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JERRY MINTER


**FREQUENCY MODULATION**

**TELEVISION**

**AUGUST**

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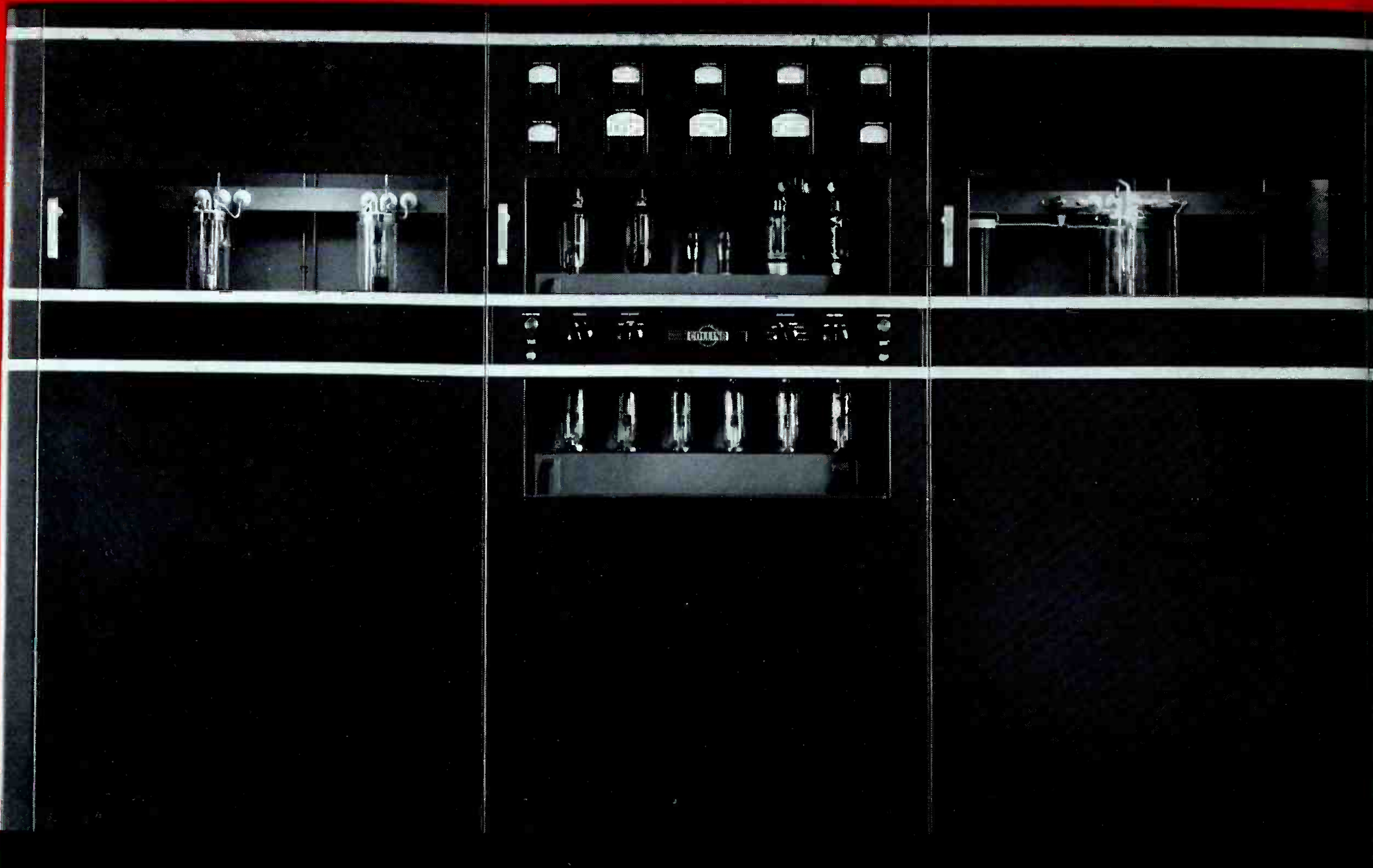

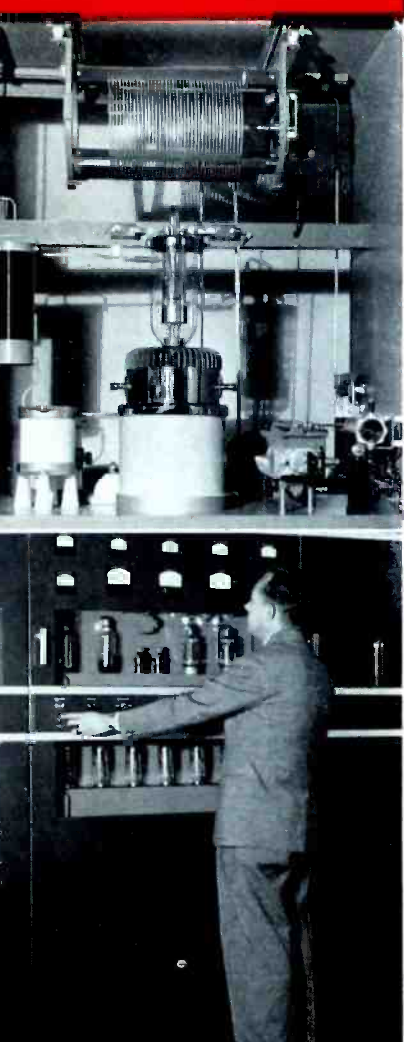

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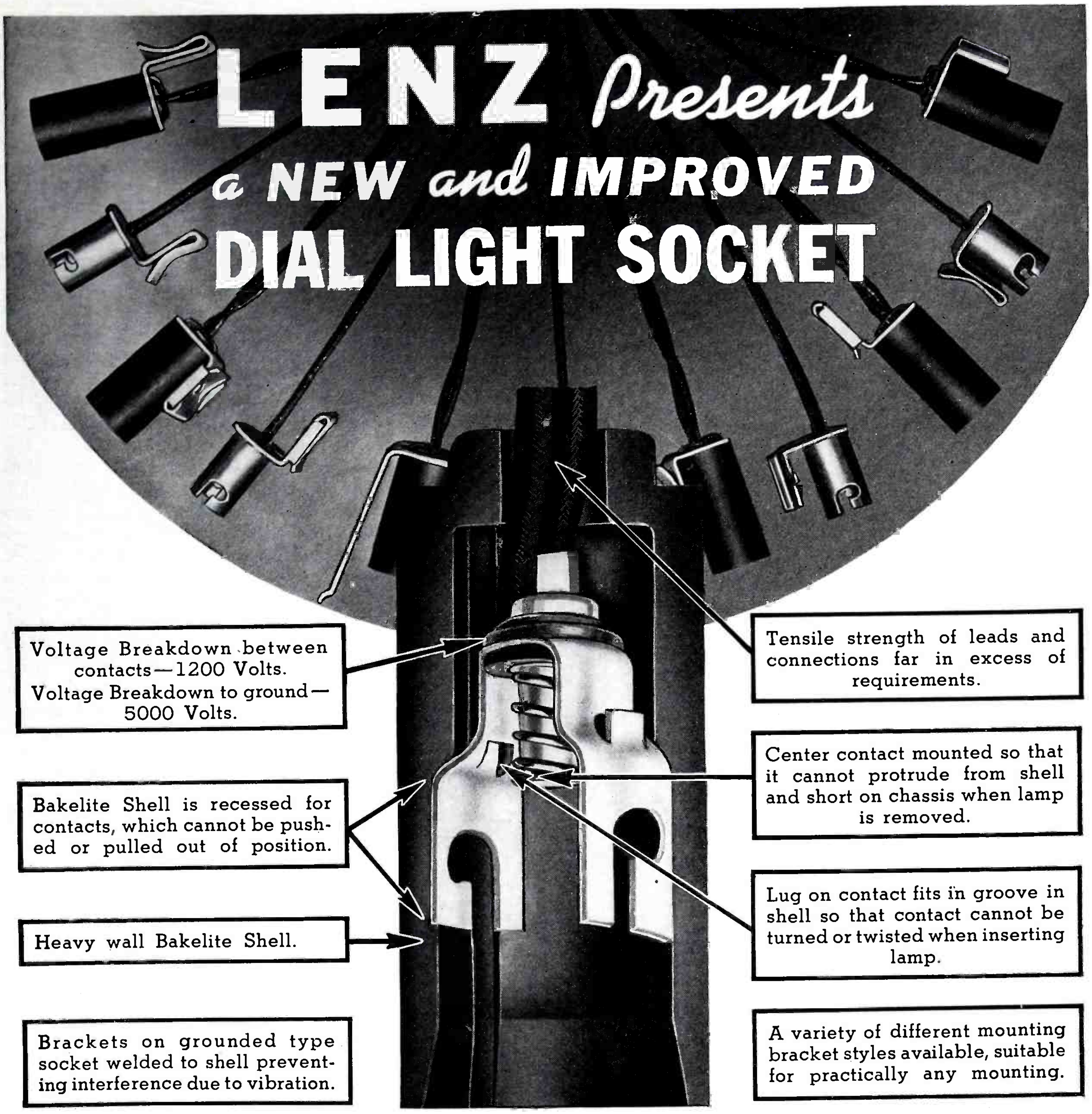


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AUGUST 1940

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VOLUME 20
NUMBER 8

RAY D. RETTENMEYER
Editor

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## • Editorial Comment •

IT finally appears that FCC approved television standards will be obtained—standards which should be satisfactory to the entire television industry. The first step in this direction was taken last month when the organization of the National Television Systems Committee was formed under the auspices of the Radio Manufacturers Association with official FCC cooperation.

The organization of the Committee, as appointed by RMA's President, J. S. Knowlson, is as follows:

Chairman, Dr. W. R. G. Baker, Director of the Engineering Department of the Radio Manufacturers Association; from Bell Telephone Laboratories, A. A. Oswald; Columbia Broadcasting Systems, Adrian Murphy, Executive Director of Television; Don Lee Broadcasting System, Harry R. Lubcke, Director of Television; Allen B. Du Mont Laboratories, Inc., Allen B. Du Mont, President; Farnsworth Television & Radio Corporation, B. Ray Cummings, Vice-President in charge of Engineering; General Electric Company, E. F. W. Alexanderson; Hazeltine Corporation, Daniel E. Harnett, Chief Engineer; John V. L. Hogan, representing consulting engineers; Hughes Tool Company, Albert I. Lodwick; The Institute of Radio Engineers, Inc., Dr. A. N. Goldsmith; Philco Corporation, David B. Smith; Radio Corporation of America, E. W. Engstrom; Stromberg-Carlson Telephone Mfg. Co., Frederic C. Young, Chief Engineer; Zenith Radio Corporation, John R. Howland, Secretary and Assistant to the President.

Details of organization and future procedure were arranged at a New York meeting. Many companies, including research and technical experts, will serve on various "panels" of the committee. Chairman Baker appointed and announced the organization of "panels" or "subcommittees" with their respective chairmen, as follows: System Analysis, P. C. Goldmark; Subjective Aspects, Dr. A. N. Goldsmith; Television Spectra, J. E. Brown; Transmitter Power, E. W. Engstrom; Transmitter Characteristics, Philo T. Farnsworth; Transmitter-Receiver Coordination, I. J. Kaar; Picture Resolution, D. E. Harnett; Synchronization, T. T. Goldsmith; and Radiation Polarization, David B. Smith.



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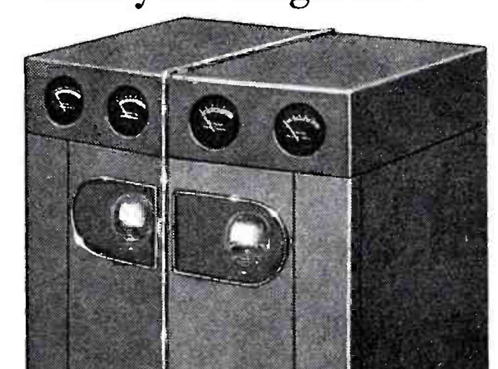

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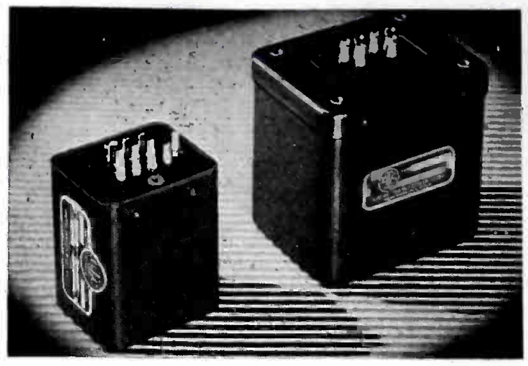

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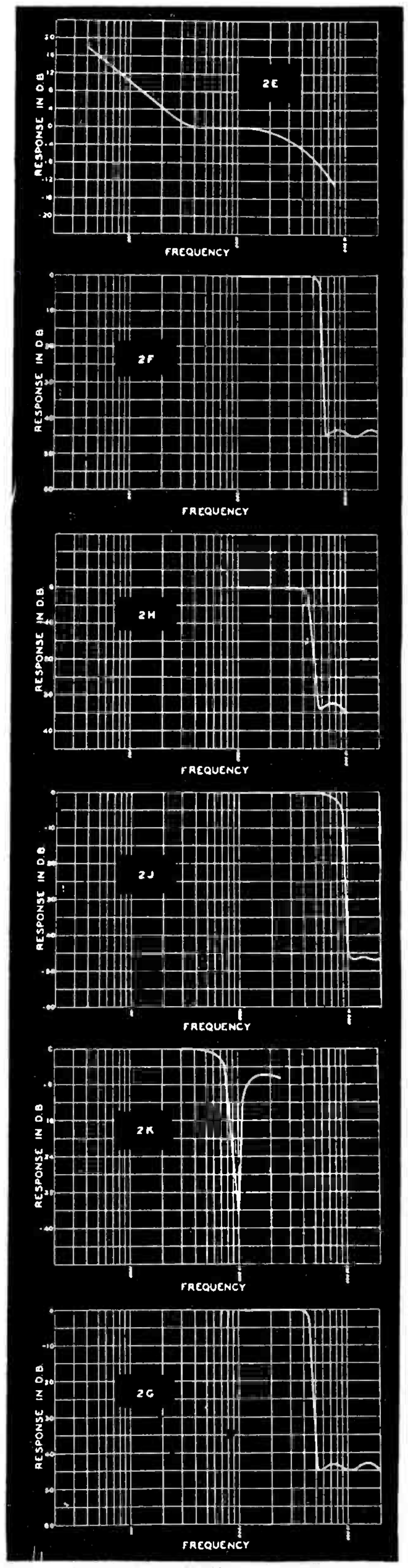


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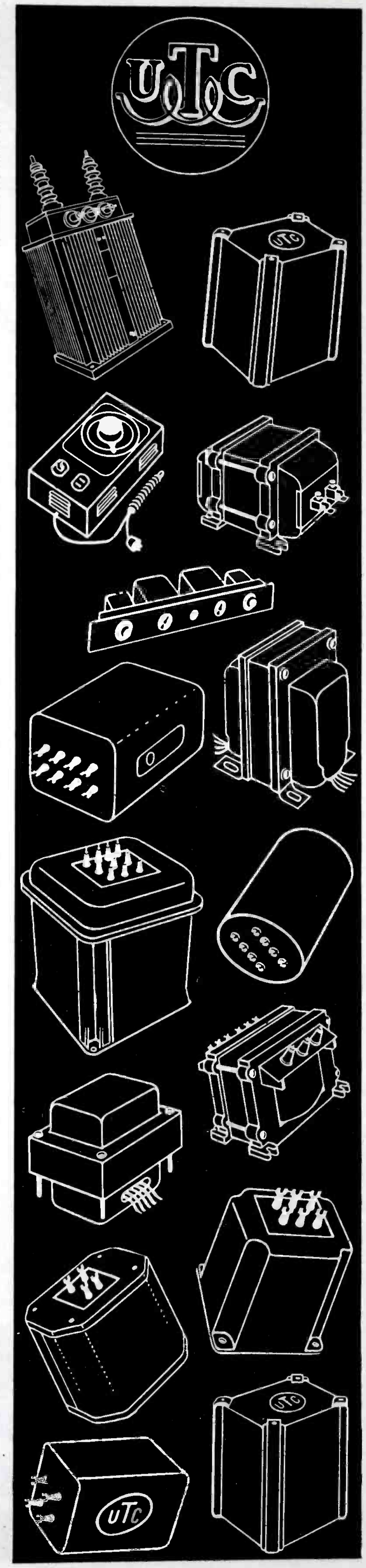

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March 7, 1940

Mr. J. A. McCullough
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San Bruno, California

Dear Mr. McCullough:

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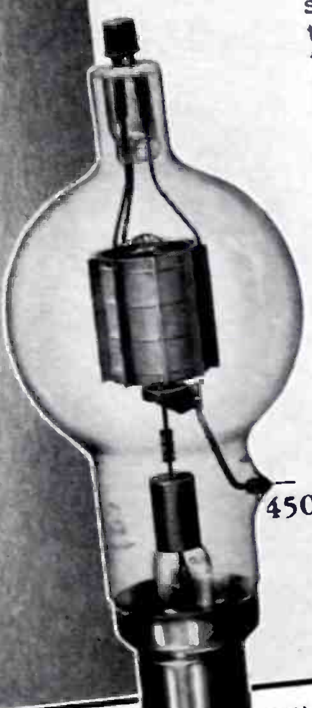

450T

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G. A. O'Reilly
Ground Station Radio Engineer

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# Tuned-Grid Tuned-Plate Oscillator

**By I. E. MOUROMTSEFF**
Special Products Engineering Dept.
Westinghouse Lamp Division
*(Bloomfield, N. J.)*

Among various types of coupling between the grid and the plate circuits in vacuum-tube oscillators, the internal feed-back through the plate-to-grid capacitance deserves particular attention. It acquires paramount importance in production of oscillations of very high frequencies. On the other hand, a closer analysis of all factors involved in the generation of this kind of oscillation can be of service in solution of the inverse problem; that is, how to reduce the tendency to parasitic oscillations resulting from incomplete neutralization in a triode, or from the residual capacitance in a screen-grid tube amplifier.

It can be shown[1] that the feed-back through the plate-to-grid capacitance, $C_{pg}$, is equivalent to the combined effect of a pure resistance, $R_x$, and of an imaginary capacitance, $C_x$, both connected on the input side parallel to the grid (Fig. 1). The numerical values of this *feed-back or reflected resistance and capacitance* can be calculated from the expressions:

$$R_x = -1/\omega A_2 C_{pg} \quad \ldots\ldots (1)$$

and

$$C_x = (1 + A_1)\, C_{pg} \quad \ldots\ldots (2)$$

where ω designates circular frequency of the oscillation, while $A_1$ and $A_2$ are the real and imaginary components of the r-f plate-to-grid voltage ratio, A:

$$A = -e_p/e_g \quad \ldots\ldots (3)$$

The same factor, A, can also be expressed through the tube and circuit parameters:

$$A = \mu Z_a/(Z_a + R_p) \quad \ldots\ldots (4)$$

Note: It may be noted that for sustained oscillation in any circuit the coupling, K, between the plate and grid circuits, determining the amount of the feed-back must be equal to the reciprocal value of the voltage ratio; that is, $K = 1/A$.

In eq. 4, $\mu$ is the amplification factor of the tube,
$R_p$ is the internal tube resistance,
$Z_a$ is the complex impedance of the load circuit:

$$Z_a = X_r + j\, X_i \quad \ldots\ldots (5)$$

with $X_r$ and $X_i$ designating the series resistance and series reactive components of $Z_a$. Here, $X_r$ is different from the parallel load resistance, $R_L$, which will enter into our discussion later; $R_L$ is always used in calculation of output power from tube characteristics, while $X_r$ is introduced merely as a convenient parameter in this discussion. From (4) and (5) it follows that the voltage ratio factor, A, is generally also complex and can be split into real and imaginary components:

$$A = A_1 + j\, A_2 \quad \ldots\ldots (6)$$

The most important fact about the outlined representation is that the feed-back resistance, $R_x$, in the grid circuit becomes *negative* for any assumed anode frequency, $\omega_a$, that makes the output circuit inductive, that is, for $\omega_a < \omega_r$, the latter being the resonance frequency of the load circuit. The presence of a negative resistance in a circuit is a sure sign that the circuit may generate oscillations. In our case, it will actually occur if the feed-back resistance, $R_x$, is numerically smaller than $R_g$, the positive parallel resistance inherent in the grid circuit. The frequency of the generated oscillations is then:

$$\omega_x = 1/\sqrt{L_g\,(C_g + C_x)} \quad \ldots\ldots (7)$$

that shows that frequency depends on both the real and the feed-back capacitances in the grid circuit.

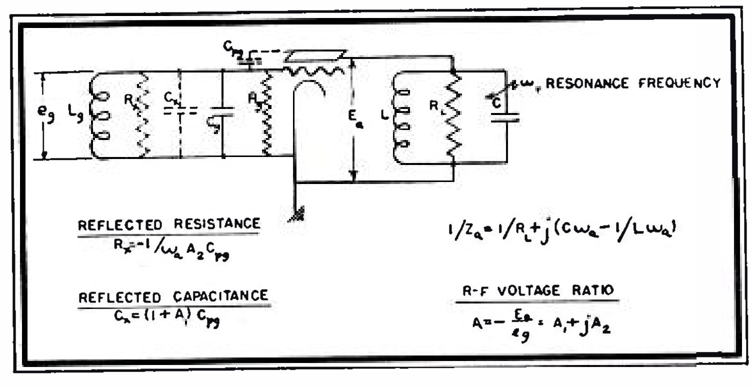
Obviously, all our equations are consistent with each other only for the condition when the calculated frequency of the grid circuit and the frequency of oscillation assumed to be existent in the anode circuit (and used for the computation of $R_x$ and $C_x$) are equal to each other, that is, if

$$\omega_x = \omega_a \quad \ldots\ldots (8)$$

This is the only physically possible frequency for a chosen set of the tube and circuit constants. Indeed, only in this case will the oscillation amplified by the tube react on the grid circuit in the manner necessary to sustain itself.

After these preliminary remarks, we shall analyze more closely the factors involved in the production of oscillations—whether useful or parasitic—caused by the feed-back through the plate-to-grid capacitance of the tube. We shall preferably use a graphical method as it gives

**Fig. 1. Diagram of plate reaction on grid circuit.**



a better insight into the role of the individual factors and their mutual relations.

Equations (3), (4) and (5) permit one to express $A_1$ and $A_2$, the real and the imaginary components of the voltage-ratio factor, as functions of the tube and circuit parameters:

$$A_1 = \mu \frac{R_p X_r + Z_a^2}{(R_p + X_r)^2 \times X_1^2} \text{ and}$$

$$A_2 = \mu \frac{R_p X_r}{(R_p + X_r)^2 \times X_1^2} \qquad \text{.............. (9)}$$

These equations imply that both $A_1$ and $A_2$ depend on the actual frequency in the anode circuit, which we do not yet know. However, for any *assumed* frequency, $\omega_a$, in the output circuit we can calculate $Z_a$, $X_r$, and $X_1$. This we shall do by means of an *inversion diagram* as described below.[1]

In the study of vacuum tubes, it is convenient to represent the load circuit as consisting of three parallel branches (Fig. 1)—one containing only the load resistance, $R_L$, another being purely inductive and the third capacitative. The admittances of such a circuit are connected by the algebraic relation:

$1/Z_a = 1/R_L + (C\omega - 1/L\omega)$ .......................... (10)

Vectorially it is represented in Fig. 2. If the resulting imaginary component is inductive, it is plotted upward; if capacitative, downward.

By way of example, in all subsequent diagrams and calculations the following numerical data are used:

| | |
|---|---|
| Resonance frequency, | $f_r = 10^6$ cycles/sec. |
| Amplification factor, | $\mu = 12$ |
| Plate-to-grid capacitance, | $C_{pg} = 14.5$ mmfd. |
| Inner tube resistance, | $R_p = 2000$ ohms |
| Parallel load resistance, | $R_L = 2500$ ohms |
| Parallel resistance in grid circuit, | $R_g = 8250$ ohms |

For evaluating the vectors $Z_a$, $X_r$ and $X_1$, entering into the previous formulae, we shall utilize Fig. 2 and prepare from it an inversion diagram in the following manner (Fig. 3). After replotting the previous diagram we produce the vector $1/R_L$ and plot its reciprocal value, $R_L$, to an arbitrary scale. Then, taking $R_L$ as a diameter, we describe a circle and extend all $1/Z_a$-vectors to their intersection with the circle. From the geometrical relations of the constructed diagram, one can derive that the new vectors represent the $Z_a$-values for assumed frequencies. Their horizontal and vertical projections are equal to $X_r$ and $X_1$, respectively. Just a glance at the diagram reveals that

$Z_{a\,max} = X_{r\,max} = R_L$
and ................................ (11)
$X_{1\,max} = R_L/2$

By supplementing the diagram with the horizontal vector, $R_p$, on the left of the origin, we also obtain a vector the square of which represents the denominator in the expressions for $A_1$ and $A_2$. Thus, both of these quantities can be computed, and the only question to be decided is the frequency scale on our diagrams. This can be derived from a simple relation between the "detuning" of the circuit, D, and the quantity, Q, of the circuit used. It is known that:

$D = \omega_a/\omega_r - \omega_r/\omega_a = \text{approx. } 2(\omega_r - \omega_a)/\omega_r$ ............ (12)
and
$Q = R_L/\omega_r L$ ...................................... (13)

A simple transformation of equation (10) permits one to conclude that for the 45°-phase angle of $1/Z_a$, the relation exists:

$D = 1/Q$ .......................................... (14)

Indeed, in this case

$1/R_L = C\omega_a - 1/L\omega_a$ ................................ (15)
and
$1/R_L = 1/L\omega_r\ (\omega_a/\omega_r - \omega_r/\omega_a)$ .................... (16)

Wherefrom one arrives at the relation (14). Thus, the frequency pertaining to the phase angle of 45° can be computed from the equations (12) and (14).

In a similar way, by equating D to 2/Q, 3/A....., and to 1/2Q, 1/3Q....., one is able to determine frequencies for various other vectors.

Thus, the inversion diagram permits the calculation of the voltage-ratio components, $A_1$ and $A_2$; also, for any assumed frequency (more or less departing from

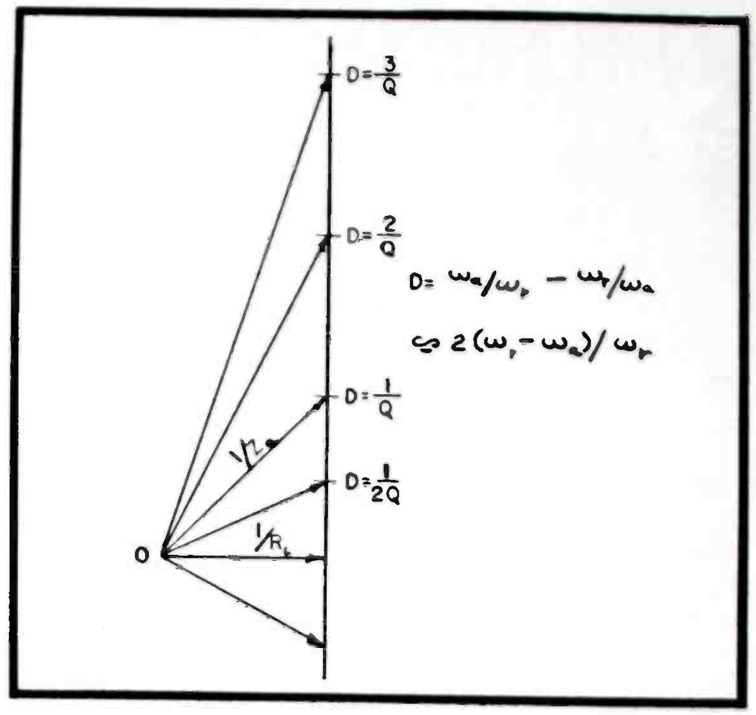
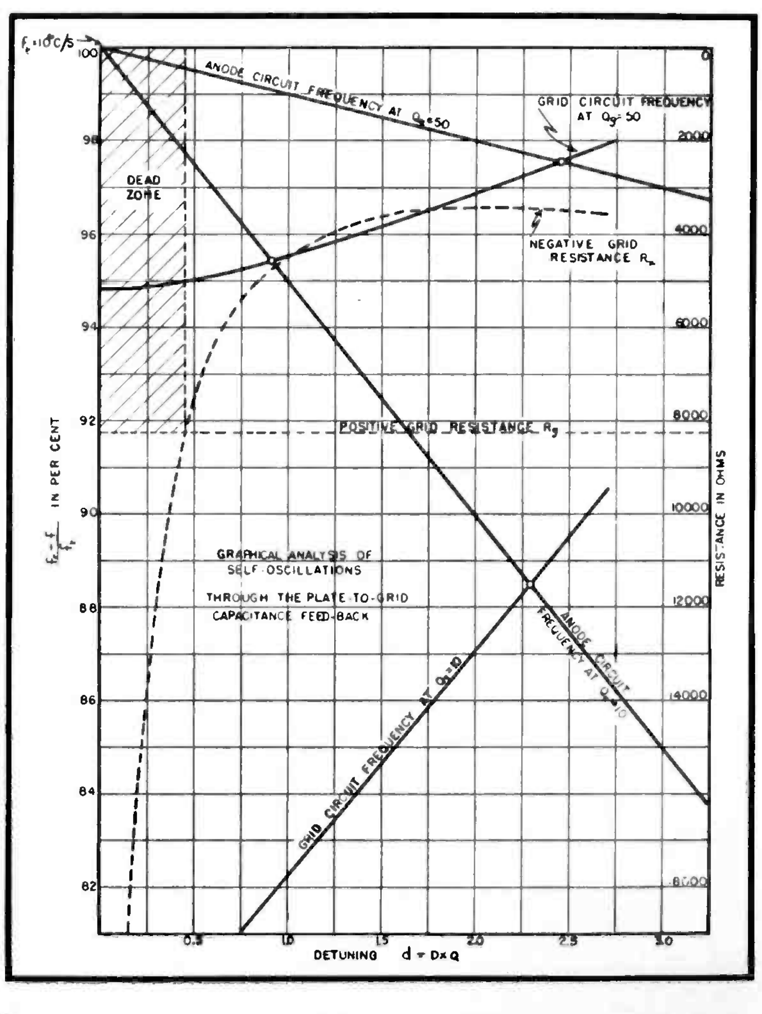



**Fig. 2. Vector diagram of load circuit.**

**Fig. 5. Other curves for the graphical analysis.**



the resonant frequency) the values of the equivalent load impedance, $Z_a$, its phase angle and also the phase angle of A. All this enables one to calculate the negative resistance, $R_x$, and the feed-back capacitance, $C_x$, in the grid circuit, and finally, the exact frequency of oscillation generated in the grid circuit.

The results of calculation of A, $A_1$, $A_2$, $C_x$ and $C_a$ are plotted in Fig. 4 and $R_x$ is shown in Fig. 5. All curves are referred to "detuning" of the anode circuit; however, instead of D, the product DQ is used as an independent variable, since this makes the results more comparable to each other. In this case the abscissa, 1.0, always corresponds to the 45°-detuning for any value of Q (see equation 14). One may designate $d = DQ$ as *relative detuning*. The "dead zone" in Fig. 5 corresponds to the condition when the feed-back resistance, $R_x$, is numerically greater than the positive parallel grid resistance, at which condition—according to the previous discussion—no oscillation is possible at all.

It has already been pointed out that only that oscillation will be sustained, the frequency of which, $\omega_x$, is equal to the assumed frequency in the output circuit, $\omega_a$, used in evaluation of $Z_a$, $X_r$, and $X_i$. In order to find the numerical values of the actual frequency of sustained oscillations, the anode frequency, $\omega_a$, and also the calculated frequency in the grid circuit, $\omega_x$, are plotted in Fig. 5. As ordinates of these curves the values $(f_r - f_x)/f_r$ are plotted in per cent instead of cycles per second. This allows for the use of the same anode frequency curves with any other assumed resonant frequency, different from the $10^6$ cycles/sec. of our numerical example. The intersection of the $\omega_x$ and $\omega_a$ curves indicates the looked-for frequency, the only feasible one under the chosen conditions. One may notice that Fig. 5 contains anode frequency curves corresponding to the values of Q, — 50 and 10. In accordance with a previous statement, the oscillations can actually be sustained only if the intersection of the anode and grid frequency curves falls in the region where the feed-back resistance, $R_x$, is numerically smaller than the positive grid resistance, $R_g$.

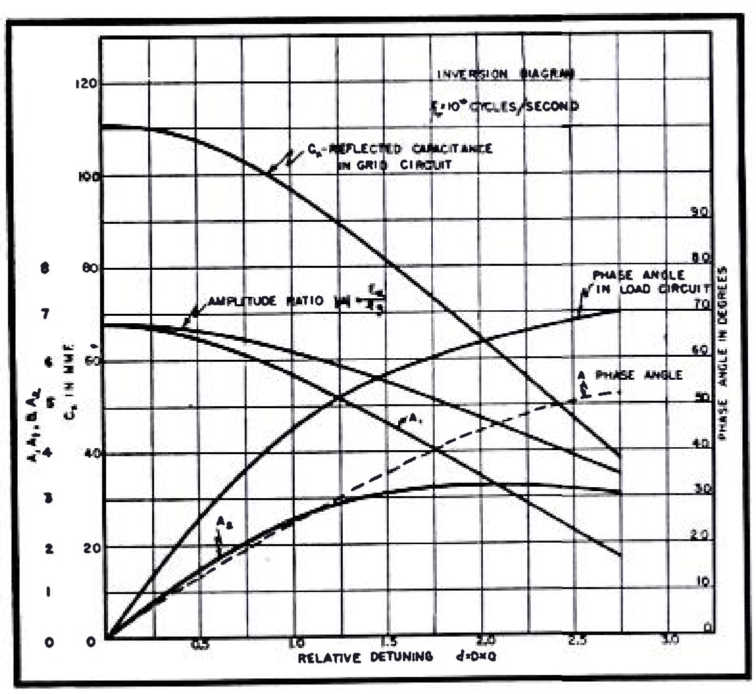



**Fig. 4. Curves of A, $A_1$, $A_2$, $C_x$ and $C_a$.**

The sequence of calculation is as follows: First, by means of the inversion diagram, the vectors, $Z_a$, $X_r$ and $X_i$ are determined for an assumed frequency $\omega_a$ or—what is the same—for an assumed detuning, d. These

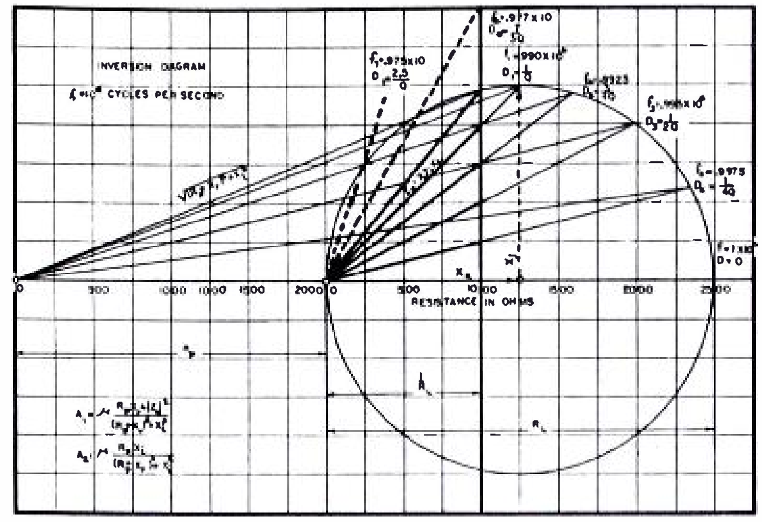



**Fig. 3. An inversion diagram prepared from Fig. 2.**

values are used in the calculation of $A_1$ and $A_2$ (equation 9). Then, $R_x$ and $C_x$ are computed from equations (1) and (2). Further, from the known $C_g$ and calculated $C_x$, one finds $\omega_x$, the would-be frequency of the grid circuit under the assumed conditions. One can use the relation:

$$\omega_x/\omega_r = \sqrt{C_g/(C_g + C_x)} \quad \ldots\ldots (17)$$

Repeating the same process for several assumed frequencies in the anode circuit, one can draw curves for both $\omega_r$ and $\omega_x$, and from their intersection determine the frequency of feasible oscillation.

In order to render the role of the grid circuit constants more conspicuous, it is convenient to apply a similar procedure to the grid circuit of different $Q_g$'s; then, one has to derive values of $C_g$ from the relation

$$1/C_g\omega_r = R_g/Q_g \quad \ldots\ldots (18)$$

#### Discussion of Results

The curves of Figs. 3, 4 and 5 permit a quick conclusion as to the role played by each factor involved in the production and sustaining of oscillations in tuned-plate tuned-grid oscillators. It is obvious that the factors favoring generation of oscillations, when these are desired, will also favor starting of annoying parasitic oscillations when the tube is used as an amplifier. Therefore, all measures which are to be recommended for production of oscillations in the first case must be reversed for suppression of parasitic oscillations in the case of an amplifier. Now, we shall more closely analyze the influence of various factors.

(1) The role of the grid resistance, $R_g$, is obvious. The higher this resistance, the narrower is the "dead zone" in Fig. 5. Hence, for production of oscillations, $R_g$ must be large, for suppression of parasitics—low. Normally, $R_g$ is determined by the grid excitation loss, both in the bias and in the tube. The loss is due to the electronic current from the filament to the grid. One must be reminded that in this analysis, we designate by $R_g$ *parallel* grid resistance. In application to the amplifier, one can arrive at a conclusion that more effective suppression of undesirable oscillations can be achieved by connecting an additional resistance across the grid circuit, as this would widen the "dead zone."

(2) The role of the Q's of the anode and grid circuits is also rather obvious. By varying the Q's, one can shift the point of intersection of the anode and the grid frequency curves. As is clear from Fig. 5, the smaller the $Q_a$ the steeper is the anode frequency curve; also, a

*(Continued on page 22)*

# FREQUENCY VS. PHASE MODULATION

**By HERBERT J. SCOTT**
Dept. of Electrical Engineering
University of California

IT has been stated on occasion that frequency modulation and phase modulation are essentially one and the same thing, that the distinction between them is purely academic, and that to differentiate between them is to quibble. As will be demonstrated subsequently there is in general, except for a very special case, a very real difference between these two types of modulation.

If a carrier is modulated with a *single frequency* the expression for a *phase* modulated wave becomes,

$$e = A \sin(\omega t + m \sin pt) \qquad (1)$$

where; A is the amplitude of the wave (a constant in this type of modulation), $\omega/2\pi$ the carrier frequency, $p/2\pi$ the modulating frequency, and the modulating index $m = k_1\phi$. The modulating index expresses the deviation in phase, $\phi$, with modulation and is related to the amplitude of the modulating signal through the factor $k_1$. In particular it will be noted that the modulation index is *independent* of the frequency of modulation.

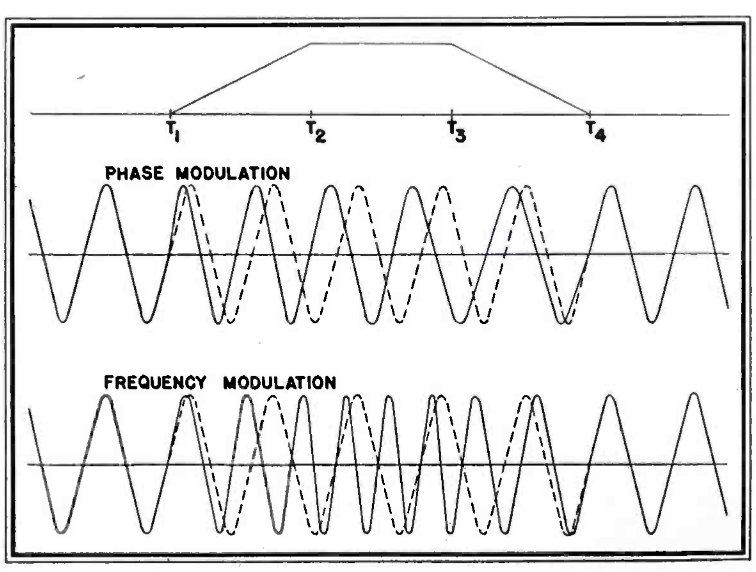
If the same carrier is now frequency modulated with the same single frequency the corresponding expression for the *frequency* modulated wave becomes,

$$e = A \sin(\omega t + M \sin pt) \qquad (2)$$

where; A, $\omega$, and p are defined as above. In this type of modulation, however, the modulating index $M = k_2\omega/p$. Note particularly that M is directly proportional to the deviation in carrier frequency $k_2\omega/2\pi$, is related to the amplitude of the modulating signal through the factor $k_2$, and is *inversely proportional to the frequency of modulation* $p/2\pi$.

Equations (1) and (2) are identical in form except for the modulating indices m and M. These equations may be more conveniently expressed as follows,

**Fig. 2. Phase modulation and frequency modulation resulting from a trapezoidal modulating wave.**




$$e = A \sin(\omega t + x \sin pt)$$
$$= A \sum_{n=-\infty}^{\infty} J_n(x) \sin(\omega + np)t \qquad (3)$$

where now x is to be replaced by m for phase modulation or by M for frequency modulation. The quantity $J_n(x)$ is a Bessel's function of the first kind, of order n, and argument x. Formal expansion of the right hand member of equation (3) yields,

$$e = A \left\{ \begin{array}{ll} & J_0(x) \sin \omega t \\ & \text{(carrier)} \\ \text{(lower sidebands)} & \text{(upper sidebands)} \\ -J_1(x) \sin(\omega - p)t & +J_1(x) \sin(\omega + p)t \\ +J_2(x) \sin(\omega - 2p)t & +J_2(x) \sin(\omega + 2p)t \\ -J_3(x) \sin(\omega - 3p)t & +J_3(x) \sin(\omega + 3p)t \\ \cdots\cdots & \cdots\cdots \\ (-1)^n J_n(x) \sin(\omega - np)t & +J_n(x) \sin(\omega + np)t \end{array} \right\} \qquad (4)$$

Examination of equation (4) shows a group of sidebands, symmetrically located with respect to the carrier and separated from the carrier by integral multiples of the modulating frequency, the absolute magnitude of any upper sideband being equal to that of its companion lower sideband. It is evident that if m is now made equal to M the resulting modulated waves themselves, and consequently their spectral components, will be identical for *both* frequency and phase modulation. Once produced it is impossible to determine, by any means, which system of modulation created them.

From some such superficial examination and the evidence it presents, may have arisen the notion that frequency modulation and phase modulation are alike. Such a point of view, however, is almost akin to assuming that because two intersecting lines have a point in common they must therefore be coincident.

A striking illustration of the difference between frequency modulation and phase modulation is obtained when the spectral distributions resulting from such modulations are viewed side by side as in Fig. 1. The modulating intensity is of such a value that the carrier deviation is nominally ± 60 kc for frequency modulation. This gives a value of $M = 4$ when the modulating frequency is 15 kc. For phase modulation the modulating intensity is chosen such as to make $m = 4$.

When the modulating frequency is 15 kc the frequency modulated wave and the phase modulated wave are identical corresponding to the previously discussed condition. As soon as this frequency of modulation is departed from, a marked difference between frequency modulation and phase modulation becomes apparent.

In phase modulation the modulating index corresponding to a given degree of modulation, having once been set at a value of 4, remains constant regardless of the frequency of modulation and consequently results in a *fixed* number of sidebands[1]. It will also be noted that the envelope of the carrier and the sideband magnitudes is of the same shape for all frequencies of modulation except that the frequency span progressively diminishes with a corresponding decrease in modulating frequency. Again, the magnitude of the carrier is the same for all frequencies of modulation, and the magnitude of any given order sideband likewise remains the same for all modulating frequencies.

In frequency modulation the modulation index depends upon the carrier deviation *and* upon the modulating frequency. Since for a given degree of modulation the carrier deviation is fixed, the modulation index becomes increasingly larger as the modulating frequency becomes progressively lower. The result is that while the spectral width remains approximately constant as the modulating frequency is decreased, the number of sideband frequencies present steadily increases. The envelope of the carrier and sideband magnitudes for one frequency of modulation bears no fixed relationship to the envelope for any other frequency of modulation. This is quite different from the observations made in connection with phase modulation.

The foregoing discussion is based upon a modulating wave of sinusoidal form whose frequency is changed in discrete steps, and has but one value at any one time. In order to transmit intelligence a simple sinusoidal wave form will not suffice. Such transmission requires a *complex* wave form of modulation and this makes the difference between frequency modulation and phase modulation even more pronounced.

To demonstrate this more convincingly a trapezoidal modulating wave is shown in Fig. 2. In this same figure are shown respectively the resulting phase and frequency modulated waves. The solid lines show the wave before, during, and after modulation. The dotted lines show the wave as it would have appeared during this period had no modulation been applied.

For phase modulation the *phase* increases linearly during the interval from $T_1$ to $T_2$ and is given by[2] $\phi = k_1 t$. The resulting wave during this interval is then,

$$e = A\sin(\omega t + k_1 t) \qquad (5)$$

This may be written

$$e = A\sin(\omega + k_1)t \qquad (6)$$

and in this form brings out an interesting point. As soon as modulation commences the frequency abruptly changes from $\omega/2\pi$ to a new value $(\omega + k_1)/2\pi$ and remains *constant* at this new value until $T_2$ is reached. At this point the phase has been advanced by an angle $k_1 T_2$ radians. From $T_2$ to $T_3$ the phase advance remains constant and the frequency in this interval again becomes $\omega/2\pi$, *the original frequency of the unmodulated carrier.* In the interval from $T_3$ to $T_4$ the phase is retarded (frequency decreased) until at $T_4$ the wave becomes once more that of the undisturbed carrier.

If the modulating magnitude or the time interval $T_1$

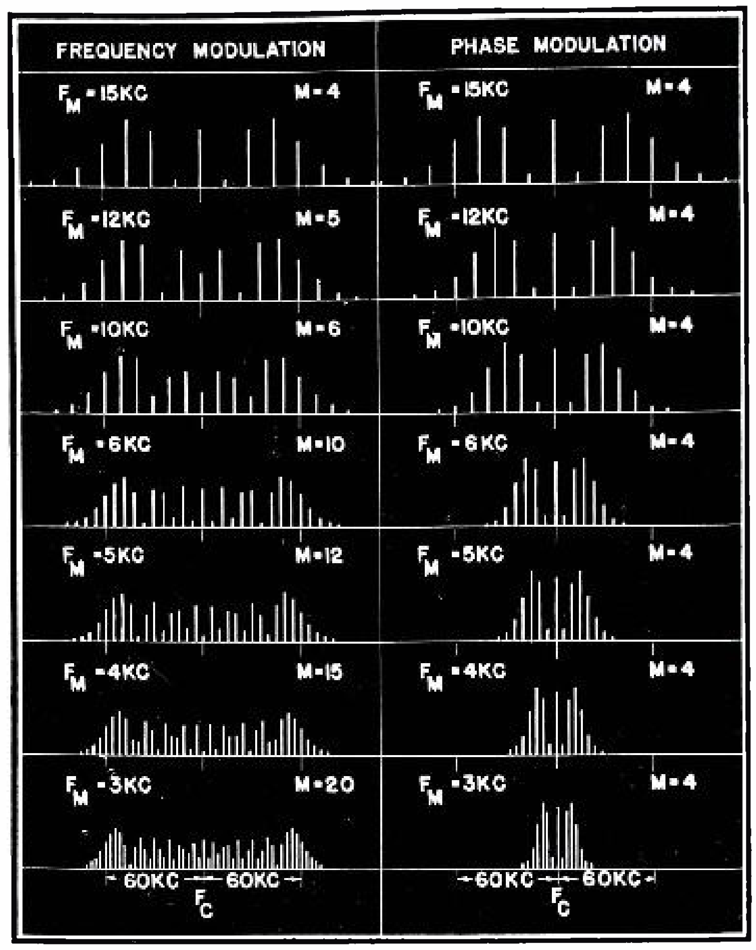



**Fig. 1. Spectral distributions resulting from frequency modulation and from phase modulation. Degree of modulation constant, frequency of modulation as indicated.**

to $T_2$ were properly adjusted, the phase advance at the end of this interval could be made equal to $2\pi$ radians (360 degrees). The modulated wave in the interval $T_2$ to $T_3$ would then be *coincident* with the unmodulated wave in the same interval. Were one to be confronted with the wave in this interval before modulation and again during modulation, *no physical measurement could discern the difference between the two.*

In frequency modulation the *frequency* increases linearly with time from $T_1$ to $T_2$ and during this interval the wave is given by the expression,

$$e = A\sin(\omega t + \tfrac{1}{2} k_2 t^2) \qquad (7)$$

which may be written

$$e = A\sin(\omega + \tfrac{1}{2} k_2 t)t \qquad (8)$$

(compare equations (7) and (8) with equations (5) and (6) respectively).

In contrast with the phase modulated wave, the frequency modulated wave continually *increases* in frequency in the interval $T_1$ to $T_2$ and the instantaneous frequency is $(\omega + \tfrac{1}{2} k_2 t)/2\pi$. From $T_2$ to $T_3$ its frequency remains constant at whatever value it has upon reaching $T_2$ and from $T_3$ to $T_4$ its frequency gradually decreases from this value until at $T_4$ the wave is once more coincident with the undisturbed carrier.

When the results of frequency modulation are thus compared with the results of phase modulation there is seen to be a very real distinction between them. It becomes evident at once that they are neither one and the same thing nor is one just a "special case" of the other.

[1]Theoretically there is an infinitude of sidebands but values of $J_n(4)$ for $n > 7$ are essentially zero.
[2]Assuming that $t = 0$ at $T_1$.

# Synchronized Frequency Modulation

WITH the expansion of frequency-modulation broadcasting it is interesting to follow the equipments as they are made available.[1] Among the new developments is Synchronized Frequency Modulation which makes its appearance in the Western Electric 503A-1, 1000-watt transmitter. It is the purpose of the following paragraphs to explain the operation of synchronized f-m.

According to Doherty,[2] foremost among the practical problems in frequency modulation is frequency stability . . . a term which takes on a new meaning, since it refers to average frequency. In conventional a-m systems, a crystal oscillator provides the desired stability. However, in a mode of transmission employing deliberate variation of frequency over a wide range, the direct use of the crystal as the source of the oscillations would necessarily give rise to a conflict between the factors which stabilize the frequency and those which are to produce the desired variation. Yet the mean frequency in f-m transmission is subject to the same regulation prevailing for the carrier frequency in amplitude modulation.

Now the mean frequency in a frequency modulated signal may be defined as the *total number of cycles* occurring in a second, whatever their distribution in time over this interval may be; so that a logical and direct procedure in maintaining the mean frequency at the assigned carrier value, would be to count continuously the number of cycles per second, comparing this with the number generated by a precise fixed-frequency standard, and adjusting the source of the oscillations to keep the two always exactly the same. This is in effect what is accomplished in synchronized frequency modulation. The procedure has a close parallel in electric power system practice, where cycle counting by means of synchronous motor clocks permits accurate control of the average frequency.

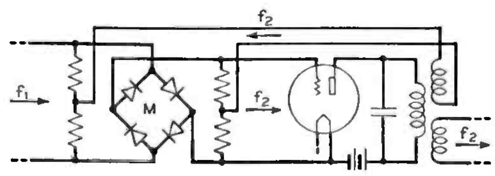
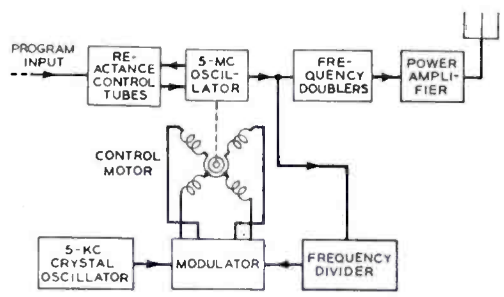



**Fig. 1. Schematic of a frequency divider.**

**Fig. 2. Block diagram of the f-m transmitter.**




It is not necessary, however, to count millions of cycles each second, for the frequency may be reduced to any desired degree through the new technique of *frequency division*, whereby a low frequency is obtained which is an exact submultiple derived directly from the original frequency and having its variations reduced in proportion. The frequency divider, a tool of considerable promise in the communication field, consists basically of a modulator (M. Fig. 1) and a vacuum tube amplifier. The frequency $f_2$ appearing in the output of the modulator is the difference between the frequency fed back from the output, which is $f_2$ itself, and the frequency $f_1$ applied to the device; that is, $f_2 = f_1 - f_2$. This requires that $f_2 = f_1/2$, so that we have an exact halving of the frequency; and the output wave, although produced by a regenerative action, is under complete control of the input by virtue of the modulation process through which it originates.

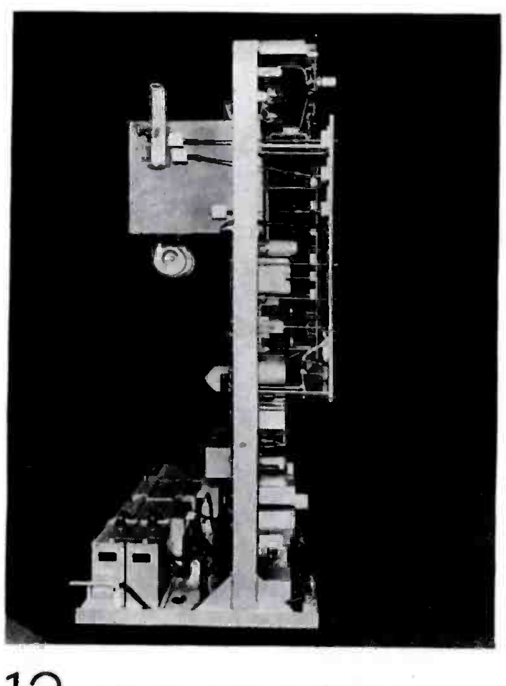
Using a modulator of the copper-oxide type, which recent refinements have rendered suitable for use at frequencies of several megacycles, the frequency divider becomes a compact and simple device. By cascading a series of such dividers, we obtain for synchronizing purposes a frequency as low as desired, in exact submultiple relationship to the carrier frequency. In synchronized frequency modulation the dividing process ends up with a frequency of about 5000 cycles, or 1/8000 of the carrier frequency.

Referring now to the block diagram of the system, Fig. 2, the role of the frequency divider becomes apparent.

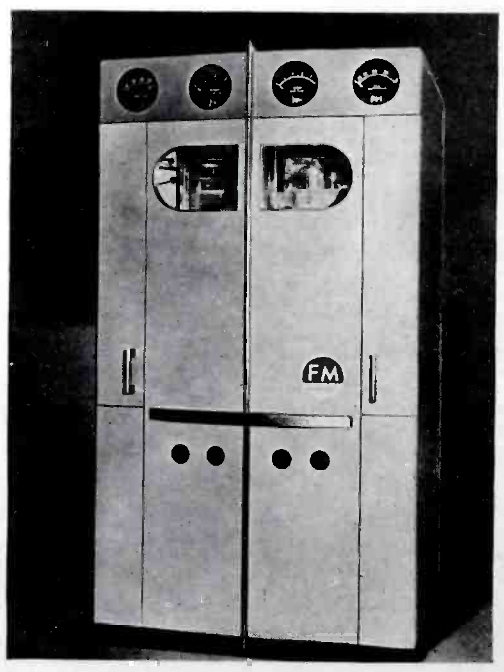
[1] "New U-H-F Transmitters", p. 10, December, 1939, COMMUNICATIONS. "Notes on F-M Transmitters", by Frank A. Gunther, p. 11, April, 1940, COMMUNICATIONS.
"Frequency Modulation Receiver Design", by Richard F. Shea, p. 17, June, 1940, COMMUNICATIONS.

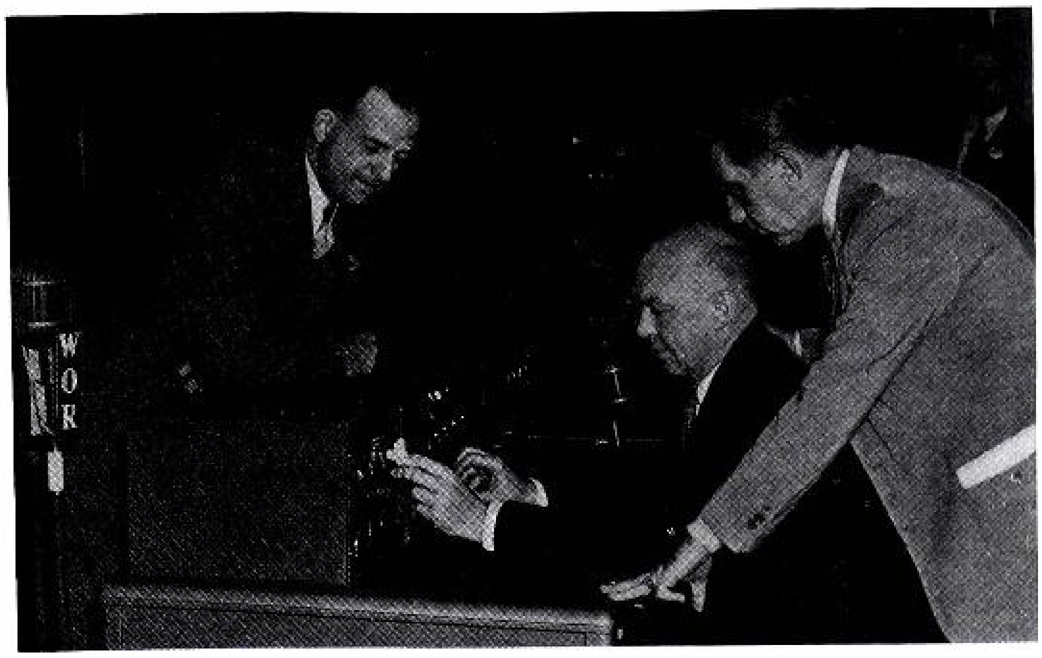
[2] Data from article "Synchronized FM", by W. H. Doherty, p. 3, August, 1940, *Pick-Ups*.

**Side view of the transmitter with cabinet removed.**



**Front view of the synchronized f-m transmitter.**






**Left to right: J. R. Poppele, Edwin H. Armstrong and Alfred J. McCosker at the dedication of W2XOR, WOR's new f-m station.**

The divider is energized from the output of a frequency-modulated oscillator operating at about five megacycles, and its function as a part of the synchronizing system is to insure the constancy of the mean frequency of this oscillator, and hence of the final output frequency (42 to 50 megacycles) to be obtained by doublers following the oscillator.

There has been in use for some years a method of synchronizing two frequencies wherein the frequencies are combined in a modulator to produce a rotating magnetic field whose speed and direction of rotation correspond to the amount and sense of the frequency difference. As a small armature, geared to the tuning condenser of one oscillating source, brings the frequency back toward synchronism, the speed of rotation of the field decreases and the armature slows down, coming to rest when exact synchronism is attained.

At first thought, one would not expect such a device to be applicable in an ultra-high-frequency system because the departures in frequency are so great as to be beyond the capacity of a mechanical system to follow; but when the frequency is reduced by the dividing process to the order of 5000 cycles, or 1/8000th of the output frequency, we find that variations of hundreds of kilocycles in the output frequency are represented by variations of only tens of cycles, so that with a low-frequency crystal oscillator as the comparison standard we obtain a rotating magnetic field readily followed by the armature. So effective and immediate is the control that if the output frequency through some cause were to depart suddenly by as much as *four hundred kilocycles* from its assigned value, it would be returned to exact synchronism in two to three seconds; while gradual changes in frequency of as much as several megacycles will also be corrected because the change is followed continuously.

It is well known in frequency modulation theory that the phase deviations are directly proportional to the frequency swing and inversely proportional to the audio rate at which the swing is produced. The frequency swings employed in wide band frequency modulation are so great as to entail phase deviations of thousands of degrees; that is, the frequency modulated wave is alternately advanced and retarded by many complete cycles with respect to an unmodulated comparison wave. When a high order of frequency division is introduced, however, the frequency swing becomes small while the audio rate is unchanged, so that the phase departures due to modulation are then only a few degrees.

The magnetic field in the control device therefore oscillates only slightly at audio frequencies about its mean position, and the oscillation is not followed by the motor because of its inertia; the slightest change in *mean* frequency, however, produces a continuous rotation of the field and is corrected at serves two important purposes: to reduce the whole phenomenon to a time scale suited to electromechanical operations, and to obscure the effects of modulation so that only changes in the mean frequency, or total number of cycles per second, can influence the frequency con- once. The frequency divider, then, trol mechanism.

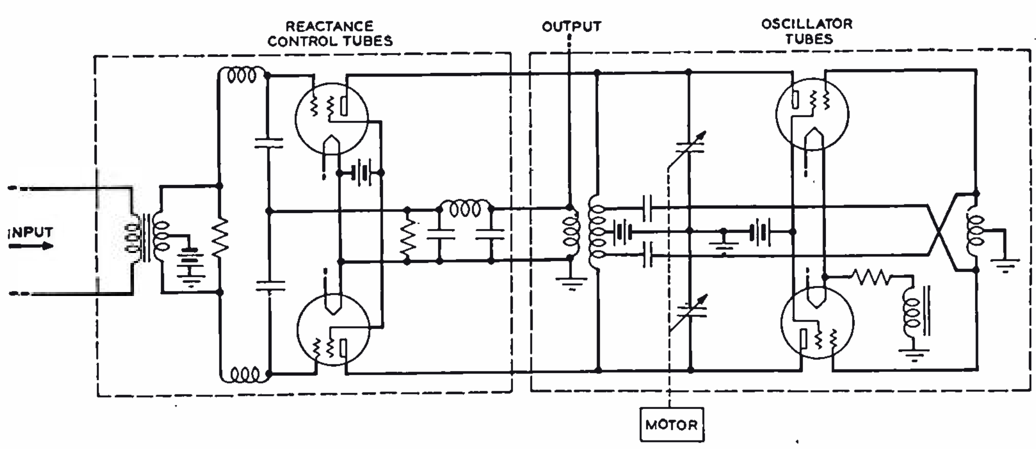



**Fig. 3. Schematic of the frequency modulation circuit.**

Not long ago 50 kilocycles was regarded as an extremely low frequency for quartz crystals. The appearance of a 5-kilocycle crystal oscillator in the block diagram of Fig. 2 is a reminder that advances in the frequency range of radio equipment are not being confined to the high-frequency end of the spectrum. This is a low temperature co-efficient crystal oscillator giving the same per cent stability as obtained in the best broadcast crystals. The stability is well under one part in a million per degree Centigrade, making temperature control unnecessary.

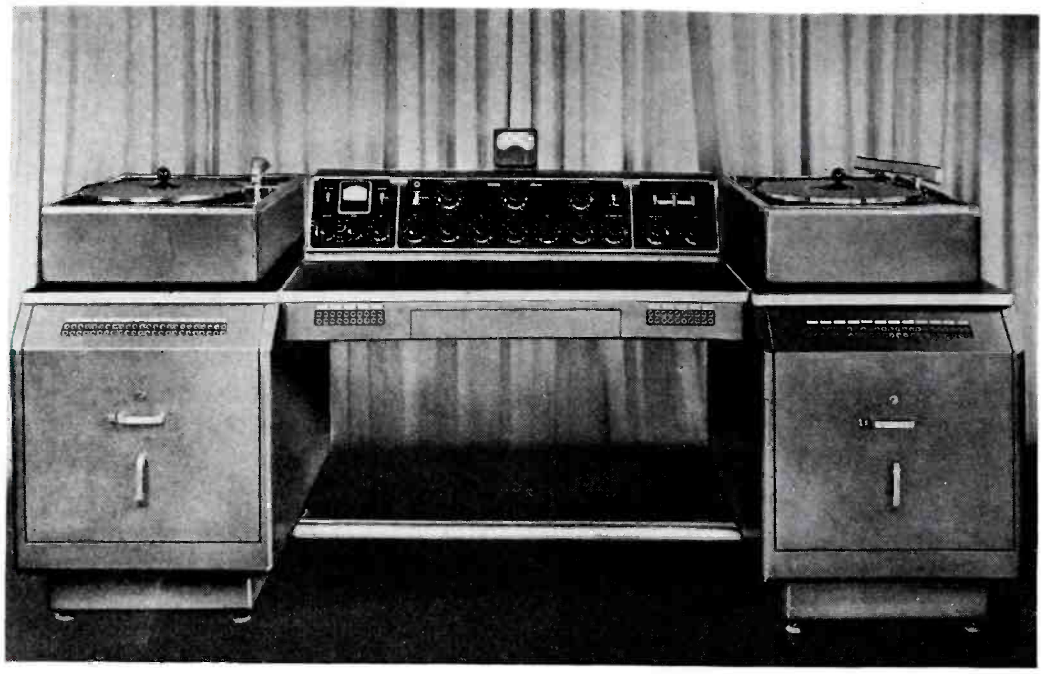
The system of frequency control described above is unique in a number of characteristics. For one thing, the stability is identically that of a single crystal oscillator, unaffected by any beating process with other oscillators, or by changes in gain or frequency characteristics of associated circuits. There are no temperature-controlled networks for converting frequency changes into amplitude changes, opening the door to errors due to gain fluctuations; everything in the control system is kept in terms of frequency. In the second place, the actual control exercised on the oscillator to maintain its mean frequency is mechanical, involving a variable condenser; and being mechanical, when the oscillator is brought to the correct frequency it is *left* there, without the necessity of any sustaining voltage.

Mechanical control, moreover, completely relieves the frequency modulating elements of any connection with the stabilization of the mean frequency, so that these elements may always be operated at the optimum point for linear modulation. Finally, the entire synchronizing system, including the crystal oscillator, is completely external to the program-carrying part of the transmitter.

#### Circuit Design

Frequency modulation holds such promise as a vehicle for high-quality noise-free broadcast service that it obviously justifies the production of a transmitting circuit design of extremely low distortion and background noise level. By modulating at a carrier frequency of five or six megacycles, where the phase deviations are large, the difficulty encountered at low initial frequencies from phase modulation due to power supply hum and microphonics is removed. In synchronized frequency modulation, moreover, the complete separation of the two functions of modulation and frequency stabilization permits the use of push-pull reactance control tubes for modulating the oscillator, so that ripples in bias or plate voltage supplies do not modulate the frequency. The balanced circuit (Fig. 3) employed for these tubes and for the oscillator, together with other refinements in design, permit a frequency excursion of hundreds of kilocycles on either side with very low distortion.

Following the modulated oscillator in the 503A-1 equipment are four pentode stages, three of them being doublers, and all extremely simple in design. At the final output frequency two triode stages increase the power to 1000 watts for transmission to the antenna. These triode stages use the 356A and 357A, stemless and baseless ultra-high-frequency tubes in the molded hard glass type envelope. The 357A is rated at 350 watts plate dissipation, with full voltage rating up to 100 megacycles. Two of these in the 40-50 megacycle range deliver 1000 watts for f-m with great ease.



**Showing the new desk type speech-input equipment adaptable for use with either frequency modulation or amplitude modulation stations.**

#### Speech Input

Special speech-input equipment[3] designed to accompany the f-m transmitter is shown in accompanying illustration—a block diagram of the equipment is given in Fig. 4. This apparatus was designed to meet the following requirements:

(1) Wide frequency range—from 30 to well beyond 15,000 cycles.

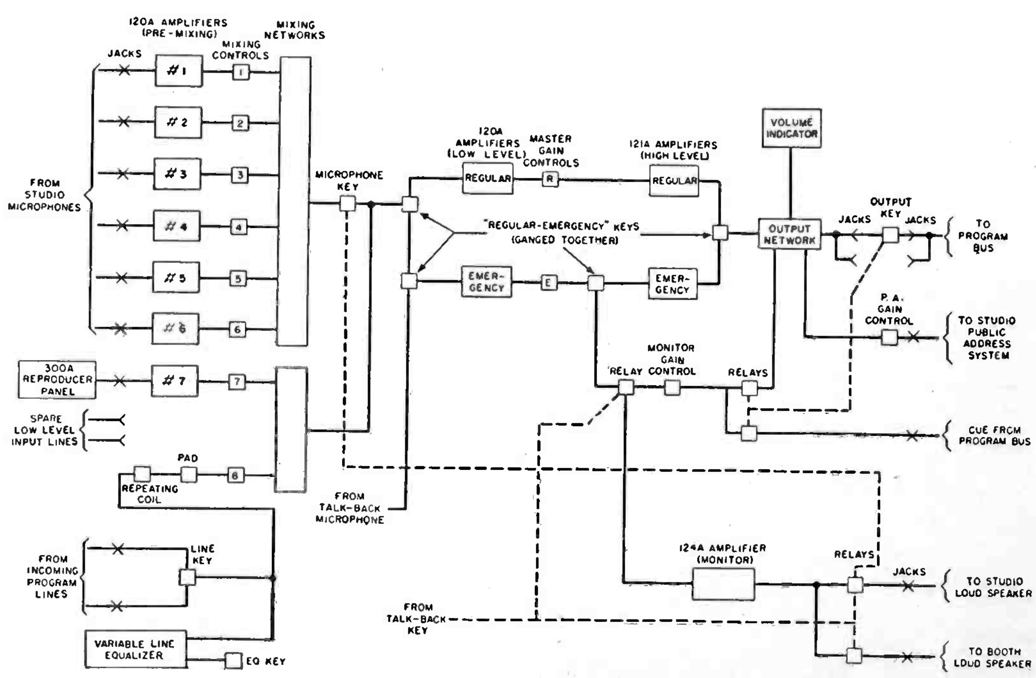



**Fig. 4. Block diagram of the new speech-input equipment.**

(2) Adequate net gain—ample margin over normal requirements to promote ease and flexibility of operation.

(3) Improved signal-to-noise ratio—a close approach to theoretical limits throughout the normal range of control adjustments, and better than 60 db at normal gain.

(4) Adequate overload margins—ability to deliver normal program level over a wide range of control adjustments and with uniformly low harmonic distortion.

(5) Simplicity of control—maximum compatible with full operating flexibility of a complex system.

(6) Maximum reliability—ability to deliver program even during partial equipment failure.

(7) Advanced mechanical design—compactness without sacrifice of accessibility for servicing; clean-cut, modern appearance and flexibility to permit assembly of equipment in various ways to meet various operating requirements.

(8) Use of standard components—the desk assembly is composed of standard apparatus available from the Western Electric Company.

(9) New basic structure—harmonizing with modern steel studio furniture design, with the probability that structural units of this type will be standardized.

A table unit with pedestal and turntable at each end is shown in an accompanying illustration. Such an assembly together with a separate power-supply cabinet will take care of elaborate studio installations.

[3]See article "WOR Installs Custom-built High-Quality Speech-Input System Designed for FM", by J. E. Tarr, p. 10, August, 1940, *Pick-Ups.*

# Book Reviews

*AN INTRODUCTION TO VECTOR ANALYSIS, by B. Hague, published by Chemical Publishing Co. of N. Y., 148 Lafayette Street, New York City,* 1939, 118 *pages, price* $1.50.

Year by year the communications engineer finds himself increasingly confronted with the necessity for at least a working knowledge of vector analysis. More and more, the literature of communication engineering implicitly assumes a knowledge on the part of the reader of this most important subject. Nevertheless, the practicing engineer can ill afford the time necessary to study three-hundred to five-hundred pages which are mainly devoted to the geometrical and mechanical applications of vector analysis in order to obtain some comprehension of its significance with regard to electricity and communication engineering. The present book, therefore, will enable the engineer to acquire a knowledge of vector analysis in the minimum of time, for, except when he is describing the underlying mathematical principles, the author restricts himself almost exclusively to their application to electrical phenomena.

The author's name is, of course, familiar to all electrical engineers as the author of *Alternating Current Bridge Methods,* which is a standard on this subject. As painstaking and authoritative a work as his book on a-c bridges is, it must nevertheless be confessed that it is written in a rather dry and pedantic style. The reviewer calls this to attention in order to contrast it with the beautifully lucid and interesting style which Professor Hague offers in *An Introduction to Vector Analysis.* For, not only does the author write with unusual clarity but he is also very careful to point out possible sources of confusion to the reader. Whereas most authors on vestor analysis take for granted that their readers will have no difficulty with the non-cummutative nature of vector products, Professor Hague carefully calls this, as well as other sources of confusion, to the attention of the reader.

For either the engineer who is unfamiliar with vector analysis or the engineer who is desirous of reviewing the subject this book is highly recommended.

R. L.

*APPLIED ACOUSTICS, second edition, by H. F. Olsön and F. Massa, published by P. Blakiston's Son and Co., Inc.,* 1012 *Walnut Street, Philadelphia, Pa.,* 1939, 494 *pages, price* $5.50.

The second edition of this text has been considerably enlarged and revised, with the result that the communication and acoustic engineer will find its usefulness greatly enhanced. Although the authors assume a familiarity of differential equations and vector analysis on the part of the reader, sufficient descriptive matter has been included so that the engineer who has grown mathematically rusty will also find this book of great utility in his work.

The treatment is mainly theoretical in character but, as its title implies, this book is mainly concerned with the practical aspects of electro-acoustic equipment. The major emphasis of the book is placed upon an analysis of microphones, telephone receivers, and loudspeakers, although other topics, such as acoustical measurements, architectural acoustics, measurement of noise, and physiological acoustics are also dealt with. It is unfortunate, however, that such sketchy treatment has been given to the chapter on Electrical Apparatus For the Acoustical Laboratory, and it is to be hoped that future editions will find this section considerably enlarged. Only two minor errors were noted: (1) footnote 17 on page 136 should read Vol. X, No. 3, and (2) in Fig. 5.38 on page 139 the numeral 5 is missing from the large dot at the extreme left of the figure.

This book is recommended to all communication engineers.

R. L.

*TABLES OF THE EXPONENTIAL FUNCTION* $e^x$, *by the Federal Works Agency, Works Projects Administration, conducted under the sponsorship of the National Bureau of Standards,* 1939, 535 *pages, price* $2.00. *(Remittances should be made payable to the "National Bureau of Standards.")*

The title of this book is self-explanatory, the book consisting of a series of tables for this function both for positive and negative values of x. Unlike similar tables of this function which appear in practically all engineering handbooks and which have almost no utility, the present tables have great practical value.

In all there are eight different sets of tables which differ in practical utility depending upon the particular problem in hand. The number of significant figures is unusual, and varies from twelve to nineteen.

Although it is already noted in the heading, the price is once again called to your attention, for at $2.00 this book becomes a necessary addition to every engineer's library.

This book is unqualifiedly recommended.

R. L.

*SUPERSONICS: THE SCIENCE OF INAUDIBLE SOUNDS, by R. W. Wood, published by Brown University, Providence, R. I.,* 1939, 158 *pages, price* $2.00.

Although the existence of inaudible "sound" waves has been known for a long time it was only since the first World War that any intensive investigation was made regarding the properties of supersonic waves. In recent years research work on supersonic waves has accelerated to such an extent that not only have many of their physical characteristics been discovered but, in addition, these waves have found industrial application.

Despite the fact that a considerable number of papers pertaining to supersonics are appearing in various journals, books on this subject are indeed rare. Accordingly, this book makes a welcome addition to the engineer's library, particularly in view of the fact that its author was one of the pioneers in the field.

Professor Wood's excellent summary of supersonics is divided into two main sections, the first concerning itself with the historical development of supersonics, while the latter deals with the physical and biological effects of these waves. The properties of supersonic waves are so unusual that it is greatly to be hoped that this book becomes as widely read as it well deserves.

R. L.

*HIGH FREQUENCY ALTERNATING CURRENTS, second edition, by K. McIlwain and J. G. Brainerd, published by John Wiley and Sons, Inc.,* 440 *Fourth Avenue, New York City,* 1939, 530 *pages, price* $6.00.

The fact that all available copies of the first edition of this book have become exhausted appears to have occasioned the authors some surprise. Just why this should be so is a little difficult to understand, for the fame of this excellent text is so far-reaching that in the reviewer's opinion not only will this new second edition become likewise exhausted, but all future editions will meet the same well-deserved fate. And this, not despite, but just because of its rigorous mathematical treatment.

Although some sections of the book presuppose a knowledge of differential equations and vector analysis on the part of the reader, the majority of the work can be read without familiarity of these topics. A grasp of the symbolic treatment of alternating current theory is, however, prerequisite.

While the authors verbally qualify the validity of equation 18e on page 35 and point out that it is true only for a very special case (and incidentally one rarely encountered in practice), it is confusing as it stands. The full equation might just as easily have been given, for the additional terms would certainly have caused no dismay in the reader of a book as full of equations as this one. On page 107, the lowest ordinate numeral of Fig. 55 should be 1.0 and not 0 as given. The reference on page 132, four lines from the bottom, to Fig. 133, is obviously wrong.

The authors' analysis of the triode is unusually good. On the other hand, they dispose of pentodes in less than half a page. Nor do they return to them when they discuss radio-frequency amplification, for their analysis is based on r-f triode amplifiers which are completely in the discard today. The discussion of Electromechanical Systems should prove of considerable interest to the communication engineer.

As is customary in books written by members of the teaching staff of a university, no answers to the problems are given, thereby considerably lessening the value of the book for the practicing engineer.

R. L.

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## BOOK CATALOG

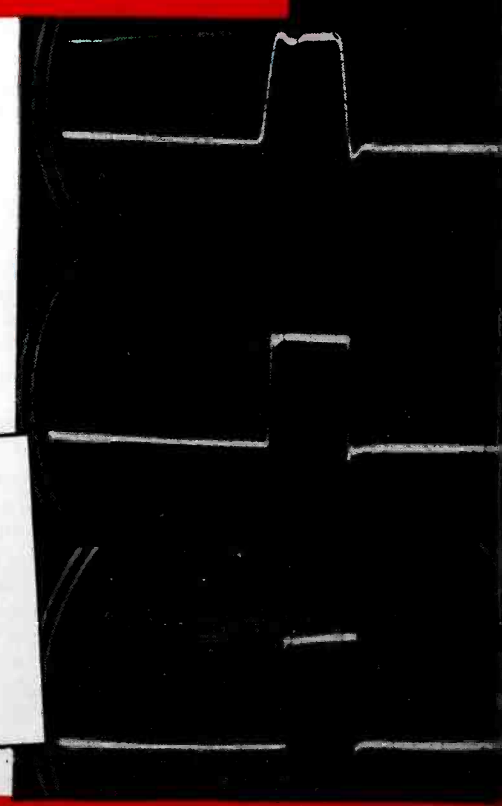
The new 1940 catalog of technical and scientific books of the Chemical Publishing Co., Inc., 148 Lafayette St., New York City, N. Y., contains many new titles. Books in nearly all technical and scientific fields have been added to the previous works.


# TOP PERFORMANCE BY EVERY TEST

## AND EVERY G-E FM TRANSMITTER PASSES 100%

QUICKLY, accurately, G-E engineers determine by thorough square-wave measurements the over-all performance characteristics of *every* General Electric FM transmitter. Carefully they check frequency stability in the G-E "torture chamber" at temperatures from freezing to 122 F.

Noise level, cross modulation, linearity—from every angle General Electric proves the capabilities of each unit before it goes onto the job. That is your assurance of dependable, high performance at low cost.



These typical square-wave measurements show an a-f characteristic that is flat within ±0.5 db from 15 to 16,000 cycles.

# Some Notes on Diode Detection

By ALBERT PREISMAN

*RCA Institutes, Inc.*

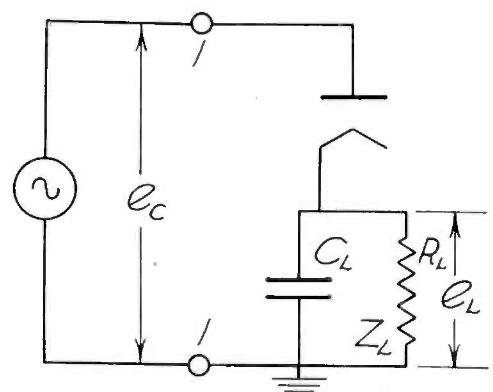
### 1. Diodes-Rectification Curves

THE diode detector is in almost universal use today as a demodulator, but its circuit design is far removed from that of the early Fleming valve. Its present-day popularity is due to its ability to handle reasonably large signals without overloading, its characteristic of linear detection of large signals, and ability—if required—to furnish the variable bias necessary for a-v-c with simplicity.

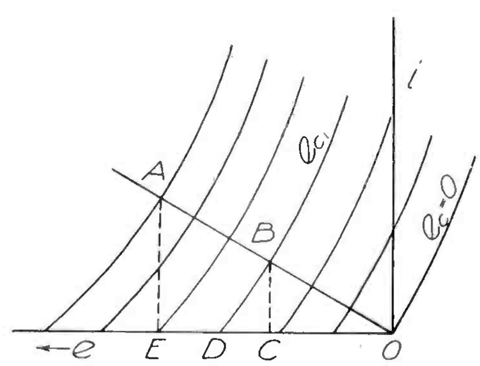
A linear detector is one whose output is a faithful copy of the envelope of the modulated high-frequency wave impressed upon it, i.e. the former is directly proportional to the latter. Such detection is facilitated by the use of an ideal diode. An ideal diode is one whose impedance to current flow in one direction is zero, and whose admittance to current flow in the opposite direction is zero. Actual diodes practically meet the latter condition, but not the former, since they present a finite, usually non-linear resistance in the conductive direction.

An elementary diode circuit is shown in Fig. 1. The high-frequency source impedance is assumed zero. The condenser $C_L$ is assumed to have negligible reactance at the high carrier frequency, $f_c$, and to have negligible susceptance at the low modulation frequency $f_m$. We shall first, however, assume that $e_c$ is unmodulated, that is of constant amplitude. The rectified output voltage, $e_L$, will therefore be d-c. Its magnitude, as well as that of the current flow through the diode, could be found by the methods described in another paper[1], but these become very involved when $e_a$ is assumed to be modulated, and various impedances present in the actual circuit are taken into account. Accordingly, a simpler, although more approximate method is to be sought.

For steady amplitude of $e_c$, $e_L$ can be found experimentally as well as graphically. Thus, first—by the Compensation Theorem—the impedance $Z_L$ can be replaced by a zero impedance d-c source, e, so poled as to oppose current flow through the diode. If a constant value of $e_c$ is impressed as well, and its peak amplitude exceeds e, rectified current will flow through the diode, and its d-c component, i, can be measured. If e is varied from the peak value of $e_c$ to zero, i will vary from zero to some maximum upper limit and give rise to an e-i curve similar to an ordinary $i_b$-$e_b$ curve for a triode. If $e_c$ is now adjusted to another fixed value, another curve will be obtained, and so on until a whole family is developed. This is shown in Fig. 2. The curves are approximately equidistant in spacing, and are known as *rectification curves* for the diode. It will be noted that they proceed from right to left as $e_c$ is increased, and also that these curves can be obtained at a conveniently low frequency for $e_c$.




**Above: Fig. 1—Below: Fig. 2.**




If e is now replaced by $Z_L$, i and $e_L$ (in place of e) must satisfy the relationship that $iR_L = e_L$. For any value of $e_c$, i can be found from the graphical solution for $R_L$ and the diode in series. Thus, the load line for $R_L$ is laid off as OA, and its intersection with the diode family at B gives the output current BC for a peak a-c (carrier) voltage, $e_{c1}$. The diode curves can represent by their slope the non-linear internal resistance of the diode, $r_d$. Then DO can represent an equivalent d-c voltage generated in the circuit, of value equal to the maximum value of $e_{c1}$. The portion DC is lost as voltage drop across the diode, and the remainder CO is available across $Z_L$. If $Z_L >> r_d$, then it can be seen that CO is practically equal to the maximum value of $e_{c1}$, and DC approaches zero. This is usually the case in actual practice. The diode thus approaches the ideal diode in performance.

Now suppose $e_c$ represents a modulated wave, of the form

$$e_c = E_c\,(1 + m \sin \omega_m t) \sin \omega_c t \quad \ldots\ldots\ldots\ldots (1)$$

where $\omega_m = 2\pi f_m$, the low, modulation angular frequency, and $\omega_c = 2\pi f_c$, the higher, carrier angular frequency. It is evident from Eq. (1) that the amplitude of $e_c$ may be regarded as made up of a constant value, $E_c$, and a variable quantity, of peak magnitude $mE_c$, and varying at a frequency $f_m$. Suppose in Fig. 2 that $e_{c1} = E_c$. Then BC and CO represent the constant components of the output current and voltage respectively. The variation in amplitude due to $mE_c$ produces a path of operation along AO. Assume, specifically, that $m = 1$ (100% modulation) and that this carries the operation from B to A, and from B to O. The output current then varies from BC to AE and down to 0 (zero), or it has an a-c component as well as the d-c component BC which alone is produced when there is no modulation ($m = 0$). Similarly, the output voltage has a variable or a-c component of peak-to-peak value EO, as well as the d-c component CO. It will be noted that the output voltage and current can follow the input carrier up to 100% modulation. If the diode family consists of straight lines of equidistant spacing, then from the geometry of the figure it is evident that the output voltage will be a constant fraction of the envelope of the input, and there will be no distortion, i.e. linear detection. In particular, if $r_d = 0$, the diode curves will be vertical, and $e_L$ will be equal to the

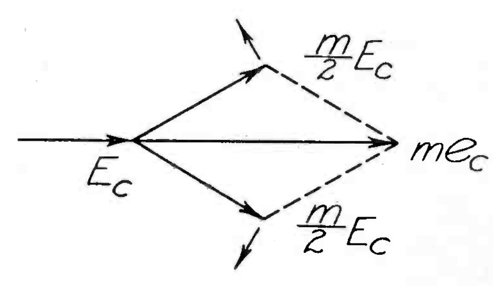



**Fig. 5.**

[1] "Graphics of Non-Linear Circuits"—Part I, Section IV—A. Preisman, *RCA Review*, July, 1937.

amplitude of the envelope, that is, the above-mentioned fraction will be unity. This ratio or fraction is known as the detection efficiency. If the diode family is curved, and $R_L$ is low relative to $r_d$, it is evident that the detection efficiency will vary over the path of operation, and hence over the modulation cycle, and also will be less than unity.

#### 2. Input Impedance

The output load, $Z_L$, or more specifically, $R_L$ (assuming $C_L$ has negligible susceptance at the modulating frequency) presents a certain equivalent resistance to the source of $e_c$. The value of this resistance, call it $R_c$, involves $r_d$ as well, and in practice is influenced by the other impedances in the circuit, such as that of $C_L$ (if appreciable), and of the source. However, a fair approximation to $R_c$ may be had as follows.

The actual current wave shape through the diode is a series of narrow pulses, which represent the charging current of $C_L$ as it charges up to the peaks of $e_c$ through $r_d$. As $e_c$ passes from its positive peak value down to its peak negative value, $C_L$ discharges more slowly through the higher resistance $R_L$. The action of the diode is thus determined by two time constants: the charging time constant $C_L r_d$, and the discharge time constant $C_L R_L$. For a constant amplitude carrier of peak value $E_c$, the output voltage $e_L$, the input voltage $e_c$ of peak amplitude $E_c$, and the diode current $i_d$, are as shown in Fig. 3. The wave form for $e_L$ is exactly that for an ordinary condenser input filter. If $R_L >> r_d$, and $1/\omega_c C_L$ is negligibly small, the ripples in $e_L$ are negligible, and $e_L$ is practically equal to $E_c$. In this case $i_d$ consists of exceedingly narrow impulses. It is an interesting fact that such a wave form for $i_d$, when analyzed into a Fourier Series, yields a fundamental a-c component, $i_c$ (of frequency $f_c$), which is practically twice the d-c component $i_o$. This is practically true regardless of the wave form of the pulses, so that we may set

$$i_c \cong 2i_o \quad \ldots\ldots\ldots\ldots\ldots (2)$$

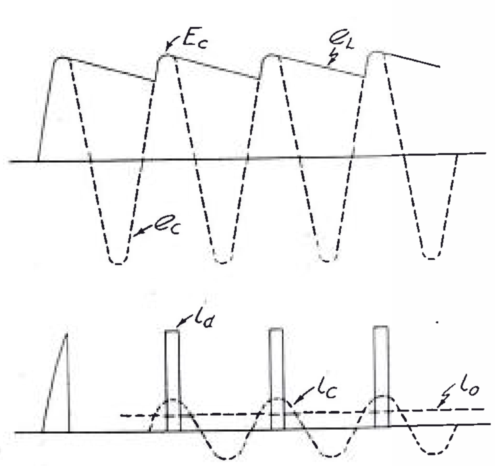



**Fig. 3.**

As the pulses become narrower and narrower, (2) approaches an equality more and more closely. This is the case as $R_L$ exceeds $r_d$ more and more, and such are the conditions in normal practice. If $R_L$ does not greatly exceed $r_d$, then as the pulses become broader their shape is of more consequence, and $i_c$ decreases from its value of $2i_o$. Over quite a large range of practical values of $R_L$ and $r_d$, such as, for instance, specific values of ¼ megohm and 5000 ohms, respectively, (2) is practically an equation. From the above it is apparent that $R_L$ appears to the source of $e_c$ as one-half of its actual value,—i.e.

$$R_c \cong \tfrac{1}{2} R_L \quad \ldots\ldots\ldots\ldots (3)$$

A further refinement is to take $r_d$ into account as well as $R_L$, so that somewhat more accurately

$$R_c \cong \tfrac{1}{2} (R_L + r_d) \quad \ldots\ldots (4)$$

However, as stated above, $r_d$ is usually negligible in broadcast practice as compared to $R_L$.

If $e_c$ is a modulated wave, further complications arise. If the envelope drops too rapidly, $C_L$ in conjunction with $R_L$ may fail to follow this drop, and the inward peaks of modulation will be "clipped" and the output distorted. The rectifier is now rectifying the envelope as well as the carrier wave itself. If $f_c >> f_m$, this clipping can be avoided by a suitably low choice of the time constant $C_L R_L$. Fig. 3 must now be replaced by Fig. 4. Actually the pulses of $i_d$ during inward modulation will be less than those during outward modulation because of the fact that $C_L$ cannot discharge as rapidly

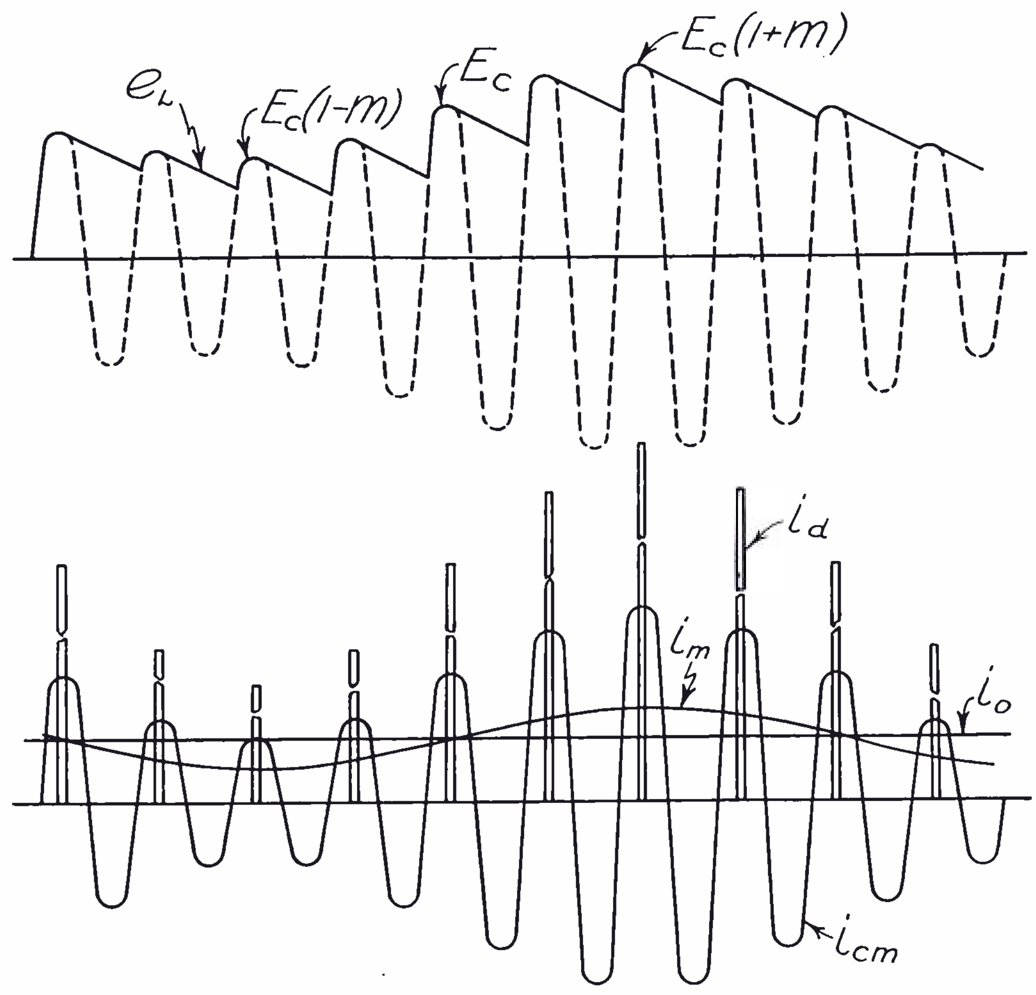



**Fig. 4.**

through $R_L$ as it can charge through $r_d$, but this factor will be ignored, and the pulses represented as varying symmetrically with the envelope. This corresponds to the pulses and components of $i_d$ of Fig. 3 being modulated by the factor m sin $\omega_m t$. As a consequence, instead of $i_o$, we have an output current composed of the d-c component $i_o$, acting as a (d-c) carrier for an a-c wave $i_m$ of frequency $f_m$. Similarly, the high frequency component is $i_{cm}$, which in turn can be broken up into a term of constant amplitude, $i_c$, of frequency $f_c$, and two side bands, $i_{cm1}$, of frequency $(f_c + f_m)$, and $i_{cm2}$, of frequency $(f_c - f_m)$.

It is evident that, similar to Eq. (2), the following approximate equations hold:

$$i_c \quad 2i_o$$

$$i_{cm1} + i_{cm2} \quad 2i_m \quad \ldots\ldots (5)$$

If, tentatively, $i_{cm1}$ is assumed equal to $i_{cm2}$, then

$$i_{cm1} = i_{cm2} = i_m \quad \ldots\ldots\ldots (6)$$

We now have, in addition to Eq. (4), the following:

$$R_{em1} = R_{em2} \cong (R_L + r_d) \quad \ldots\ldots\ldots\ldots\ldots(7)$$

where $R_{em1}$ and $R_{em2}$ are the apparent impedances which $(R_L + r_d)$ presents to side band voltages of frequencies $(f_c + f_m)$ and $(f_c - f_m)$ respectively.

These two apparent impedances can be combined into a single equivalent impedance of $2(R_L + r_d)$ in the following manner.

In Fig. 5 is shown a vector representation of a modulated wave[2]. The carrier voltage vector, of constant amplitude, is shown as $E_c$. Around it rotate in opposite directions at a frequency $f_m$, the two side-band voltages, each $mE_c/2$ in peak amplitude. Their resultant at any instant, $me_c$, is clearly collinear with $E_c$, and varying at the frequency $f_m$ from a positive peak amplitude of $mE_c$ to a negative peak amplitude of $-mE_c$. The equation of this voltage is therefore

$$e = me_c = (m \sin \omega_m t)\, E_c \sin \omega_c t \quad \ldots\ldots\ldots\ldots\ldots(8)$$

We may regard the quantity $(m \sin \omega_m t)$ as a modulating operator of the carrier voltage $E_c \sin \omega_c t$, and its product by $E_c$ also represents the equation of the envelope. At any rate, to $e = me_c$, the apparent impedance of the diode load is

$$R'_{em} = (\tfrac{1}{2})\, R_{em1} = (\tfrac{1}{2})\, R_{em2} = (\tfrac{1}{2})\,(R_L + r_d) \quad \ldots(9)$$

Now suppose that $C_L$ does not have negligible susceptance to $e_L$, that is—$C_L$ by-passes $R_L$ appreciably at the modulation frequency $f_m$. Then the a-c component of $i_L$ must be broken up into two quadrature components: one, $i_{LC}$ through $C_L$ that leads the a-c component of $e_L$ by 90°, the other, $i_{LR}$ through $R_L$, that is in phase with $e_L$. The total must have been produced by a modulated carrier current capable of producing this through the diode.

Consider a modulated carrier current whose equation is

$$i = I\,[1 + m \sin(\omega_m t + \theta)]\, \sin \omega_c t \quad \ldots\ldots\ldots\ldots\ldots(10)$$

This can be expanded into the form

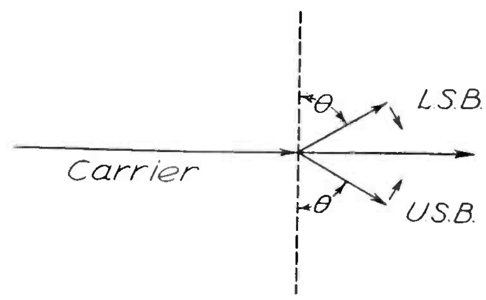
$$i = I \sin \omega_c t - (mI/2)\cos(\omega_c t + \omega_m t + \theta) + (mI/2)\cos(\omega_c t - \omega_m t - \theta) \quad \ldots\ldots\ldots\ldots\ldots(11)$$

The process of detecting or demodulating this current consists (among other things) of beating the side bands with the carrier. The two difference beat frequencies will be

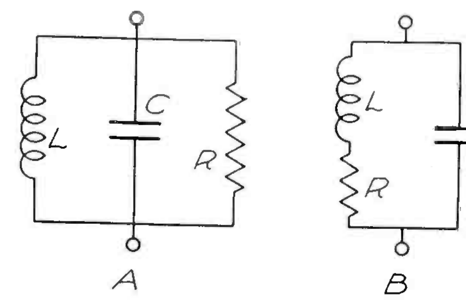
$$i_d/2\,[\sin(\omega_m t + \theta) + \sin(\omega_m t + \theta)] = i_d \sin(\omega_m t + \theta) \quad \ldots(12)$$

[2]See, for instance, Laport—"Characteristics of Amplitude Modulated Waves," *RCA Review*, April, 1937.

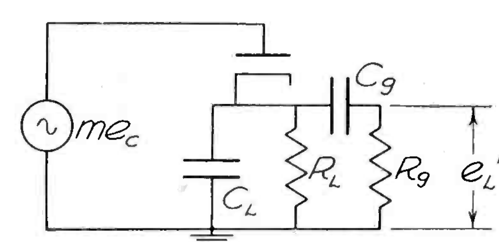
The vector diagram for the current defined by Eqs. (1) and (11) is shown in Fig. 6 at the instant $t = 0$. The reference axis is along the vertical. If $\theta$ were not present, the lower and upper side bands would be respectively vertically up and down at this instant of



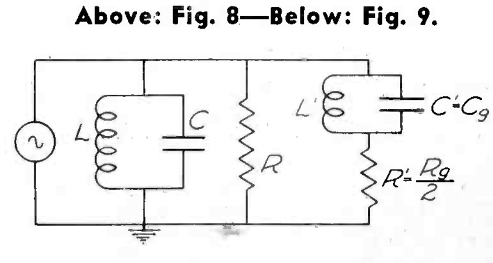

**Above: Fig. 6—Below: Fig. 7.**




time. The presence of $\theta$ causes both vectors to take up the positions shown. Thus the lower side band (L.S.B.) lags from the vertical by $\theta$. Their resultant nevertheless remains collinear with the carrier, so that no envelope distortion occurs. The envelope, however, regarded as a wave shape, leads its position for $\theta = 0$ by the quantity $\theta$. This wave, upon detection, gives rise to an a-c current of frequency $f_m$, that leads the voltage $e_L$ by the angle $\theta$. But it is evident from the preceding discussion that $e_L$ is in phase with the envelope of the modulated voltage wave $me_c$. Hence a modulated current i




**Above: Fig. 8—Below: Fig. 9.**




whose envelope leads that of the modulated voltage $me_c$ by the angle $\theta$, and whose upper and lower side bands therefore lead and lag, respectively, the corresponding side bands of the voltage wave, give rise to a demodulated current $i_d$, which leads the demodulated voltage $e_L$, by the same angle $\theta$.

We may at least expect the converse to be true, namely—if $i_d$ leads $e_L$ by $\theta$, then the modulated high-frequency current has upper and lower side bands that lead and lag those of the modulated high-frequency voltage. At the same time, since the actual current is in the form of pulses, the sum of the side band currents is twice $i_d$.

We shall now attempt to find a form of impedance which, when placed directly in the high-frequency circuit, has the same effect upon the current components as $Z_L$ acting through the diode. This effect can be expressed in this manner: the diode translates the side bands down in the spectrum from the carrier frequency $f_c$ to a zero frequency carrier. The lower side band would therefore have a negative frequency (be below zero in frequency). A negative frequency is physically equivalent to a positive frequency, i.e., the angle of the lower side band is reversed in sign, and thus the two side bands become directly additive. In the case of the upper and lower side band currents respectively leading and lagging the corresponding voltage side bands, the translation through detection produces the following results. The lower side band voltage becomes one of negative frequency and hence becomes directly additive to the upper side band to form the total voltage $e_L$ (a-c component) of frequency $f_m$. The lower side band current becomes one of negative frequency and lagging angle $\theta$. Upon reversing to obtain a positive frequency wave, $\theta$ changes sign, too, to become a leading phase angle. The lower side band current thus becomes in phase with the upper side band current after detection, and is therefore directly additive to the latter, to form a total current, $i_L$ (a-c component) which leads $e_L$ by the angle $\theta$. This is a very striking and interesting feature of detection.

The high-frequency impedance equivalent to $Z_L$ is a parallel resonant circuit shunted by a resistance, Fig. 7(A). Suppose it is anti-resonant at $f_c$. Then

$$LC = 1/\omega_c^2 \quad \ldots\ldots\ldots\ldots(13)$$

The impedance at this frequency is evidently simply R, and this corresponds to $(R_L + r_d)/2$ by Eq. (7). Now consider the admittance A of the LC portion of the circuit at the frequencies $(f_c \pm f_m)$, i.e.—at the upper and lower side band frequencies, respectively.

$$A = 1/j\,(\omega_c \pm \omega_m)\,L + j\,(\omega_c \pm \omega_m)\,C \quad \ldots(14)$$

Fq. (14), in conjunction with Eq. (13), gives

(*Continued on page 22*)

[3]See Wheeler—"Design Formulas for Diode Detectors"—*IRE Proc.*, June, 1938.

# VETERAN WIRELESS OPERATORS ASSOCIATION NEWS

W. J. McGONIGLE, President — RCA BUILDING, 30 Rockefeller Plaza, New York, N. Y. — GEORGE H. CLARK, Secretary

### E. A. Nicholas—Life Member

WE introduce to our members and readers, this month, one of our most prominent Life Members, E. A. Nicholas, President of the Farnsworth Television and Radio Corporation. His wireless career began as a wireless operator with the United Wireless Company back in 1909 for whom he served on many vessels. Later he was promoted to shore duty at New Orleans and then became Superintendent of the RCA Marine Division on the Great Lakes. In 1924 he served as Assistant to Mr. Sarnoff, then Vice-President and General Manager of the Radio Corporation of America.

Excepting a short period during which Nicholas headed a wholesale distributing company in Chicago, his service in the radio business has been with the Radio Corp. of America and its predecessor, Marconi Wireless Telegraph Co. of America, which he joined in 1912. In the World War period he was a wireless operator in the merchant marine.

During the early period of his association with these companies Nicholas held various positions and in recent years he served as Sales Manager of the eastern division of Radio Corp. of America, Vice-President in charge of sales of RCA-Victor Co. and Manager of the licensing division and a member of the advisory board of RCA.

Nicholas, in 1930 and 1931, was President of E. A. Nicholas, Inc., of Chicago, wholesale distributor of RCA products.

The Farnsworth company, with plants at Fort Wayne and Marion, Ind., manufactures and sells radio receivers and radio-phonographs, television transmitting and receiving apparatus and other special electronic devices.

The company, incorporated Dec. 13, 1938, acquired its plants in April. The start of production in September was preceded by five months of preparatory planning and building, during which Nicholas carried the principal burden. He formulated the licensing policies adopted by the company and carried on negotiations with prospective licensees. One result of this work was the signing of a broad patent license agreement between Farnsworth and Radio Corporation of America. Under the direction of Mr. Nicholas, the company's program of research, development and manufacturing was worked out. He also guided the building of an executive staff and supervisory force, re-arrangement of the company's manufacturing plant, designing of new products, formulation of sales and advertising plans, selection of skilled workers, the beginning of production and the introduction of merchandise.

The magnitude of the task performed by Nicholas in this preparatory period, constituting a remarkable feat in organization and planning, has attracted wide attention throughout the industry.

Mr. E. A. Nicholas is one of the country's outstanding executives in the radio field. His selection, in February, 1939, by the board of directors, was based on his record of accomplishments during a long career in the industry.

We congratulate EAN on a long and notable career in the radio field and on behalf of all VWOA members we extend our best wishes for his continued success.

**E. A. Nicholas, President of the Farnsworth Television and Radio Corporation, who recently became a Life Member of the VWOA.**

### Marconi Scholarship

J. R. Poppele, Secretary and Chief Engineer of the Bamberger Broadcasting Service and Chairman of our Marconi Memorial Scholarship Committee, announces a winner for our Second Annual Scholarship at RCA Institutes in Radio and Electrical Communication.

The contest for the Scholarship was participated in by members of science clubs in high schools all over the United States and was conducted by the American Institute of the City of New York. The American Institute sponsors science clubs in over 600 high schools in 44 states and the contest was open to all members of each of them.

Mr. Poppele announces that Robert Joseph Stahl, a student of Bellarmine College Preparatory School of San Jose, California, was the outstanding participant in the contest. Robert resides at 51 North Umberland Avenue, Redwood City, Cal. He will take the course at either the New York or Chicago school of RCA Institutes.

It will be a great thrill to Mr. Stahl to have the Scholarship presented to him by the "Father of Radio" Dr. Lee de Forest, Honorary President of our Association, on a coast-to-coast network of the Mutual Broadcasting System—the program originating in the studios of KHJ, Los Angeles. The presentation will be made on Saturday, August 17, 1940.

### De Forest Day

The combined efforts of all engaged in the radio and allied fields on the Pacific Coast are being enlisted by Leroy Bremmer, Secretary of the Los Angeles-Hollywood chapter of our Association, in the furthering of the plans for "DE FOREST DAY" at the San Francisco World's Fair on Saturday, September 7, 1940. There is no question but that this event will be a most outstanding one in the life of the "Father of Radio" and all interested are invited to participate. Communications relative to this "day" should be addressed to Leroy Bremmer at the National Schools, Los Angeles, or to Gilson Willets, Chairman of the San Francisco Chapter, in San Francisco. Our best wishes from the East Coast for the success of "DE FOREST DAY."

## Oscillator— continued from page 9

low $Q_g$ brings down the entire grid frequency curve. The combination of $Q_a$ and $Q_g$ may determine whether there will be oscillation at all, and if so, what its frequency will be. Generally, one may state that the point of the intersection falling too far to the right on the chart of Fig. 5, indicates a large phase angle of feasible oscillations, hence, poor efficiency. In each particular case power output and efficiency can be calculated by the graphical method using the constant current chart for the tube[2].

If the intersection of the frequency curves falls within the "dead zone," no oscillation can be produced. Generally speaking, this would usually correspond to a low $Q_a$ and a high $Q_g$. From this viewpoint a low Q in the anode and a high Q in the grid circuit are favorable for the suppression of oscillation in an amplifier.

(3) The value of $C_{pg}$ determines both the feed-back resistance $R_x$ and the reflected capacitance $C_x$. With a decreasing $C_{pg}$, the first factor increases, hence the dead zone widens, and it becomes more difficult to produce oscillations. This justifies the use of the screen grid tubes in amplifier circuits. In these tubes, as is well known, the plate-to-grid capacitance is reduced to its physical minimum.

(4) Amplification factor, $\mu$ appears in the expression (9) for $A_1$ and $A_2$, both of which are directly proportional to $\mu$. From equations (9), (1) and (2), one can conclude that a high $\mu$ makes the dead zone narrow by rendering $R_x$ small; also, it brings the grid frequency curve downwards by increasing the reflected capacitance, $C_x$; hence it shifts the point of intersection away from the dead zone. Thus, generally speaking, a high $\mu$ favors oscillations.

(5) It is more convenient to examine the role of the load resistance $R_L$ simultaneously with that of the tube internal resistance, $R_p$; one can designate

$$k = R_p/R_L \qquad (19)$$

We also have

$$d = DG \qquad (20)$$

Using *k* and *d* in equations (10) and (9), one can express the components of the voltage-ratio, A, as follows:

$$A_1 = \frac{\mu\,(1 + 1/k)}{k\,(1 + d^2) + 2 + 1/k} \qquad (21)$$

and

$$A_2 = \frac{\mu d}{k\,(1 + d^2) + 2 + 1/k}$$

From these last expressions, it is obvious that either a very large *k*, or a very small *k* leads to low values of $A_2$, hence high $R_x$. This means that both very high and very low values of $R_p/R_L$ are not favorable for generating oscillations, but they are useful for suppression of parasitic oscillations in an amplifier.

By differentiating the expression for $A_2$ with respect to k, one will find that a maximum of $A_2$ corresponds to

$$k = 1/\sqrt{1 + d^2} \qquad (22)$$

Since in practical cases *d* should not be too large in order to avoid a large phase angle, one can assume that

$$0 < d < 1 \qquad (23)$$

Then, the optimum values of *k* are given by

$$.7 < R_p/R_2 < 1$$

On the other hand, for a fixed *k*, one can find the maximum for $A_2$ by differentiating $A_2$ with respect to d and setting $\delta A_2/\delta d = 0$. The solution is

$$d = 1/k + 1 \qquad (24)$$

and

$$A_{2\,max} = \mu/2(1 + k) \qquad (25)$$

(6) *Resonance frequency of the circuit.* In the expression for $R_x$ (equation 1), the factor $A_2$ is independent of magnitude of the resonant frequency, $\omega_r$; it varies only with detuning, d. On the other hand, the actual frequency of oscillation $\omega_a$, departs but very little from the resonance value $\omega_r$. Therefore, for any particular value of $A_2$, the higher the order of magnitude of resonance frequency, the lower is the feed-back resistance, $R_x$, hence the more readily oscillations start, all other conditions being kept unchanged. However, a limit to an indiscriminate increase of generated frequencies, may be established by the inter-electrode capacitance, which can become greater than required for the building of an adequate resonance circuit. In addition, when the resonant frequency becomes very high (say ultra-high), the finite transit time of electrons introduces new limitations, one of them being rapidly decreasing parallel grid resistance, $R_g$; this widens the dead zone.

If, in a particular case, one can evaluate the positive grid resistance, $R_g$, one can calculate the limiting value of the inter-electrode capacitance which permits generation of oscillations due to the feed-back through the tube. It will be found from the following relation:

$$/R_x/ = 1/\omega_a A_2 C_{pg} < R_g \qquad (26)$$

Using expression (25) for $A_2$ and $\omega_r$ for $\omega_a$, one has

$$C_{pg} = 2(1 + k)/\mu\,\omega_r R_g \qquad (27)$$

#### Summary

The above analysis gives one an insight into the mechanism of the generaiton of oscillations in a tuned-plate tuned-grid circuit; it accounts for the role of various individual factors contributing to or interfering with this phenomenon; and permits one the prediction, at least approximate, of the limiting condiitons for starting or not starting oscillations.

#### References

[1]H. Barkhausen. "Elektronen-Röhren," 2. Band Verstärker, §30. 1933.

[2]I. E. Mouromtseff and H. N. Kozanowski. "Analysis of the Operation of Vacuum Tubes as Class C Amplifiers," *Proceeding of I.R.E.*, July, 1935.

## Diode Detection— continued from page 20

$$A = \frac{1 - 1 \mp 2m/\omega_c - \omega^2{}_m/\omega_c^2}{j\,(\omega_c \pm \omega_m)\,L} \qquad (15)$$

$$\approx \pm 2j\omega_m C$$

since if $\omega_c >> \omega_m$, the term $\omega^2_m/\omega_c^2$ is negligibly small. The corresponding impedance is therefore

$$Z \approx \mp j/2\,\omega_m C \qquad (16)$$

This is an inductive reactance to the lower side band voltage, and a capacitive reactance of equal amount to the upper side band voltage. It will thus give rise to leading upper, and lagging lower side band currents, which, upon detection in a pure resistive load, will give rise to a current $i_L$ which leads the voltage $e_L$. Hence the impedance Z is potentially equivalent to $Z_L$. The equivalence is complete if[3]

$$C = C_L;\ L = 1/\omega_c^2 C;$$
$$R = (R_L + r_d)/2 \quad \ldots(17)$$

because the high-frequency currents total to double $i_L$. More accurately, $r_d$—the diode resistance—should appear in series with Z for complete equivalence, but in practical cases can usually be ignored, as well as in the third equation (for R).

For $\omega_m << \omega_c$, Fig. 7 (B) is practically equivalent to 7 (A) if

$$R = (\omega^2 L^2)/R' \quad \ldots\ldots(18)$$

and so either circuit may be used to represent the apparent impedance that the diode and load present to the high-frequency source.

One further type of diode load of practical interest is shown in Fig. 8. The circuit $C_g R_g$ permits the grid of the following tube to be coupled to the diode load without assuming the d-c bias voltage developed across $R_L$. Its equivalent high-frequency impedance is represented in Fig. 9. The condenser $C_g$ appears as L′ and C′ in parallel, and these are in series with R′, which represents $R_g$. At sufficiently high modulation frequencies, $C_g$ practically passes out of the picture, as does therefore L′ and C′, so that R′ becomes simply parallel to R, as does $R_g$ and $R_L$.

• • •

## OVER THE TAPE

### CHANGE OF ADDRESS

The offices of the Bakelite Corporation and the Halowax Corporation have been moved from 247 Park Ave., New York City, to the Carbide and Carbon Bldg., 30 East 42nd St., New York City. The change was effective July 1st.

### SYNTHANE PRESS ROOM

Synthane Corporation, manufacturers of Bakelite-laminated products, have just completed a new press room addition to their Oaks Plant. The first piece of equipment placed in operation was a new laminating press for handling special lengths up to 100″.

### WORNER PRODUCTS APPOINTMENTS

L. L. Worner, President of the Worner Products Corporation, manufacturers of photo-electric equipment, have announced the appointment of Robert H. Campbell as General Sales Manager. Mr. Campbell is well known in both the radio and electrical field.

### GENERAL CEMENT CATALOG

The General Cement Mfg. Co., 1041 Kilburn Ave., Rockford, Ill., have recently issued a 32-page catalog covering their line of radio chemicals and products. Copies may be secured by writing to the above organization for Catalog No. 141.

### FINCH BULLETIN

Finch Telecommunications, Inc., 1819 Broadway, New York City, have recently issued a very interesting bulletin describing their facsimile equipment for aviation work. Technical details are given. Write to the above organization to secure a copy.

### INSTRUMENT LANDING

The International Telephone Development Company, subsidiary of the International Telephone and Telegraph Corporation, has received an order amounting to $537,547 from the Civil Aeronautics Authority to manufacture and install airplane instrument landing systems at the airports of six cities in the United States: La Guardia Field, N. Y.; Municipal Airports at Chicago, Cleveland and Kansas City; Mines Field, Los Angeles, and Meacham Field, Fort Worth.

With the exception of experimental installations, these instrument landing systems will be the first ever contracted for by the United States Government for utilization by the commercial airways.

### DAVEN CATALOG

The Daven Co., 158 Summit St., Newark, N. J., have recently issued an interesting and informative catalog containing data on their line of attenuators, attenuator networks, rheostats, potentiometer controls, resistance boxes, output power meters, volume-level indicators, power-level indicators, decade voltage dividers, logarithmic resistors, program line equalizers, transmission measuring sets, etc. Copies may be secured by writing to the above organization.

### DUMONT CATALOG

The Allen B. Du Mont Laboratories, Inc., 2 Main Ave., Passaic, N. J., have made available a 68-page catalog covering their line of cathode-ray instruments. This booklet contains a great deal of data on oscillographs, cathode-ray modulation monitors, oscillograph assemblies, amplifiers, regulated power supplies, electronic switches, television signal generator, as well as a complete line of television studio and transmitting equipment. A copy of the catalog may be secured by writing to the above organization.

### GENERAL INDUSTRIES BULLETIN

The General Industries Co., Elyria, Ohio, have recently issued a bulletin describing their line of phonograph motors, recording assemblies, automatic record changers, etc. Copies may be secured by writing to the above organization.

### N. U. DIRECTORS

Reelection of S. W. Muldowny as President of National Union Radio Corporation was announced following the organization meeting of the Board of Directors. At the same time W. R. Wilson was named Treasurer and E. O. Sandstrom, formerly Acting Secretary, was elected Secretary and Assistant Treasurer.

At the company's annual meeting, held in Wilmington, Del., the following persons were elected to the Board of Directors: S. W. Muldowny, President, National Union Radio Corporation; Henry L. Crowley, President, Henry L. Crowley Manufacturing Company; Paul V. Galvin, President, Galvin Manufacturing Company, Chicago; Penn Brook, Vice-President in charge of factories for Sears-Roebuck and Company, Chicago; W. R. Wilson, Controller, Philco Corporation; and Fred D. Williams, Assistant to the President, Philco Corporation.

### FINCH APPOINTMENT

W. G. H. Finch, President of Finch Telecommunications, Inc., of Passaic, N. J., manufacturers of facsimile apparatus and other equipment, has appointed James W. Baldwin as Assistant to the President and in charge of new offices in the Bowen Building, 815 Fifteenth Street, in the Nation's Capitol. Mr. Baldwin, former Secretary of the Federal Radio Commission, comes to the Finch organization with a great deal of experience.

### SYLVANIA APPOINTMENT

The appointment of Walter R. Jones, Emporium, Pa., to the post of Director of Commercial Engineering, Radio Tube Division, is announced by Hygrade Sylvania. Mr. Jones will administer and guide the activities of the company's engineering consultants and the popular Sylvania radio service schools which are conducted all over the country by Walter Jones and George C. Connor of New York.

### WARD LEONARD BULLETIN

Ward Leonard Electric Co. have announced a new, revised circular No. 507, which describes their line of radio resistors. Write to the above organization at Mount Vernon, New York.

### UTC CATALOG

United Transformer Corp., 150 Varick St., New York City, has issued bulletin PS-404 covering their 1940-41 line of equalizers, filters, audio transformers, microphone cable transformers, automatic voltage regulators, etc. The manufacturer will send catalog on request.

### CALLITE ACQUIRES HARRIS ALLOYS

Charles H. Kraft, President of the Callite Tungsten Corporation, Union City, New Jersey, announces the acquisition of Harris Alloys, Inc., by his company. The business of the former Harris Alloys will be conducted as a division of and under the corporate name of the Callite Tungsten Corporation. Frederick T. Harris, former Harris executive head, joins the Callite Tungsten executive staff and will have full charge of operations of this new division. Sales activities will be under the personal supervision of D. R. Donovan, General Sales Manager of the Callite Tungsten Corporation. This assimilation of the business of the Harris Corporation enables Callite Tungsten Corporation to add to its regular list of products a line of special wires in sizes down to .001", including aluminum, stainless steels, nickel silver, everdur, brass in all grades, brush wire, commercial bronze, phosphor bronze and fourdriniering, silver, monel.

### HOLLYTRAN APPOINTS AGENT

Eddy Woodward, owner of the Hollywood Transformer Company, announces the appointment of Norman B. Neely as national sales representative. This company specializes in the design and custom building of Hi-Q chokes, audio transformers and similar items for special applications.

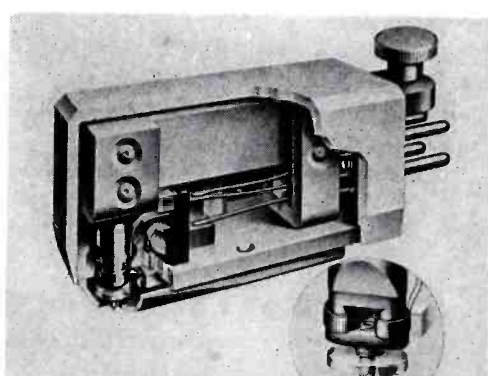
### JEFFERSON MACHINE BULLETIN

An interesting bulletin has been made available by the Jefferson Machine Tool Co., Fourth, Cutter and Sweeney Sts., Cincinnati, Ohio. This bulletin contains data on swing frame grinding and polishing machines, endless belt sanders, gyratory foundry riddle, turret lathes, etc. Write to the above organization.

## THE MARKET PLACE

### UNIVERSAL PICKUP

A universal phonograph reproducer which plays both the vertical cut and lateral cut records has been introduced by the Western Electric Company, 195 Broadway, New York City. At the flip of a switch, the new reproducer may be converted for either type of "cut." This convenience simplifies operation and maintenance. It also eliminates all chance of selecting an incorrect reproducer when shifting to records of different types because the transfer can be made without lifting the stylus of the reproducer from its groove in the record. The response of the 9A is essentially flat up to nearly 10,000 cycles for both types of recording. Most lateral cut records are made at a slightly higher level than those recorded by the vertical method. Bell Telephone Laboratories, therefore, designed the 9A reproducer with increased sensitivity for vertical recordings and thus made the output volume of the instrument approximately the same for both types.



Two main assemblies, a self-contained vibrating system and a permanent magnetic circuit, comprise the internal mechanism.

The vibrating system of the 9A differs basically from conventional dynamic reproducers in that it employs two adjacent voltage-generating coils instead of one. These coils are mounted on a common framework of duralumin and vibrate axially in a radial magnetic field. Supporting this structure and mounted midway between the two coils, is a flat triangular shaped spring that can flex up or down and twist axially but cannot flex sideways because of its wide cross section. Recorded sound vibrations, in the form of undulations in the record groove, impart motion to the coil structure through a thin duralumin tube. This tube, which extends downward from the midpoint of the coil structure, carries the stylus at its lower end. During the reproduction of vertical records, the rise and fall of the tube carries both coils up and down simultaneously with it. Lateral cut records, on the other hand, swing the tube sideways like a pendulum. Consequently the coils continue to move up and down; but with lateral records they travel in opposite directions after the fashion of a see-saw. Hence, by switching the electrical connections of the coils (series aiding or series opposing depending

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The magnetic circuit consists of a rectangular bar of magnetic alloy to which are riveted two soft iron "U" shaped yokes, one of which carries the center pole-piece. The two yokes are secured directly to the outside pole plate which serves as a mounting for the vibrating system.

Since good studio practice calls for the introduction of varying amounts of equalization during the recording process to overcome the effects of noise originating in the record material and other distorting factors, an equalizer capable of introducing a series of complementary characteristics has been designed for use with the 9A reproducer. This unit is known as the 171A repeat coil and K. S. 10066 switch. It serves not only as an adjustable equalizer, but as a means of matching the impedance of the reproducer to the input of the amplifiers. These input values may be 30, 250, 500, or 600 ohms.

### FREQUENCY MONITOR

The Browning Laboratories, Winchester, Mass., have recently announced a custom-built laboratory calibrated frequency monitor for checking any three bands of

frequencies from 1.5 to 60 mc. A 100 kc crystal is used as a secondary standard. This may be readily checked against WWV. E-c oscillators are used to cover a band of frequencies from 50 to 100 kc wide. The transmitter frequencies which it is desired to check are included in this band. The circuit is so arranged that the electron-coupled oscillators can be accurately checked at numerous points by means of the 100 kc crystal. Transmitter frequencies are checked by the zero-beat method which is indicated visually on a cathode-ray tube and aurally by means of phones plugged into a jack provided. The accuracy of this frequency monitor is better than 1/100 of 1%, it is said.

### F-M RECEIVERS

The Meissner Mfg. Co., Mt. Carmel, Ill., have announced two models of frequency-modulation receivers. Both receivers are identical except that one is a table model, while the other is a console. A bulletin may be secured from the manufacturer.

### MERCURY-VAPOR RECTIFIER

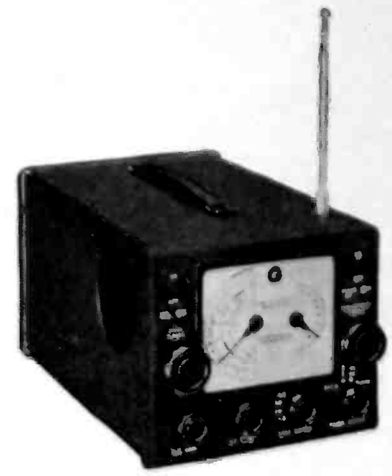
A new GL-869-B mercury-vapor rectifier employing a horizontal mesh filament has been introduced by the General Electric Company, Schenectady, N. Y. The filament structure of the new tube, proved in both the GL-266-B and GL-857-B transmitting tubes, makes it possible to double the average anode current of previous 869's when filaments are connected in quadrature. The GL-869-B is expected to find considerable application in 50-kw frequency modulated transmitters. It supersedes the GL-869-A.

When using in-phase filament excitation, the GL-869-B is capable of a maximum inverse anode voltage of 20,000 volts with natural ventilation. The average maximum anode current at 25 cycles and above is 10 amperes.

For quadrature filament excitation, the average maximum anode current is 15 amperes. The maximum peak inverse anode voltage is 15,000 volts using forced ventilation.

### PORTABLE COMMUNICATIONS RECEIVER

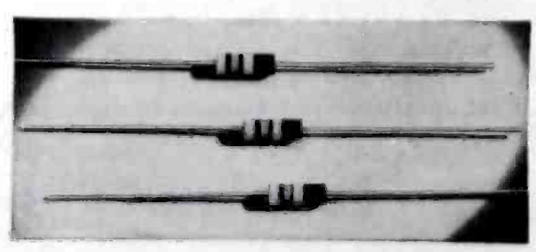
Something new in receiver equipment is the Hallicrafters Model S-29 "Sky Traveler" which combines the universal convenience features of the modern 3-in-1 portable with those of a communications receiver. The "Sky Traveler" is housed in a crackle-finished aluminum case with



carrying handle, 7 inches high, 8½ inches wide and 13¼ inches deep. The weight with self-contained batteries is 18 pounds. Operation is from any 110-volt a-c or d-c line or from the batteries. Battery life is prolonged by a built-in charging circuit. Nine tubes provide one r-f and two i-f stages, mixer, detector and avc, two audio stages, beat oscillator, automatic noise limiter, and line rectifier. The tuning range is continuous from 542 kc to 30.5 megacycles in four steps and electrical band-spreading is provided for all parts of this range. Sensitivity averages better than 2 microvolts in all ranges. The Hallicrafters, Inc., 2601 S. Indiana Ave., Chicago, Ill.

### MOLDED RESISTORS

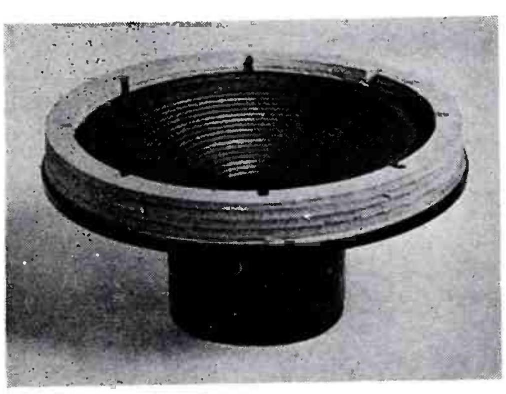
Erie Resistor Corporation announces a new molded type resistor designed to carry ½ watt load at 40° C., indefinitely. Designated as Type 523, this unit is compact, measuring but ⅜" long x .143" diameter.



Wire leads 1½" long are paralled to the axis of the body. This wire has a specially developed alloy-coated surface that resists oxidization and retains its excellent solderinging characteristics for a long period of time. Samples of the new Type 523 will be sent on request by writing Erie Resistor Corporation, 640 West 10th Street, Erie, Pa.

### LOUDSPEAKER

Representing an important improvement in loudspeaker design, a new type of "accordion edge" loudspeaker has been developed by the RCA Manufacturing Co., Inc., Camden, N. J., to reproduce low frequen-



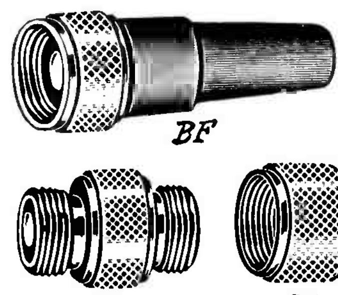
ciese with a small speaker in a small cabinet. Although only 7 inches in diameter, the new instrument has a frequency response of from 80 to 7,000 cycles. The new loudspeaker makes effective use of a folded or "accordion edge" cone support principle which permits the cone to move more freely when driven by the permanent magnet speaker mechanism. The cone was developed by Dr. H. F. Olsen in the RCA Radio Research Laboratoreis at Camden.

### PORTAPOWER

An interesting development in the equipment of marine transmitters is the Portapower unit manufactured by Electronic Laboratories, 122 W. New York St., Indianapolis. The Portapower combines a power supply for the complete marine transmitter and receiver, the supply being controlled by means of a microphone switch. One section of this power supply delivers filament power for the transmitter and receiver, while the other two sections deliver high-voltage direct-current output. It is constructed for either 12, 32 or 110 volts in marine applications. Also the wattage rating of this unit can be as high as 400 watts.

### BRUNO CONNECTORS

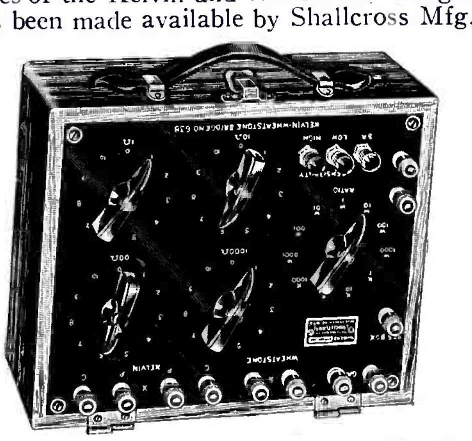
The new Bruno Baby Connectors are designed for the entire public address and allied industries. Of the locking type, these new sizes have been especially designed




for output or loudspeaker connections, and are equipped with half-inch 27 thread to prevent accidental mixing of cables with microphone input. Complete literature may be had by writing Selectar Mfg. Corp., 30 West 15th, New York City.

### KELVIN-WHEATSTONE BRIDGE

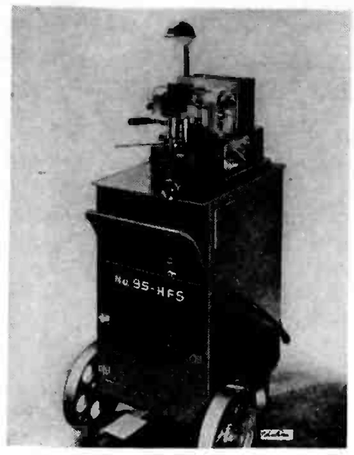
A portable instrument combining the features of the Kelvin and Wheatstone bridges has been made available by Shallcross Mfg.



Co., 10 Jackson Ave., Collingdale, Pa. Total resistance range is from 0.00001 ohm to 11.11 megohms. It may be secured either with or without a galvanometer. Write to the manufacturer for bulletin 146-2G.

### PORTABLE BUTT WELDER

The Eisler Engineering Co., 740-770 So. 13th St., Newark, N. J., has developed a new portable butt welder with a burr grinder and annealer mounted on a fabricated case which is portable. One of its many applications is in the repair of stranded wire where this unit is used to burn out (electrically) the faulty section and fuse all the strands, thereby preventing the ends from unraveling. The final steps are to butt weld the two ends, and grind off any burrs that may have appeared. The grinder has an adjustment to compensate for the wear in the abrasive and there are two special bushings to guide




the stranded wire to the grinder, which squares off the ends so as to make neater butt welds.

### BROADCAST CRYSTAL UNIT

A new crystal unit for broadcast service, the G30, with a guaranteed temperature-coefficient of less than one part per million per degree C has been announced by General Electric, Schenectady. The reallocation of broadcast station frequencies resulting from the Havana Treaty, and F.C.C. demands for more precise frequency control, have led to the development of this new crystal cell.

Approved by the F.C.C., the G30 is guaranteed to maintain transmitter frequency within ± 10 cycles at any specified point in the broadcast band. The low temperature-coefficient of the quartz plate is possible through the use of special x-ray equipment during manufacture which determines the angle of crystal cut. From a cold start, the new Thermocell is ready to go on the air in less than fifteen minutes. Only 2 watts of heating power are required, because of the small size and careful thermal design of the unit.

A hermetically-sealed metal-tube shell houses the G30 unit. The Thermocell's hermetic sealing protects the crystal from atmospheric dust, dirt, fumes and moisture. Of small size (¼ inch in

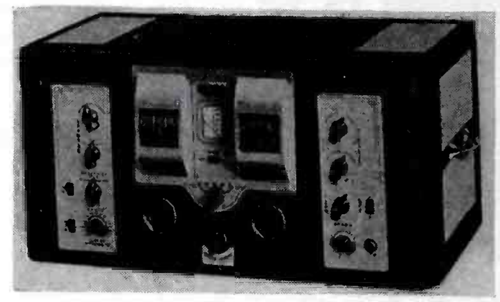
diameter), the unit fits standard octal sockets and is easily adapted to an existing circuit. Crystal electrodes are of stainless steel and the thermostat contacts are of nonoxidizing platinum-iridium. A sturdy bimetallic thermostat makes positive contact, preventing arcing and local heating of the thermostat.

The control ratio is more than 50:1. Control ratio is defined as ratio of room temperature change to change in mean crystal temperature.

Adjustment of the new Thermocell to almost the exact frequency is made at the crystal laboratory. Since the frequency is slightly dependent on the circuit in which the unit is used, final adjustment is made after installation in the crystal circuit by means of a trimmer condenser. Oscillator units in which the G30 is to be installed may be shipped to the General Electric crystal laboratory at Schenectady for free installation and adjustment, or the installation can be made by the station operator.

### COMMUNICATIONS RECEIVER

The new Howard 490 communication type receiver is designed for the advanced amateur, short wave listener and commercial operator. It features 14 tubes, coverage in six bands from 540 kc to 43.5 mc



(550 to 7 meters), calibrated bandspread, variable selectivity i-f, variable fidelity audio system, temperature compensated oscillator, air tuned i-f transformers and split stator ceramic insulated tuning condensers. Howard Radio Co., 1731-35 Belmont Ave., Chicago, Ill.

### ELECTRONIC BRIDGE

This new Hickok instrument, besides being used as a conventional bridge, can be used as a percentage bridge and also as a synchrometer. It permits electrical triangulation not possible with conventional


bridges. The circuit, which is shielded, features a cosine galvanometer, has built-in standards and provisions for external standards. The galvanometer cannot be injured by unbalance of the bridge, it is said. A-c power is used for operation. Despite the fact that it operates on a-c, the electronic bridge has null balance the same as all standard d-c bridges. Measurements include capacity, resistance, inductance, impedance, power factor and frequency. For complete information write the maker, The Hickok Electrical Instrument Co., 10514 Dupont Ave., Cleveland, Ohio.

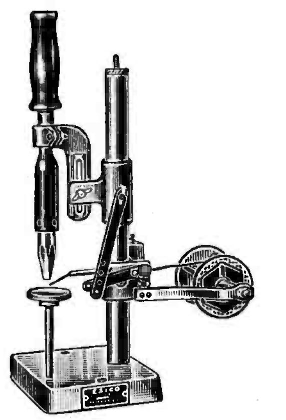
### RESISTORS

A complete line of commercially non-inductive IRC power wire wound resistors, from 10 to 200 watts and with any type of mounting, has been announced by the International Resistance Co., 401 N. Broad St., Philadelphia. These units utilize the Ayrton-Perry type of winding which is

said to assure full wattage ratings. IRC Resistance Data Bulletin IV and IV-A will gladly be sent upon request.

### SOLDERING MACHINE

In the accompanying illustration is shown a treadle operated soldering machine. In this unit the iron is advanced straight downward, the solder being fed



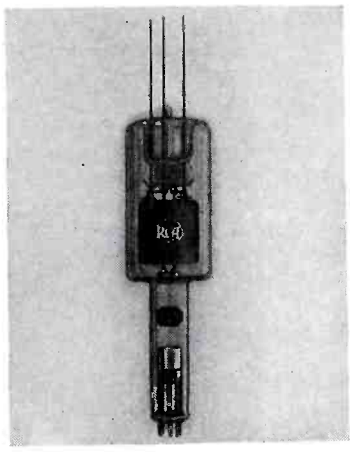
forward as iron returns from work. The feeding of the solder may be regulated. Solder up to 1/8" diameter may be used. Electric Soldering Iron Co., Inc., Deep River, Conn.

### RCA TUBES

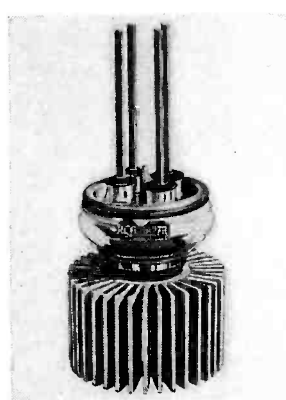
The RCA Radiotron Division, RCA Manufacturing Co., Inc., Harrison, N. J., have recently made available a number of new tubes designated as follows: 117N7-GT, 827-R, 825, 893-R, 5W4-GT, 880, 889-R and VR75/30.

The 117N7-GT is a multi-unit tube containing a half-wave rectifier and a beam power amplifier in the same envelope (T-9). It is intended primarily for use in portable battery/a-c/d-c receivers. The power amplifier unit delivers 1.2 watts with 100 volts on plate and screen. The heater is designed for operation directly across a 117-volt power-supply line.

The 827R is a new air-cooled radiator type of u-h-f transmitting beam power amplifier. It has a maximum plate dissipation rating of 800 watts in Class C telegraph service, and is particularly suitable for



use in frequency-modulation and television transmitters. It has several unique design features which include multiple-ribbon filament leads, two multiple-ribbon grid leads to minimize the effect of lead inductance, and an entrant metal header. The header serves not only as a low-inductance terminal for the screen but facilitates isolation of the input and output circuits. As



a result, neutralization is usually unnecessary except at the very high frequencies.

The RCA-825 is a new type of multi electrode tube in which the electron stream is inductively coupled to the output circuit. The 825 is designed for use as a power amplifier at frequencies above 300 megacycles and in such service is capable of handling power outputs up to 35 watts, depending on bandwidth and type of service. Because of its high transconductance and its adaptability to tank circuits having low effective capacitance, the 825 is especially suited for wide-band service, such as television and frequency modulation. The 825 may also be used as an oscillator and harmonic generator. Opening up new possibilities for research and development in the ultra-high-frequency transmitting field, the 825 will be of particular interest to experimenters and engineers working at frequencies above 300 megacycles.

The 893-R is an air-cooled radiator type of transmitting triode capable of dissipating 20 kilowatts of power in Class C telegraph service. The 893-R may be operated with maximum ratings at frequencies as high as 5 megacycles.

The 5W4-GT is a high-vacuum full-wave rectifier with T-9 bulb for use in a-c receivers having relatively low current requirements.

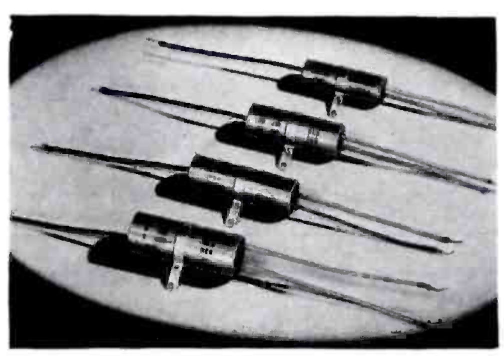
The 880 is a three-electrode tube with water-cooled plate capable of dissipating 20 kilowatts of power in Class C telegraph service. This new tube is designed especially for operation with full input at frequencies as high as 25 megacycles. Operation with reduced input is permissible up to 100 megacycles.

The 889-R is a three-electrode tube with fin-type radiator for air-cooling the plate. The maximum plate dissipation in Class C telegraph service is 5 kilowatts. This new tube is intended for use with full input at frequencies as high as 25 megacycles. Operation with reduced input is permissible up to 100 megacycles.

The VR75/30 voltage regulator tube is of the cold-cathode, glow-discharge type. It has an operating voltage of approximately 75 volts and a maximum operating current of 30 milliamperes.

### TUBULAR CAPACITORS

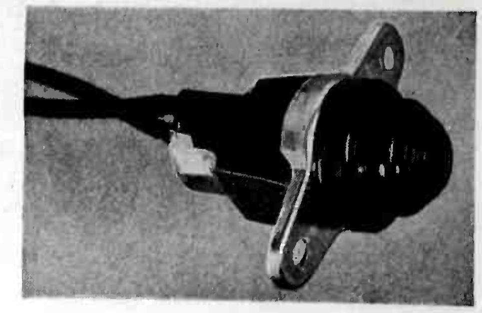
A new line of dry electrolytic tubular capacitors has just been announced by Cornell-Dubilier. This new Type BRH capacitor is available in capacities ranging up to 500 mfd at 25 volts, 1000 mfd at 15 volts, and 2000 mfd at 6 volts. The BRH capacitors are similar in physical appear-



ance to the standard Type BR. They are enclosed in vented aluminum containers over which a varnished cardboard tube is drawn for protection. Dimensions vary from ⅝" x 1 1/16" for the smaller sizes, to 1" x 2½" for the higher capacities. A choice of either pigtail or lug terminals is provided. Cornell-Dubilier Electric Corp., South Plainfield, New Jersey.

### PILOT LIGHT ASSEMBLY

A jewel pilot light assembly, developed by Alden Products Co., Brockton, Mass., fastens to the panel with either eyelets, rivets or screws through two holes in the mounting plate. The indicating jewel is made from a translucent thermo-setting



material which is available in a large variety of colors. It is also available with a U-shaped bracket on the back of which mounts the lamp detachable from the mounting plate if preferred.

### VARITRANS

New design changes have been made in the UTC Varitran units to increase their ruggedness and improve reliability. In addition to glass insulated wire throughout all sizes, multiple contact units now employ

ballast coils for uniform contact loading. The Varitran units are available for 115 or 230-volt service with respective output voltages of 0-130 and 0-260 volts. United Transformer Corp., 150 Varick Street, New York, N. Y.

### FLASHLIGHT BULB EXTENSION

Frequently the engineer has need for a closely spotted light in hard-to-get-at places. Wiring or other impediments prevent location of a light or flashlight right at the point where illumination is needed. To take care of this need a novel flashlight bulb extension has been developed. It extends the light through any maze of wires, behind or around equipment and brings the bulb right to the point desired. Made in

lengths from 6" to 36"; the extension has a plug which will screw into any flashlight with the bulb in the opposite socket. Being bendable, it can go down into intricate mechanisms. The device is made by the Sierra Aircraft Company, Sierra Madre, California.

### HAMMERED METAL EFFECTS

Finishes resembling hammered silver, copper, bronze, and other ornamental metals can be applied to products made of any kind of metal and also bakelite molded plastics by the use of a new type of finish, known as "Hammertone," which was recently developed by Maas & Waldstein Company, 438 Riverside Ave., Newark, N. J.

In finishing a surface with Hammertone, a base coat of the desired color is first sprayed on. This is followed immediately by a spatter coat of Hammertone liquid, which produces the hammered pattern in the base coat. The product is then baked at medium heat for an hour, which produces a lustrous metallic finish that is tough and durable.

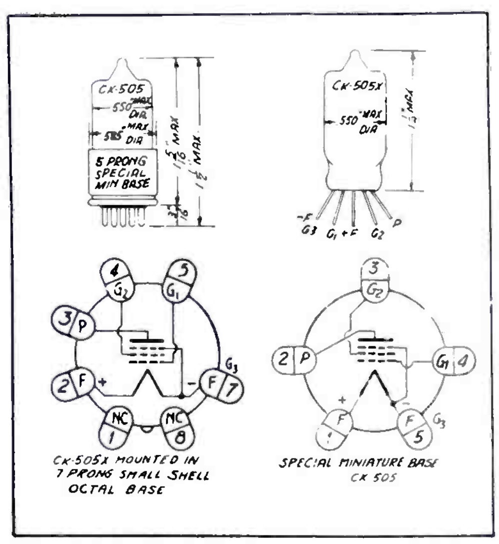
### DUREZ 8685

Durez Plastics & Chemicals, Inc., North Tonawanda, N. Y., announce a new arc-resistant phenolic molding compound called Durez 8685. This new material was formulated specifically to prevent tracking where there is combined electrical spark and rubbing action. It has very high resistance to carbonization under an arc. While slightly slower to cure in the mold, its electrical properties are materially increased by baking after molding, it is said. It will also hold its high dielectric strength when used at moderately high temperatures. Durez 8685 is somewhat more flexible in its final set than standard materials and does not crack under the expansion and contraction of metal inserts around which it may be easily molded.

### HEARING AID TUBES

Raytheon Production Corp., 55 Chapel St., Newton, Mass., have announced two new hearing aid tubes designated as types CK-505 and CK-505X. Both are filament type pentodes. These tubes require less

than half the filament power of their present hearing aid tubes and are a little shorter in size. A pair of these tubes in a resistance coupled amplifier has a voltage gain of about 225 at 30 volts of B battery. The total B drain for both tubes is 54 microamperes and the total A drain for the two tubes in series is 30 milliamperes at 1.25 volts. In addition to hearing aid ap-




plication these tubes may be used where extremely small size and low battery drain are the primary tube requirements.

### ATR BATTERY ELIMINATOR

The American Television & Radio Co., 300 E. Fourth St., St. Paul, Minn., have announced a B battery eliminator for battery portables, which operates from flashlight cells or storage batteries. The unit is designed to deliver 90 volts at 10 ma, operating from a 7.5-volt source such as 5 Type-D cells in series or a 3-cell storage battery. The total weight of the eliminator, including the flashlight cell container and cells, is approximately 3 lbs.

## Index to Advertisers





### WHEN YOU CHANGE YOUR ADDRESS

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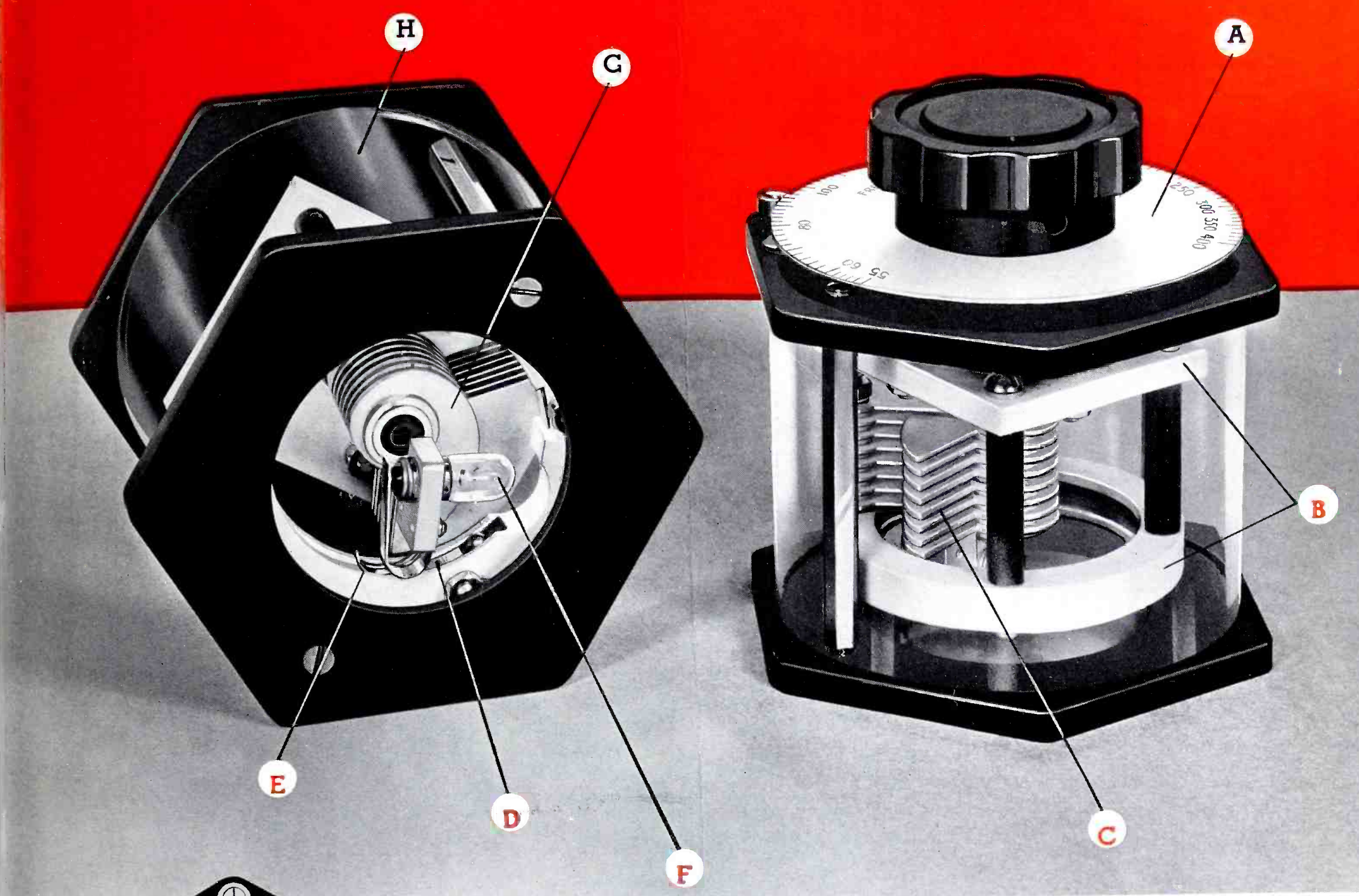
# *New* WAVEMETER for ULTRA-HIGH FREQUENCIES





Direct-reading Dial calibrated from 55 to 400 Mc, with open scale. Six-sided end plates for resting wavemeter horizontally in a number of positions.

## FROM 55 TO 400 MEGACYCLES

FOR SOME TIME there has been an increasing demand for a simple, small, accurate and portable wavemeter of the tuned-circuit absorption type for use at the very high frequencies. General Radio now announces the new Type 758-A Wavemeter to meet these needs.

Unique design involves the use of a variable single-turn inductance the value of which is changed continuously as the capacitance is changed. The condenser shaft is equipped with a positive sliding contact which bears on a single turn of flat strip, thus changing capacitance and inductance simultaneously and providing an enormous frequency range from a single coil.

When used with circuits generating some power a lamp indicator, coupled to the wavemeter circuit through a fixed loop, is used for resonance indication. For measuring very low power oscillators or receiving circuits, resonance is indicated by the usual absorption-type wavemeter means such as plate or grid milliammeter dip.

### DESIGN AND CONSTRUCTION FEATURES INCLUDE:

- **A** **EXTREMELY WIDE RANGE**—55 Mc to 400 Mc
- **B** **GOOD INSULATION**—Isolantite of small area
- **C** **NEW CONDENSER**—extra heavy, silver-plated brass plates
- **D** **SINGLE TURN INDUCTOR**—flat silver strip securely fixed in Isolantite ring
- **E** **VARIABLE INDUCTANCE**—2-finger silver spring making low-resistance contact with single turn
- **F** **LAMP RESONANCE INDICATOR**—small lamp coupled to wavemeter circuit through fixed loop
- **G** **STRAIGHT-FREQUENCY-LINE PLATES** — heavy construction
- **H** **TRANSPARENT PROTECTIVE CASE**—resonant illumination of lamp can be seen in any position of wavemeter

**This wavemeter, with an accuracy of 2%, is priced at only $28.00**

WRITE FOR BULLETIN 626 FOR INFORMATION

**GENERAL RADIO COMPANY** CAMBRIDGE MASSACHUSETTS

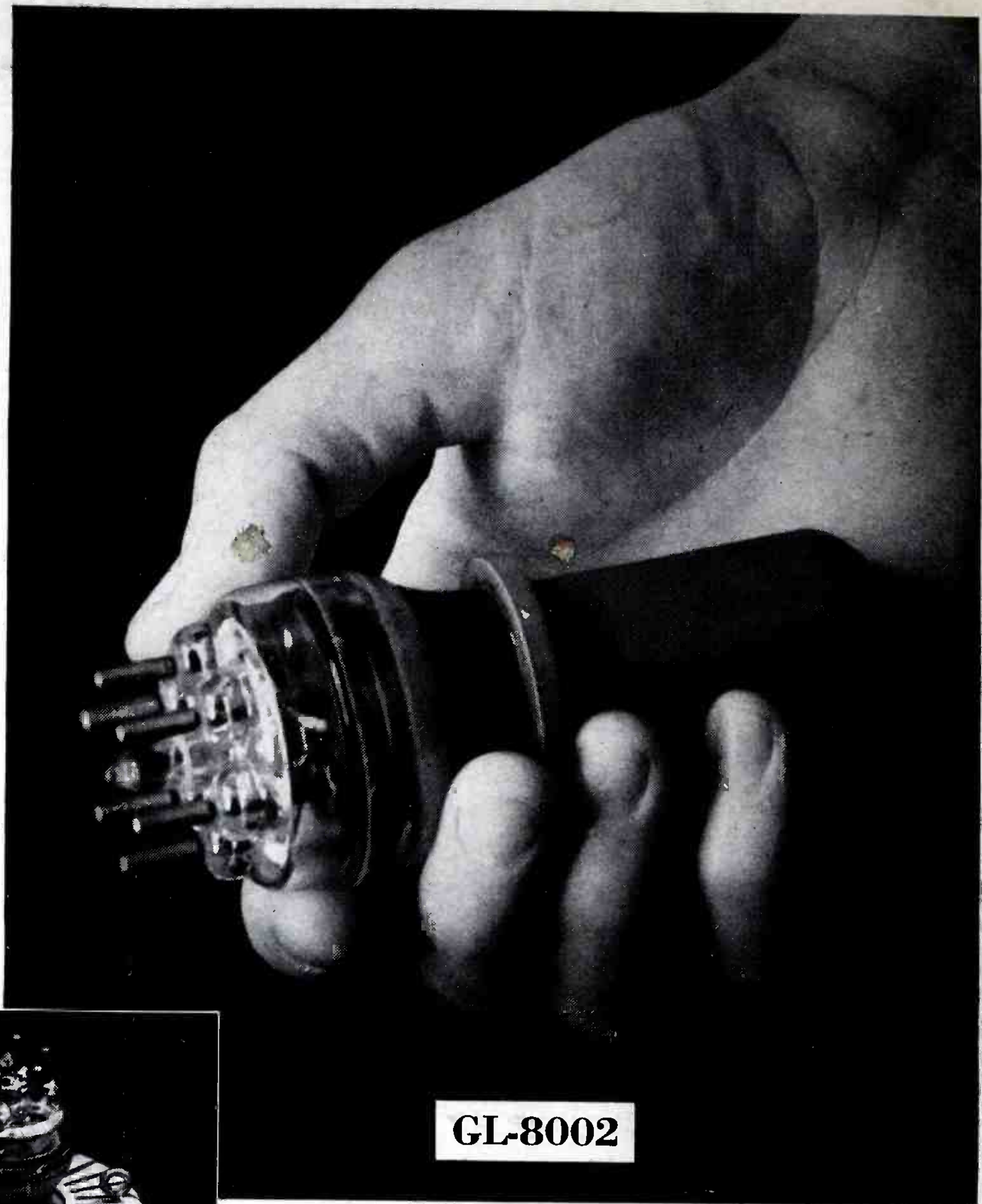
★

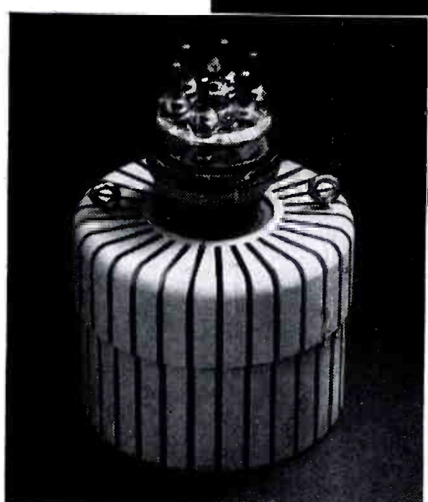
# HARDLY A HANDFUL

## But Its OUTPUT Is 1800 Watts up to 150 MC!

★

**And the air-cooled GL-8002-R puts out 1800 watts up to 120 MC.**



*GL-8002-R's radiator is milled from solid copper. Air passages are of uniform width for more efficient cooling. Total height, including radiator, is 5 3-8 inches; total diameter, 3 11-16 inches.*

**Characteristics—GL-8002 and GL-8002-R**

| | |
|---|---|
| Fil Volts . . . . . . . . . . . . . . 16 | Mutual Conductance . . . . . . 8850 mmhos |
| Fil Current . . . . . . . . 39 amp | **Class C Telegraph Rating** |
| Direct Interelectrode Capacitances, Approx.: | |
| Grid to plate . . . . . . 9 mmf | Plate Volts . . . . . . . . . . 3000 |
| Grid to filament . . . . 8 mmf | Plate Input, Watts . . . . 3000 |
| Plate to filament . . 0.5 mmf | Plate Dissipation, Watts 1200 |
| **GL-8002—$180** | **GL-8002-R—$250** |

BUILT to speed the conquest of the ultra-highs, General Electric's latest are marvels of vacuum-tube design.

These tiny triodes lick many basic U-H-F design problems almost singlehanded! And they're top performers at lower frequencies as well.

Small in size, their leads are short. But, in addition—their three grid leads make possible separation of excitation and neutralizing circuits, lowering common inductance, and thereby *greatly reducing neutralizing difficulties.*

Use of center-tapped filaments permits paralleling of filament leads for low inductance r-f feed. Reduced ratings permit operation up to 200 mc air-cooled, 300 mc water-cooled.

Two GL-8002-R's are used in the final amplifiers of both the G-E 3-kw frequency-modulation transmitter and the G-E 1-kw television sound transmitter.

GL-8002 and GL-8002-R are "sister" tubes of the GL-880, GL-889, and GL-889-R—all developed and introduced by General Electric, Schenectady, N. Y.

**GENERAL ELECTRIC**

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