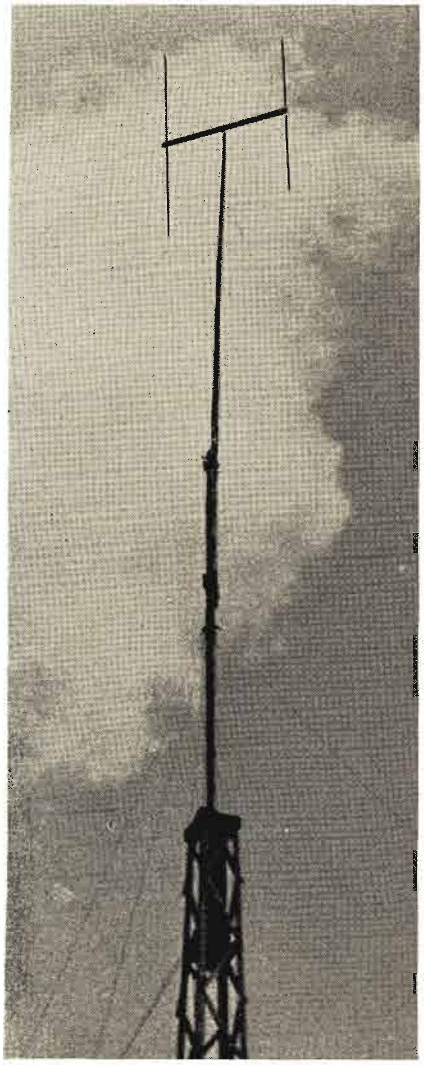
Another useful form is the diamond or rhombic antenna. This consists of 4 arms, each of 1 wave length. The transmission line connects to one of the acute angles of the diamond, and a terminating impedance is inserted in the other acute angle (this impedance having a theoretical value of 700 ohms for the surge impedance, but giving satisfactory operation in practice for values in the range from 400 to 800 ohms). The rhombic antenna is mono-directional and has the advantageous feature of acceptably receiving a 2.5-to-1 frequency band.

The double-doublet antenna, while less directional than the rhombic antenna, also covers a similar frequency range when provided with an appropriate transmission line.

Practice in antenna installations has not been standardized. One recommendation has been to use a half-wave-length dipole for more distant locations. So far as economy in installation and minimum space requirements are concerned, the antennas range in the following order: simple dipole, vee-beam, double dipole, rhombic antenna.

Near the transmitting station, excessively powerful signals may overload the receiving set, producing, for example, cross modulation between the picture and sound in the first receiving tube. When this readily recognizable effect occurs, it becomes necessary to reduce the signal voltages reaching this tube by the insertion of circuit resistance in such fashion as not to cause transmission-line impedance mismatching and consequent reflection echoes.

It is particularly necessary at points where weak signals are involved that directional antennas of considerable length shall be connected to the receiver through transmission lines with careful impedance matching since not only may echoes otherwise be received but reflected waves in an improperly engineered system of this sort may even throw the receiver out of synchronism through reflected pulses. Receiving antennas for television tend more to be a "hand-tailored" job than for audio installations. They also tend to favor reception from a single station and for a limited frequency band to a greater extent. These limitations are particularly troublesome in locations far removed from the transmitter or where



A television receiving antenna of Pye Ltd., at Middlesbrough, England.

electrical disturbances are particularly heavy. Not only should the antenna be electrically efficient but it should also be sightly and resistant to the weather as well as unchanging in its electrical characteristics. It is usually entirely justifiable to spend more for antenna materials and installation in a television instance than for audio broadcasting. The electrical requirements are more stringent, and the receiving equipment more costly; accordingly it is poor economy to use low-efficiency television receiving antennas.

G-2 Antenna Reflectors

As mentioned above, reflectors back of an antenna will sometimes increase the signal strength and reduce static coming from other directions than the signal. This applies as well to directional antenna systems. Static and electrical disturbances show themselves in various ways. Most disturbances (for example, those from automobile ignition systems) cause "spottiness," that is, transient white or black splotches generally extending parallel to the scanning lines of the picture and irregularly distributed over the picture. At best they detract from entertainment value, and at worst they practically ruin picture quality. The necessary signal strength to over-ride disturbances from automobile ignition systems has been stated to be at least 5 mv/m in city locations. When only more usual disturbances and those inherent in the receiver are considered, a satisfactory picture is stated to be obtained with a signal strength of 1 mv/m. Suburban and rural signal strengths above 5-10 microvolts/m have been said to be satisfactory, but such values represent a lower limit of service.

Interference from diathermy generally appears as broad dark bands horizontally across the picture which may obliterate the picture altogether, or as "watered-silk" patterns which produce a peculiar shimmering or wavering of the picture. In either case, audience enjoyment is reduced or destroyed. In view of the relatively high power of diathermy outfits, and their frequent proximity to the television receiver, there seems to be no practicable solution for the elimination of such interference other than careful shielding (and even oscillator-frequency selection) of the diathermy outfit. Inherently the gen-

(*Continued on page* 35)

FUNDAMENTALS OF TELEVISION ENGINEERING

(*Continued from page* 20)

or performance. A photograph of this tube is shown in Fig. 5, and a sectioned schematic diagram is shown in Fig. 6.

Referring to Fig. 6, the image to be televised is focused upon the plane photo-electric cathode near the end of the tube. By virtue of the potential existing between the anode and this cathode, an electronic image is released from the opposite side of the cathode. With the aid of special focusing arrangements, the electronic image impinges upon the mosaic at the opposite end of the tube. The only major difference between this mosaic and the one in the basic Iconoscope is that this one is not treated for photo-electric emission. The electronic image hitting the mosaic knocks off secondary electrons from the globules. In this manner, the potential distribution corresponding to the image distribution of light and shade is set up over the face of the mosaic. The secondary electrons are carried off by the anode and leave a deficiency of electrons or a positive charge on each tiny condenser which each globule forms with the metallic backing plate. The value of these charges depend upon the number of secondary electrons given off, and this in turn depends upon the number of photo-electrons representing that particular part of the electronic image. The electron gun and deflecting system scans the mosaic in the usual way, and the signal is taken off as in the ordinary Iconoscope.

The advantages of this pick-up tube lie mainly in the fact that the sensitivity is increased to about ten times that of the old Iconoscope due to the fact that secondary emission is more effective than photo-emission in building up the charges on the mosaic. Another advantage is that the photo-cathode is so close to the end of the tube. This allows the use of short focal length lenses, and, consequently, a large aperture optical system can be used. The spurious shading signal generation effect is still present in this tube, though in at least some cases is slightly less severe.

(*To be continued*)

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INSULATION IN TELEVISION

IT is at once recognized that the design and construction of television receiving sets, as projected up-to-date, imposes new and higher standards of electrical insulation.

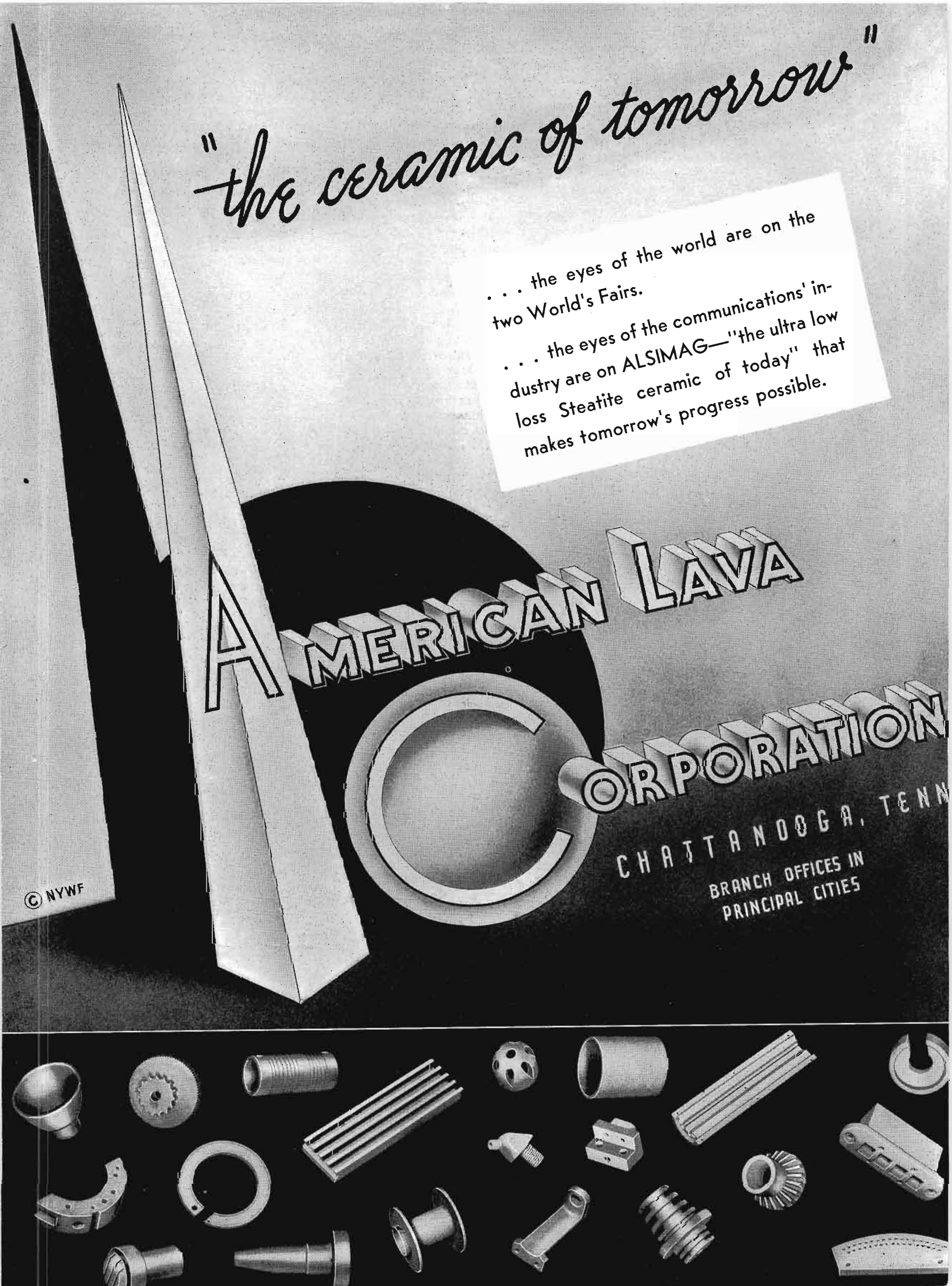
The reasons underlying the necessity for such new standards become apparent upon even a casual consideration. One need only realize that in the television development, we have a piece of apparatus, to be placed in the hands of a curiously-minded public, carrying voltages of from ten to twenty times those of the ordinary radio receiving sets. What is perhaps even more important in support of this broad consideration is the terrifically high step-up in frequencies to be handled in television sets; an increase of from one million cycles in radio to 50 million cycles in television. In dealing with such a frequency, it becomes at once evident that the minutest electrical energy loss is likely to be translated into a tangible loss of efficiency or precision in the finally reflected image. Accordingly, in the transmission specifications, the absolute of "no loss" would become almost an essential stipulation. In accomplishing such an ultra electrical transmission and transformation requirement, insulation in its qualitative phase immediately becomes of primary and paramount importance.

Aside from its energy-conservation purpose in television, there is another phase of practical application which projects the quality factor of electrical insulation to the forefront in television. Whereas in the ordinary radio receiving set the effect of stray currents eluding their conductors results in nothing more than an audible disturbance, or perchance a short-circuit, readily remedied, a similar loss in a television set is likely to result in a corona effect so elusive as to reflect itself in a manner to destroy the pictorial precision so essential to ultimate successful imagery. Under such circumstances, the question of electrical insulation becomes a distinctly critical one, and must be approached as such. The designer, therefore, must be on guard constantly, to feel certain his specifications are not only exacting and precise, but, moreover, include a wider safety factor than one commonly considered sufficient in regular radio work.

It is deemed advisable to sound this cautionary note, so far as electrical insulation is concerned in a television setup. While because of design standard specifications, the elimination of hum patterns, as the result of magnetic fields, is of remote concern so far as extra insulation is concerned, there is this to be said. The protection afforded by mechanical means and conventional forms of shielding, may create a sense of false security. From such an abstract point of view alone, to resort to additional safeguarding in the electrical insulation phase, qualitatively as well as quantitatively, particularly as the cost factor is relatively so small when considered over all, would seem a justifiable cautionary step. The single high-voltage lead from the power pack is a case in point. Perhaps a lining adjunct to the critical shielding points would not be amiss in a television circuit, purely as an extra factor. Similarly to magnetic fields, leakage flux from the transformer might form a hum pattern and as such must be guarded against.

In a measure, the interlocking switch to be universally adopted for television sets, is only a primary safeguard. With this in mind, all phases should be supplemented in an electrically protective fashion, with as wide a safety factor as can be conceded practical.

WILLIAM NAUMBURG, JR.,
William Brand & Co.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IMPEDANCE MEASUREMENTS

(*Continued from page* 9)

which cause error, however, are generally self- and mutual-inductances, rather than capacitances. To illustrate the nature of the errors in a manner similar to that for the series-resonance methods, the conductance-variation circuit can be chosen. It is shown schematically in Fig. 11.

The two residual parameters shown, namely the mutual inductance, M_1, and self-inductance, L_2, are chosen as illustrative because they cause errors in the parallel-resonance methods that are similar to those caused by the residual capacitances, C_1, and C_2, in the series-resonance methods.

The mutual inductance, M_1, causes a component of voltage to be induced in the meter circuit. The voltage at the meter terminals is therefore different from that occurring across the rest of the measuring circuit and errors occurs if the simple theory, which is based on equal voltages, is used. The use of short leads and electromagnetic shielding ordinarily makes the error from this source small at broadcast frequencies.

The inductance, L_2, causes the effective capacitance across which the unknown impedance is connected to be greater than the static capacitance because of its resonating effect. This increase in apparent capacitance causes error in the determination of susceptance. As before a concrete example shows the nature of the error.

Suppose that $L_2 = 0.1$ µh and that the two settings of the sandard condenser with the unknown impedance, Z_x, in and out of circuit are 950 mmfd and 900 mmfd. The susceptance difference, as read from the condenser dial, is

$$\Delta B = \omega(950\text{-}900) \times 10^{-12} = \omega \times 10^{-12} \times 50\,.$$

The true susceptance difference is

$$\Delta B = \omega \left(\frac{950 \times 10^{-12}}{1 - \omega^2 L \times 950 \times 10^{-12}} \times \cdot \frac{900 \times 10^{-12}}{1 - \omega^2 L \times 900 \times 10^{-12}} \right)$$

$$= \omega \times 10^{-12} \frac{50}{1 - \omega^2 \times 0.1 \times 10^{-6}\,(950 + 900) \times 10^{-12}}$$

$$= \omega \times 10^{-12} \frac{50}{1 - 0.073}$$ at a frequency of 1,000 kc, and the error caused by neglecting L_2 is again about 7%.

As in the case of the series-resonance methods, the effect of close coupling can be easily analyzed if the coupling does not affect the source frequency and if the source output impedance is essentially linear. According to a dual form of Thévenin's theorem, that part of the circuit to the left of the section B-B in Fig. 11 can be replaced by an equivalent circuit consisting of a constant current, $I_{s/c}$, equal to the short-circuit current at the section B-B in shunt with a constant impedance equal to that seen looking toward the high-frequency power source from the section B-B with all active voltages zero. Fig. 11 can therefore be redrawn as shown in Fig. 12.

If the coupling capacitance, C_c, is infinitesimal, as assumed in the simple theory, the circuit conductance which is measured is simply the parallel conductance of the measuring circuit. If the coupling is finite, a certain amount of conductance is coupled into the measuring circuit from the power source, and the circuit conductance measured includes a component that can be ascribed to close coupling. As in the case of the series-resonance methods, however, if the assumptions of constant frequency and linear output impedance are satisfied, and if the coupling is held constant, no error need arise from close coupling when an unknown impedance is measured. If determinations of circuit conductance are made with the unknown impedance, Z_x, in and out of circuit, the difference will be equal to the conductance of the unknown, and the difference between the two resonant values of condenser susceptance will be equal to the susceptance of the unknown.

Bridge Methods

The major errors in bridge methods are caused by residual parameters, and it is in the methods of eliminating and compensating these unwanted components that the design of high-frequency bridges differs from that of low-frequency bridges.

A brief description of the compensation used in the General Radio Type 516-C radio-frequency bridge serves to illustrate the nature of the residual parameters. A circuit diagram is given in Fig. 13.

In the lower left-hand bridge arm, the capacitance, C_{NO}, and the inductance, L_N, are residual parameters. The residual capacitance is the sum of the capacitance to ground of one side of the standard condenser, C_N, and capacitances to ground of the output transformer and upper left-hand bridge arm. The residual inductance occurs largely in the compensated decade resistor, R_N.

Each of the residuals causes error if the unknown impedance is connected simply between the lower and the right-hand corners of the bridge with no compensating devices and the bridge is balanced as a direct impedance bridge. The errors caused by the residuals, however, can be avoided by connecting in circuit the compensating coil, L_P, and the compensating condenser, C_{PO}. When $C_{PO} = C_{NO}$ and $L_P = L_N$, the bridge arms become identical at balance and no errors arise from the residual parameters.

(*To be continued*)

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RCA BULLETINS

A number of new bulletins are now available from the RCA Manufacturing Co., Inc., Camden, N. J. One bulletin deals with the RCA monitoring amplifier Type 94-D, a second with the 76-B consolette speech-input system, while a third covers the 5-DX five-kilowatt broadcast transmitter. All three are available from the Broadcast Equipment Division of the above organization.

Technical information and service data on the Model AVT-7B-1 aircraft transmitter may be secured from the Aviation Radio Section. A specification sheet on the 303-A frequency-limit monitor can be obtained from the Engineering Products Division.

• • •

TELEVISION CAPACITORS

Cornell-Dubilier engineers have released to the trade a series of capacitors that have been incorporated into the design of television receivers. Cornell-Dubilier engineers are supplying these capacitors to required specifications. This series has been tested in actual television receivers in the C-D laboratories, and are now in use in models being produced by the trade. *Cornell-Dubilier Electric Corp.*, South Plainfield, N. J.—COMMUNICATIONS.

• • •

POLICE GENEMOTOR

Carter has placed on the market a new police type genemotor which incorporates a starting relay as an integral part of the unit. This genemotor is of the high-voltage type and designed for continuous service. It is ready to connect to the transmitter and needs no adjustments. As a safety factor oil condensers of the high-voltage type are used. *The Carter Motor Co.*, 1608 Milwaukee Ave., Chicago, Ill.—COMMUNICATIONS.

MOTION PICTURE FILM

(*Continued from page* 17)

sists of a pattern of parallel black lines separated by equal white spaces. Therefore each "line" in the resolving power test corresponds to two lines in the television scanning pattern.

The "circle of confusion" of a lens, such as is used for photography or for projecting a picture, is the image, always somewhat imperfect, of a point of the subject as produced by the lens in the plane on which the image is focused. Stating the diameter of the circle of confusion of a lens is like stating the diameter of the spot produced by the electron beam where it strikes the fluorescent screen in the television receiver. In order to determine whether the definition is good or bad we need also to know in the one case the focal length of the lens and in the other case the size of the screen on which the picture is being formed.

The resolving powers of the films at present available for 16-mm picture photography lie in the range from 40 lines per millimeter to 60 lines per millimeter. The newest films combine high speed with fine-grain structure and resolving powers of the order of 50 lines per millimeter, a value that was formerly considered high. The trend of film research is toward steadily improving resolving power.

The height of the standard 16-mm projector aperture is 7.21 millimeters. A little arithmetic shows, then, that a frame of 16-mm film having a resolving power of 40 lines per millimeter is capable of showing 288 separate lines in the height of the picture, which is equivalent to the detail of a 576-line television image. A similar frame of a film having a resolving power of 50 lines per millimeter is capable of recording detail corresponding to a 721-line television image, while a frame of film having a resolving power of 60 lines per millimeter is capable of recording detail corresponding to an 865-line television image. Therefore, as far as the detail rendering power of the photographic emulsion is concerned, 16-mm films are more than adequate for 441-line television.

Other factors which limit the fineness of definition obtained in a motion picture besides the imperfections of the images introduced by the taking and projection lenses, are inaccuracies of registration and poor contact in printing, and unsteadiness of the pull-down mechanisms in the camera and projector. All of these deteriorating influences are cumulative. In ordinary 16-mm motion-picture photography, carried on without special precautions, the accumulated blurring of the image generally amounts to as much as a thousandth of an inch, or 1/284 of the height of the picture. This is obviously not good enough for the highest quality television, though it is probably as good as the images being broadcast at the present time.

Fortunately it is by no means impossible to retain a higher standard of definition in 16-mm films. Modern high-speed camera lenses of one and two-inch focal lengths used at their maximum apertures give circles of confusion about a thousandth of an inch in diameter, but when these lenses are partly stopped down, the excellent corrections which made the high speeds possible result in the formation of images having detail finer than the film can record. For example, one well known lens of maximum aperture f1.4 gives a circle of confusion smaller than .00025 inch when used at apertures in the range from f2.8 to f5.6. These apertures give ample speed for photography under average studio lighting. Therefore the first step in producing 16-mm films of high definition is to use high-speed, well-corrected camera lenses, and keep them always stopped down to at least two stops below their maximum openings. Naturally focusing must be accurate.

In order to obtain the highest degree of steadiness in the taking of the picture, the camera should be equipped with a registration pin which enters one of the sprocket holes of the film and holds it in a definite location during the exposure. With this refinement the film can be located within about .0002 inch of its proper place for each exposure. Amateur 16-mm cameras are not equipped with registration pins on account of the cost, since the pin and its actuating mechanism must be very accurately made if it is to accomplish its purpose.

If the printer used to make duplicates from a film photographed with the precautions above described is of the step type and also has a registration pin operating in the same sprocket hole as was used to locate the picture frame in the camera, accuracy or inaccuracy of film perforation in the original film does not affect the steadiness of the picture in the print. If, however, as is often the case, the registration pin in the printer is not at the same distance from the picture aperture as the pin in the camera, so that the registration of the film for the printing of a given picture frame is determined by a different sprocket hole, it becomes necessary to have the perforations spaced very accurately if the picture is to be free from "jump." Fortunately most of the 16-mm film now on the market is perforated with an accuracy of .0002 inch or better, though the writer has examined specimens in which some of the holes were misplaced as much as .001 inch. It may be found necessary to make special tests for perforation accuracy on film that is to be used for television subjects.

Film types commonly used today for making prints have resolving powers (as previously defined) of 80 lines per millimeter or better. For the sake of improved sound track performance, however, it is likely that prints for television broadcasting will be made on special film stocks of very much finer grain than those now in general use. These films have much higher resolving power, of the order of 200 lines per millimeter. In either case the resolving power of the stock used for making the print is so much higher than that of the stock used for the original that there is very little loss of sharpness in printing, provided proper mechanical care is taken to insure good contact and registration of the two films.

With proper attention to the refinements enumerated above, a standard of definition corresponding to a resolving power of forty lines per millimeter, or better, can be maintained through to the 16-mm print. As has been shown, this provides more detail than can be transmitted by a 441-line television system.

Film weave, or side motion of the picture on the screen, commonly results from inaccurate slitting of the film, resulting in variations of width. Weave appears most objectionably when the film is guided by one edge in the camera and by the opposite edge in the printer or projector. This defect can be reduced to a negligible amount by proper design of the side guiding mechanism throughout the system, even when using inaccurately cut film, but it is possible today to obtain film in which the width is maintained so accurately that no trouble is experienced even on machines which guide incorrectly. It may be mentioned in passing that standard specifications for the edge guiding of 16-mm sound films were published by the Society of Motion Picture Engineers in 1934, but have been disregarded by certain manufacturers. It is to be hoped that engineers who have occasion to design film handling apparatus for television purposes will make themselves acquainted with the standards of this Society, and adhere to them as far as possible in the equipment they produce.

Assuming that films have been photographed in a camera having pin registration, and printed in a step printer which also registers the films by a pin operating in the same relation to the picture aperture as the pin in the camera, any "jump," or vertical motion of the final picture on the screen is attributable either to perforation inaccuracy in the stock used for the final

(*Continued on page* 30)

VETERAN WIRELESS OPERATORS ASSOCIATION NEWS

W. J. McGONIGLE, President | RCA Building, 30 Rockefeller Plaza, New York, N. Y. | H. H. PARKER, Secretary

DE FOREST DAY

IN honor of Dr. Lee de Forest's outstanding contributions to the science of radio and the "World of Tomorrow" tribute will be paid him at the New York World's Fair 1939 on Friday, September 22nd, 1939. The day will be known as de Forest Day and an appropriate program will be arranged culminating in a Jubilee Dinner Friday evening either at one of the novel restaurants at the Fair or at one of the more prominent hotels in New York. Again we urge all interested organizations to communicate with us in this connection so that a truly worthy tribute may result. Dinner tickets will be available as soon as a place is decided upon and the price determined. Your inquiry is cordially solicited.

SCHOLARSHIP

The Marconi Memorial Scholarship in Radio Engineering at RCA Institutes to be given by our Association will be presented in conjunction with our activity at the World's Fair. The winner will be selected by the American Institute of Science from among participants in their many science activities. The American Institute has an elaborate exhibit in conjunction with the Westinghouse exhibit at the Fair. A cordial reception is assured all visiting the exhibit.

AT THE FAIR

Among our members at the Fair are Geo. P. Smith, Jr., Executive in the Concessions department of the Fair; George Clark, Past President and Director of the RCA Exhibit; Arthur Lynch, Managing Director of the World's Fair Radio Club and Oscar Oehman, Operator-in-Charge at W2USA—the amateur station at the Fair. When you visit the Fair—and you're missing an opportunity of a lifetime if you don't—look any or all of them up; they'll be glad to see you. Make it in September, if you can, in order to be there on de Forest Day.

HONOLULU

Included in an interesting letter from W. Dietz, retiring Secretary of the Honolulu Chapter, are some details of their cruise on the 11th of February.

"The Honolulu Surfboard, Grass Skirt, Hula-Hula Chapter of the Veteran Wireless Operators Association celebrated the 14th Annual Cruise with a barbecue steak dinner at the residence of George Street, Chairman. If there is anything for which George Street is more noted than his barbecue steak it may well be his vegetable salad.

"Altogether there were representatives from the majority of the commercial communication companies of the islands and a goodly representation from the United States Navy.

"After the inner man was fed the meeting was called to order and Street gave a short talk on the ideals and purposes of our Association. The meeting was then open to general business and the election of officers for the ensuing year. A motion was made and seconded that Chairman Street be declared re-elected, which motion was carried unanimously. H. F. McIntosh of Mackay Radio was elected Secretary.

"Among those present were: J. A. Alverson, Territorial Radio Inspector; F. T. Bowen, CRM, USN, Wailupe; W. E. Chadwick, Press Wireless; Mr. Daigle, CRM, USAT Henderson; W. W. Dietz, RCAC; H. E. Flowers, USN; Otto Horning; Ernest P. Lewis, CRM, USN, Wailupe; Frank J. Murray, CRM, USN, Wailupe; H. F. McIntosh, Mackay Radio; A. W. Nelson, Press Wireless; S. B. Maddams, Mackay Radio; Geo. Richardson, Globe Wireless; George Street, RCAC; Robert M. Wood, USN, Wailupe; Mr. Watford, CRM, USAT Henderson and others.

"A photograph of some of the group is enclosed.

"A list of new members together with renewals and the money to cover is enclosed.

"Before saying 30 and before I turn the business over to Mr. McIntosh I want to wish you and the VWOA a bang-up 1939, to McGonigle and all our friends 73. Signed: W. Dietz, Sec'y."

A fine job, Mr. Dietz. Be assured of our sincere appreciation of your cooperation in the past and our best wishes to Messrs. Street and McIntosh for success in the coming year.

Honolulu gathering, left to right: Richardson, Street, Wood, McIntosh, Nelson, Murray, Alverson, Lewis, Maddams, Bowen, Watford, Flowers, Daigle, Dietz.

CANADA

An interesting excerpt from a recent letter from D. R. P. Coats, Public Relations Manager of CKY, one of the larger Canadian Radio Broadcasting stations. He writes: "For my part, I value my membership in the VWOA very highly. Through publication of my name in COMMUNICATIONS I have renewed friendships with a number of old friends. Only a day or two ago I received a letter from G. Tyler, ex-wireless operator of the *Teutonic* and various other ships. He is now in the Marconi Company's offices in London. Tyler studied beside me at the British School of Telegraphy, London, in 1911. He saw my name in COMMUNICATIONS and wondered if I might be the man he knew so many years ago. I could mention several other instances in which I have thus been put in contact with old friends through the VWOA section in COMMUNICATIONS. The magazine itself is well worth the price of membership, but the VWOA page, for the reason I have stated, is priceless."

PERSONALS

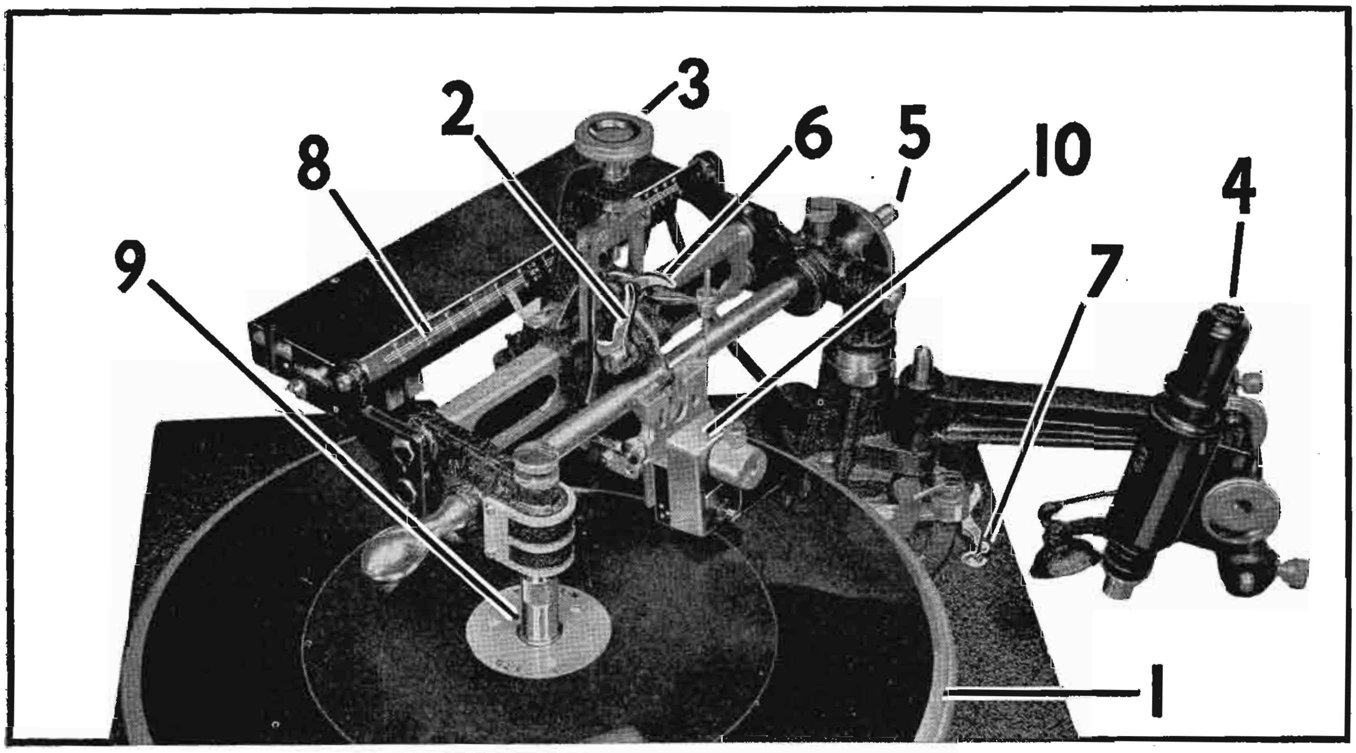
AN OLD TIMER, Mr. Carl E. Braun, ex-Chief Radio Man, 1917, on the U.S.S. *South Dakota,* writes from Oregon City, Oregon. Mr. Braun is now Vice-President of the Hawley Pulp and Paper Company, Oregon City.

ANYONE looking for G. G. Greene can catch up with him by hanging around P. O. Box 463, East Hampton, N. Y.

MEMBERSHIP

In order to retain your standing in the Association we earnestly request you to send in the necessary amount to bring you up to date. If in doubt as to your status, a post card to the secretary will bring the required data. Your cooperation in this matter will further the progress of our Association and materially reduce the work of the secretary. Please let us hear from you soon if you haven't received your 1939 card yet. Thank you.

PRESTO ANNOUNCES THEIR NEW 8-A COMMERCIAL RECORDING TURNTABLE



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3. Thumbscrew rapidly adjusts angle of needle while cutting.

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5. Spiralling feed screw makes starting and run-out grooves.

6. Lever engages cutter carriage with feed screw.

7. Combination speed change and motor switch lever prevents flats on idler wheels.

8. Four-sided, rotating scale shows recording time at 96, 112, 120 and 140 lines per inch at 78 and 33⅓ RPM.

9. Flangeless drive permits quick removal of shavings.

10. Improved high fidelity cutter records uniformly a frequency range from 40 to 8,000 cycles and gives 4 db higher playback level.

11. New, vibrationless motor.

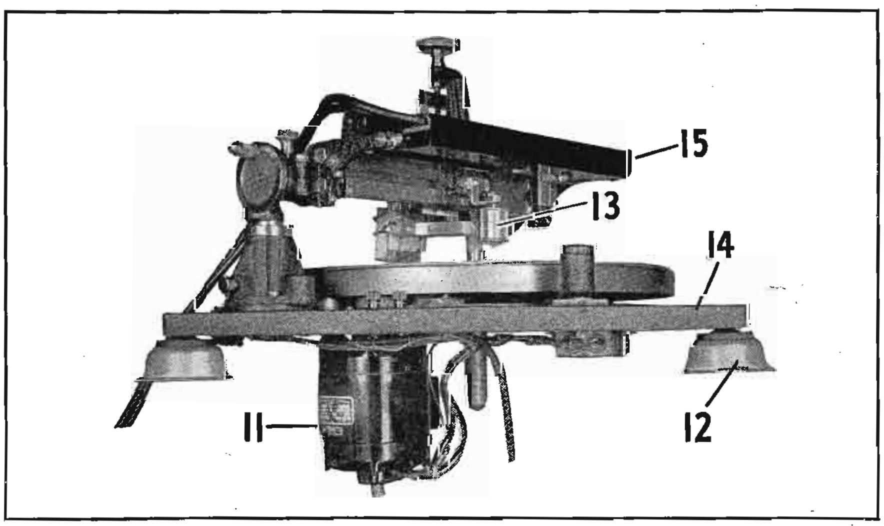
12. Shockproof mountings suppress vibration from outside sources.

13. Vertical damper suppresses transient modulation, eliminates flutter, prevents patterns due to vibration or surface irregularity in discs.

14. Sixty-pound cast iron base assures permanent alignment of turntable and cutting mechanism.

15. Automatic equalizer assures full frequency response range throughout 33⅓ RPM recordings.

Mounting dimensions of the new Presto 8-A recorder are the same as the Presto 6-C and 6-D portable recorders. A liberal trade-in allowance will be made to radio stations and studios that wish to bring their recording facilities up to date. In writing give type and serial number of your present turntables.



PRESTO

RECORDING CORPORATION 242 West 55th Street, New York, N. Y.

FILMS IN TELEVISION

(*Continued from page 27*)

print or to inaccuracies in the projection machine or scanner. Jump due to perforation inaccuracies may be expected to be of the order of 1/10 per cent or less of the picture height, which is well under half the width of a scanning line. Jump due to the scanning machine mechanism can be kept to the same order of magnitude as that in the camera by proper design and accurate workmanship. This accurate workmanship can readily be justified in the case of a television scanner, which does not have to be sold at the low price of the usual 16-mm projector.

It may be seen from the facts as outlined above that the requirements of picture detail and steadiness in 441-line television can be met with 16-mm film provided certain refinements are provided in the equipment used for taking, printing, and projecting the films. These refinements do not add appreciably to the difficulty of operating this equipment, and therefore the writer considers them practical.

The next article of the series will take up requirements of sound track.

(*To be continued*)

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MOLDED CABINETS

(*Continued from page* 15)

different problem, but one which can be satisfactorily worked out by the mold engineers if given the opportunity by the manufacturer.

Engineers experienced in plastics have created certain standards which assure uniformity of production and which guard against loss by breakage, distortion, etc. However, where the radio manufacturer is not entirely sure of his design it is sometimes advantageous to make a preliminary one impression mold from which, if additional impressions are desired, duplicates of the original impression can be made at a considerable saving over the cost of the first impression. Modern machine cutting equipment eliminates time and inaccuracy by using duplicates and original templets.

In the development of any radio case or cabinet the same general fundamentals may obtain, but at the same time each design will have certain problems perculiar to itself alone and these have to be worked out carefully in order to get satisfactory results. If the manufacturer of radios will keep this in mind and remember to bring his problems to the molder before definitely deciding on final design, he will be able to make use of the specific knowledge obtained from experience by the raw material manufacturers, the mold designer and the molder's engineering staff.

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- √ Maximum heat dissipation due to cement coating in closest bond with wire winding. More than generous wattage ratings.
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CONTROL NETWORKS

(*Continued from page* 13)

gain of 2.5 times. At 50 cycles, this is equal to a rise of 7.5 times. Consequently, by substituting quantities in expression (8), and without needlessly repeating the processes results in

$$C = .1775 \text{ mfd}$$

$$= .2 \text{ mfd (nearest standard value)}$$

and,

$$\mu = 6.6, \text{ and a gain of } 2.28 \text{ or } 7.2 \text{ db.}$$

Since the ear cannot discriminate between transitions of 1.5 db, the value 7.2 is satisfactory in that the pre-calculated figure was 8 decibels.

It is instructive to note that all the calculations have been considerably simplified for commercial purposes. Should all the minor variations been computed into the network systems, it would have resulted in high intricate design formula having little practical value. In general, the expressions used in the configuration designs are sufficiently accurate for all standard applications.

INCREASING TREBLE RESPONSE

The arrangement of the components figured in a network for high-frequency emphasis is diagrammatically pictured in Fig. 3. The amplification ratio at any high audible frequency E_2/E_1 is secured from the simple relation

$$E_2/E_1 = \frac{\sqrt{R^2 + f^2 \times L^2}}{\sqrt{(R + R')^2 + f^2 \times L^2}} \quad ..(9)$$

The value of the inductance L for maximum treble definition at 10,000 cycles (a frequency easily attained by the wide-range electrical transcriptions played via the amplifier) is formulated from:

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CONTROL NETWORKS

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$$L = \frac{\sqrt{\frac{(R+R')^2 \times \mu^2}{670} - R^2}}{f \times \sqrt{1 - \frac{\mu^2}{670}}} \quad ..(10)$$

Inserting values in the above expression for a μ of 20 and 7.5 respectively:

$$L = \frac{\sqrt{\frac{(7500 - 60{,}000)^2 \times 20^2}{670} - 7500^2}}{10{,}000 \times \sqrt{1 - \frac{20^2}{670}}}$$

$= 1.3$ henry at 100 cps for a μ of 20_{max}.
$= 7.5$ henry at 10,000 cps for a μ of 7.5_{min}.

In calculating the values for decreasing the bass response from the normal flat frequency operating curve, it is of interest to note that in Fig. 4, resistors R and R′ have no appreciable effect on the network equilibrium provided R_3 is of a greater magnitude than R. By considering such factors only values C_3 and R_3 need be taken into account.

The decrement of any frequency from the normal flat frequency response curve is computed from the equation:

$$Freq._{ratio} = \frac{1}{1 + \frac{1}{f^2 \times C_2^2 \times R_3^2}} \quad ...(11)$$

The ratio between the frequencies of the middle and lower registers can be easily compared by denoting the letter K in the following formula to have a ratio value, hence

$$K^2 = 1 - \frac{1}{f^2 \times (C_2^2 \times R_3^2)} \quad .....(12)$$

where

$$(C_2 \times R_3) = \frac{1}{f \times \sqrt{\mu^2 - 1}} \quad ......(13)$$

In designing the attenuating networks, the decremental steps should closely approximate that of the gain but in a negative direction. Hence K equals 6.3 and 2.5 at 50 cps, hence

$$C_2 \times R_3 = .000512 \text{ for } 6.3$$
$$= .001385 \text{ for } 2.5$$

Assuming R_3 to be 250,000 ohms, the capacities for the two ranges

$$6.3 = .00204 \text{ or } .002 \text{ mfd.}$$
$$2.5 = .00554 \text{ or } .005 \text{ mfd.}$$

For muting the treble register another set of expressions are used in reference to Fig. 5. The decremental steps are related to the gain and are also in the negative direction.

The diminishment of any high frequency from the normal flat frequency response curve is computed from the equation:

$$\text{Freq.}_{\text{ratio}} = \frac{1}{\sqrt{1 + f^2 \times C_3^2 \times R_x^2}} \qquad (14)$$

where,

$$R_x = \frac{R \times R'}{R + R'}$$

$$= \frac{7500 \times 60{,}000}{7500 - 60{,}000} = 6660 \text{ ohms}$$

And from expression (12), K equals 6.9 and 2.28 at 10,000 cps for the maximum and minimum attenuating steps, hence

$$C_3 \times R_x = \frac{\sqrt{K^2 - 1}}{f} \quad \ldots\ldots\ldots\ldots (15)$$

and the capacities C_3 for the two negative gain ranges will be

6.9 = .00163 mfd or .015 mfd
2.28 = .00491 mfd or .005 mfd

The insertion of the preceding network designs in a circuit for multiple control are shown in Fig. 6. The two-gang five-point switches are connected in tandem; only two two-gang switches are required. The mid-tap setting on both controls isolates the definitive circuits from affecting the normal flat frequency response curve of the amplifier.

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TELEVISION ECONOMICS

(*Continued from page* 23)

eral acceptance and economic success of television thus becomes largely dependent upon the shielding of automobile ignition systems and diathermy outfits. Since television will be established and grow in individual districts, the problem of handling such interference sources on a nation-wide scale, and in a fashion which is fair to all concerned, presents some troublesome angles.

Desirable and preferable as self-regulation would be for the gradual elimination of such interference, its failure might well lead to legislative enactments having the same aim. Some indication that such a trend is possible is found in a European country where official bodies are developing rules for the elimination of interference resulting from such sources as mentioned above. It is planned to place these rules before the local Government for later possible embodiment in the form of law.

The disturbance effect on the picture of relatively near lightning will depend partly on the severity of the storm, the

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number of discharges during a unit time, the distance of the storm, the signal carrier frequency and band width, the signal modulation method, the signal strength and requisite receiver gain, the picture brightness, and the response characteristic (linearity) of the kinescope. A quantitative statement relative to such effects is not now available; qualitatively it may be said that for brief and infrequent periods each year some visible lightning interference with average-strength signals sent by present methods will be noticed.

H. RECEPTION TRANSMISSION LINES

In view of the ultra-high frequencies and correspondingly ultra-short wave lengths used for television, and considering the necessary separation between the receiving antenna and the receiving set, a transmission line becomes necessary. Such a transmission line should be weather-proof in the outside run, sightly, capable of convenient concealment indoors (for example, behind or above picture molding or under carpets or in like locations), and should cause minimum loss of signal voltage or distortion of signal over its length.

The simplest form of transmission line is a twisted pair of insulated conductors. The rubber-insulated type of twisted pair causes a loss of signal strength of approximately 1.5-2 db per wave length at 50 mc. For lengths up to 40 or 50 feet, this is usually acceptable. For longer runs, open-wire lines or coaxial cable may be used, and such transmission lines are desirable as well toward the outer portions of the service area where signal losses in the transmission line may noticeably detract from service. The open-wire lines (consisting of 2 wires 1-2 inches apart) show the diminished loss of about 0.1 db per wave length. From the installation viewpoint, they present greater difficulty.

The transmission line and its terminations at the antenna and receiver require matching. The twisted-pair lines show a low impedance whereas the open-wire lines show a higher impedance. Fortunately, rough impedance matching of the transmission lines is fairly satisfactory. Thus if the resonant impedance of a simple doublet antenna has a theoretical value of 72 ohms, little difficulty is experienced if transmission lines presenting an impedance up to say 130 ohms are connected to such an antenna.

If marked unbalance exists in a transmission line, considerable signal voltage may be picked up in the line itself with a resulting reduction in picture quality. Transmission-line construction is of greater relative importance in video than in audio installations, and it pays to do a thorough job in this part of the television system even at somewhat increased cost.

In urban locations, the suitable antenna location may be within a restricted area, and space for transmission lines may be limited. In large apartment, office, and institutional buildings and the like, it may be most convenient and economical to distribute television signals throughout the building by means of a centralized system. Such distribution may be at radio frequencies, using one or a group of antennas and amplifiers to feed the transmission lines,

(*Continued on page* 38)

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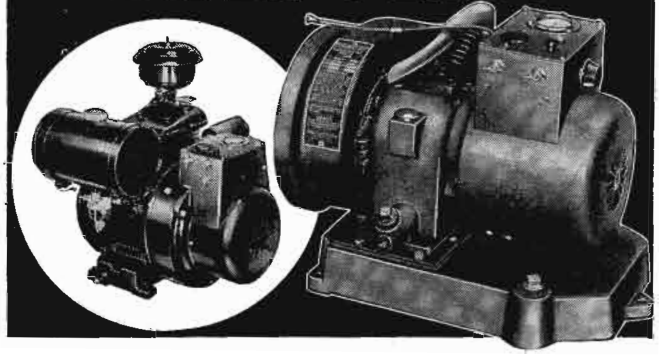

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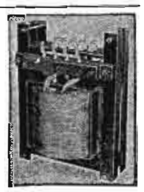

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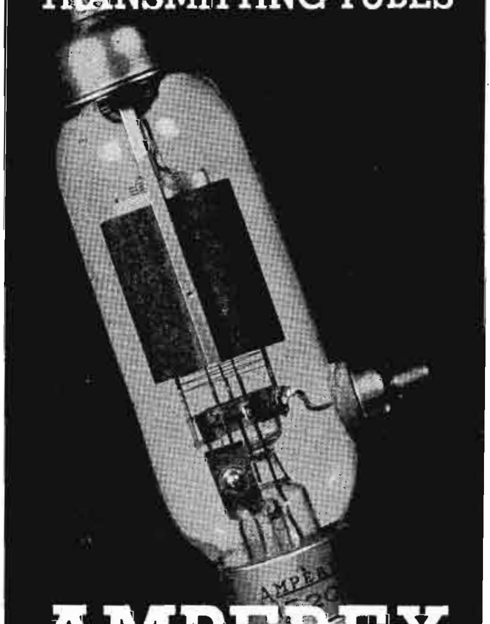

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and with suitable output (or receiver-coupling) circuits throughout the building. Alternatively, several specific programs may be selected and distributed at video frequencies. The former method requires that each user shall have a complete television receiver but ultimately provides full freedom of choice of program to the user. The latter method enables the use of slightly less elaborate terminal equipment for each user, may however somewhat restrict his extent of choice of programs, and has the advantage of transmitting lower frequencies throughout the building with diminished line losses in a properly engineered installation. Nevertheless, such a system requires great care to avoid frequency or phase distortion, as well as internal electrical echoes. Up to the present, centralized television installations have not been developed to the point where definite conclusions can be drawn as to the most economical and desirable form of installation. In part this may depend upon the number of available programs in a given vicinity, the separation of their respective carrier frequencies, the number of individual outlet stations in the building, and the average, minimum, and maximum building runs.

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HENRY HUTCHINS TO NATIONAL UNION

On leave of absence from four-year Vice-Presidency of Western Advertising Agency's Chicago Office, Henry Hutchins has returned to National Union to direct its selling program. It was back in 1930 that Henry was first appointed Sales Manager of N. U., guiding its policies during the formative years of radio parts business until 1935.

JOHNSON PURCHASE

The E. J. Johnson Co., Waseca, Minn., have just completed the purchase of the entire Centralab business pertaining to the manufacture and sale of socket contacts. This is said to include the complete equipment, inventory, good will and patent rights.

CALLITE APPOINTMENTS

Mr. J. Kurtz, Treasurer and General Sales Manager of the Callite Products Division, Eisler Electric Corporation, will personally supervise activities of both the engineering and manufacturing departments. Mr. H. M. Lusk has been elevated to the position of assistant to Mr. J. Kurtz.

CUSHWAY TO WEBSTER-CHICAGO

This year's Radio Parts Show will find Charlie Cushway, veteran sales executive, greeting his friends from the booth of the Webster Company, Chicago manufacturer of sound systems and record-playing equipment. Mr. Cushway was formerly with Thordarson.

REMCO MOVES

Radio Engineering & Manufacturing Company, manufacuturers of "Remco" portable electrical transcription equipment, for five years located at 26 Journal Square, Jersey City, N. J., have moved to their new and enlarged plant at 58 West 25th Street, New York City. Complete information covering their transcription, playback and other portable equipment may be obtained by writing to them.

BENWOOD LINZE ACQUISITION

Effective April 1, 1939, The Benwood Linze Company, St. Louis, Mo., became the domestic and export sales agents for The B-L Electric Manufacturing Company of St. Louis, Mo. No changes in officers, personnel, or corporate structure were made. Coincident with the arrangement by which The Benwood Linze Company handles the sales of B-L Electrical Products, this company acquired the interests of the Brenkert Light Projection Company in the manufacture of rectifier equipment.

DUMONT APPOINTMENTS

With the expansion of activities from cathode-ray tubes, oscillographs and allied equipment, to include television receivers, studio and transmitting equipment, the Allen B. DuMont Labs., Inc., of Passaic, N. J., have advanced Leonard F. Cramer to the post of General Sales Manager. G. Robert Mezger assists Len Cramer in the handling of industrial sales.

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