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RCA REVIEW

FOREWORD

A LARGE number of requests have been received for TELEVISION, Volumes I (1936) and II (1937). These books have been out of print for some time and no copies are available. For this reason, summaries of all papers which appeared in these books were included in TELEVISION, Volume III (1938-1941) which was published in the spring of 1947. It was felt that their inclusion would serve as useful additional reference material on the earlier work in television and related fields. This has proved to be the case.

Original publication data was prepared as a means of readily identifying the source from which the above summaries were taken. Unfortunately, in the printing of TELEVISION, Volume III (Appendix), indication of **original publication data** for these papers was inadvertently omitted.

In order that each issued volume of TELEVISION, Volume III may be corrected for maximum usefulness, copies of the Original Publication Data are available as a gummed insert to the Appendix to TELEVISION, Volume III to follow page 460. They may be obtained by writing to:

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The Manager, RCA REVIEW

⚠ WARNING

IF THE BATTERY REMAINS INSTALLED IN THE VEHICLE:

1. DO NOT CONNECT THE CHARGER AC POWER PLUG TO THE ELECTRICAL OUTLET UNTIL ALL OTHER CONNECTIONS ARE MADE!
2. Place both the AC and DC power cords in the best position to avoid accidental damage by movable vehicle parts, i.e. hoods, doors, or moving engine parts (fan blades, belts, or pulleys).
3. Check the polarity of the battery posts. If the positive (pos, p, +) post is connected to the vehicle chassis, then the vehicle has a positive ground system. If the negative (neg, n, -) post is connected to the vehicle chassis, then the vehicle has a negative ground system. Negative ground systems are the most common.
4. For negative ground systems, connect the positive (red) alligator clip, or ring terminal to the positive battery post. Then connect the negative (black) alligator clip, or ring terminal to the vehicle chassis. Do not make the negative charger clip or ring connection to the carburetor, fuel lines, or thin, sheet metal parts. Make that connection to the engine block or a heavy gauge metal part of the frame.
5. For positive ground systems, connect the negative (black) alligator clip, or ring terminal to the negative battery post. Then connect the positive (red) alligator clip, or ring terminal to the vehicle chassis. Do not make the positive charger clip or ring connection to the carburetor, fuel lines, or thin, sheet metal parts. Make that connection to the engine block or a heavy gauge metal part of the frame.
6. Connect the AC power plug to the electrical outlet.

ADDITIONAL CHARGER INFORMATION

AUTOMATIC CHARGING AND BATTERY STATUS MONITORING: The Battery Tender® Junior and Plus Chargers are completely automatic and may be left

SMALL-SIGNAL ANALYSIS OF TRAVELING-WAVE TUBE*

By

CARL SHULMAN AND M. S. HEAGY

Research Department, RCA Laboratories Division, Princeton, N. J.

Summary—An analysis is made of an idealized traveling-wave tube consisting of a hollow cylindrical shell electron beam moving parallel to the axis of a helix in free space wound with vanishingly small, perfectly conducting wire. The beam may be inside or outside the helix. Particular emphasis is given to optimum design considerations. The method of attack is based on the small-signal theory of electromagnetic wave propagation along electron beams described by W. C. Hahn in 1939.

It is found that broad optimum design criteria do exist and are presented most conveniently in the form of curves. Generally speaking, it is shown that, for maximum gain, the helix should be as small as possible with the beam as close as possible to the helix. Furthermore, for a given wavelength and helix diameter an optimum pitch, and therefore voltage, exists.

To design for minimum noise factor the beam should be as far from the helix as possible. For a given wavelength and helix diameter no optimum pitch for minimum noise factor exists. The pitch should be as small as possible which means as low a beam voltage as possible.

IT IS THE PURPOSE of this paper to analyze the helical traveling-wave tube[1, 2] with particular emphasis on discovering what optimum design considerations exist, and to try to present design information in a simple form. The general method of attack is based on the small-signal theory of electromagnetic wave propagation along electron beams described by W. C. Hahn[3] in 1939, where the problem becomes one of solving Maxwell's equations inside and outside the beam subject to the boundary conditions at the beam edge, the helix wires, and an external shield if present.

The basic tube structure to be studied is shown in Figure 1, where a source of electrons, a long wire helix, and a collector are apparent. An electron beam is caused to move at constant velocity parallel to the axis of the helix, either inside or outside of the helix. An axial radio-

* Decimal Classification: R339.2.

[1] R. Kompfner, "The Traveling-Wave Valve," *Wireless World*, Vol. 52, p. 369, November, 1946; "The Traveling-Wave Tube as Amplifier at Microwaves", *Proc. I.R.E.*, Vol. 35, No. 2, pp. 124-127, February, 1947.

[2] J. R. Pierce, "Theory of the Beam-Type Traveling-Wave Tube," *Proc. I.R.E.*, Vol. 35, No. 2, pp. 111-123, February, 1947.

[3] W. C. Hahn, "Small Signal Theory of Velocity Modulated Beams," *G. E. Review*, Vol. 42, p. 258, 1939.

frequency electric field is applied to the helix at the cathode end. The system oscillates in all the modes consistent with the manner of excitation, and at least one of these modes (usually the simplest mode) corresponds to a freely propagated wave with an axial phase velocity considerably less than the velocity of light. The electron beam velocity is adjusted to coincide with the axial phase velocity of the desired mode. Under this condition there is a synchronous exchange of energy between the beam and the electromagnetic field such that, on the average, energy is extracted from the electron beam. This corresponds to an increase in amplitude of the electromagnetic wave on the helix. The analysis presented here involves a detailed study of this process.

A rigorous solution of the problem, which takes into account transverse motion of the electrons, potential distribution in the electron beam, diameter and conductivity of the helix wire, is not very fruitful so far as design information is concerned because of the prohibitive complexity involved. Hence, an approximate solution is sought in which the following idealizations are made: the electron beam is

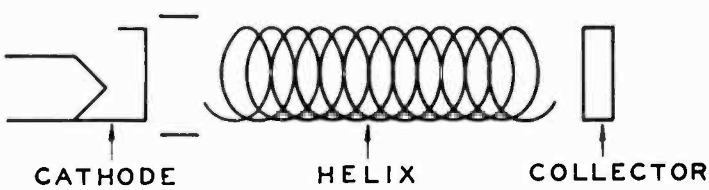

Fig. 1—Schematic diagram of traveling-wave tube.

assumed to be a hollow cylindrical shell of vanishing thickness; the electrons are constrained to move only in the axial direction (infinite axial magnetic field); the helix wire is perfectly conducting and of vanishingly small diameter; the electromagnetic field distribution in the surface containing the helix is the same as the field distribution for a helix of infinitesimal pitch; there is no external boundary, that is, the helix is in free space; and, of course, since this is a small-signal theory, the alternating-current components of all the dynamic variables are assumed to be small compared to their steady components, so that all products of alternating-current terms are neglected. This last idealization linearizes the equations and makes an analytic solution possible. The use of a hollow cylindrical beam not only removes the necessity of solving the inhomogeneous wave equation, but is suited for the solution with the beam outside the helix. Although these assumptions by no means accurately describe the situation, they lead to a simple solution which gives a reasonable first order check with experiment. No assumption is made concerning the diameter of the helix relative to the free space wavelength.

SOLUTION WITH BEAM INSIDE THE HELIX

Referring to Figure 2, which is a cross sectional view of the tube, and serves to define the coördinate system, let z be the axial dimension, a the radius of the cylindrical shell beam, b the radius of the helix. Maxwell's equations must be solved for free space in the three regions, $r < a$, $b > r > a$, and $r > b$ subject to the boundary conditions at $r = a$ and b, and the requirement that all field quantities be finite for all r, and zero for infinite r. Writing Maxwell's equations for free space, using Gaussian units, we have

$$\left.\begin{aligned} \operatorname{curl} \mathcal{E} &= -\frac{1}{c}\dot{\mathcal{H}} \\ \operatorname{curl} \mathcal{H} &= \frac{1}{c}\dot{\mathcal{E}} \end{aligned}\right\} \qquad (1)$$

where $\mathcal{E}$ is the electric field strength in statvolts/centimeter and $\mathcal{H}$ is the magnetic field strength in gauss.

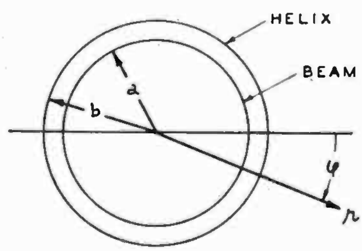

Fig. 2—Cross sectional view of traveling-wave tube defining the coördinate system used. Positive direction of z is into the page.

Since wave-like solutions only are of interest, solutions are sought where all field quantities are of the form

$$f(r, \varphi) \cdot \exp\left[i(\omega t - \Gamma z)\right] \qquad (2)$$

where $i = \sqrt{-1}$ $\omega =$ applied angular frequency

$\Gamma =$ propagation constant

On expansion of Equation (1) and insertion of the wave-like form, all the field quantities may be expressed in terms of the axial electric and magnetic fields as follows:[4]

[4] See for instance, John R. Carson, Sallie P. Meade, and S. A. Shelkunoff, "Hyper-Frequency Wave Guides—Mathematical Theory", *Bell Sys. Tech. Jour.*, Vol. XV, No. 2, pp. 310-333, April, 1936.

$$\left.\begin{aligned}
E_\varphi &= \frac{\Gamma}{i\eta^2}\left[\frac{1}{r}\frac{\partial E_z}{\partial \varphi} - \frac{\omega}{\Gamma c}\frac{\partial H_z}{\partial r}\right] \\
E_r &= \frac{\Gamma}{i\eta^2}\left[\frac{\omega}{\Gamma c}\frac{1}{r}\frac{\partial H_z}{\partial \varphi} + \frac{\partial E_z}{\partial r}\right] \\
H_\varphi &= \frac{\Gamma}{i\eta^2}\left[\frac{1}{r}\frac{\partial H_z}{\partial \varphi} + \frac{\omega}{\Gamma c}\frac{\partial E_z}{\partial r}\right] \\
H_r &= \frac{\Gamma}{i\eta^2}\left[\frac{\partial H_z}{\partial r} - \frac{\omega}{\Gamma c}\frac{1}{r}\frac{\partial E_z}{\partial \varphi}\right]
\end{aligned}\right\} \tag{3}$$

where $\mathcal{E} = E\ (r, \varphi) \exp\left[i\ (\omega t - \Gamma z)\right]$

$\mathcal{H} = H\ (r, \varphi) \exp\left[i\ (\omega t - \Gamma z)\right]$ $\qquad \eta^2 = \dfrac{\omega^2}{c^2} - \Gamma^2$

Both E_z and H_z are solutions of the wave equation

$$\frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial H_z}{\partial r}\right) + \frac{1}{r^2}\frac{\partial^2 H_z}{\partial \varphi^2} + \eta^2 H_z = 0 \tag{4}$$

and take the form

$$\left.\begin{aligned} H_z &= Z_m(\eta r)\ e^{im\varphi} \\ E_z &= \mathfrak{Z}_m(\eta r)\ e^{im\varphi} \end{aligned}\right\} \text{ where } m = 0, 1, 2, 3 \text{ - - - - -} \tag{5}$$

$Z_m(\eta r)$ and $\mathfrak{Z}_m(\eta r)$ are linear combinations of Bessel functions of the first and second kind.

It is recalled that, in the theory of continuous-wall waveguides, two independent waves can be set up; one for which $H_z = 0$, called the *TM* or "*E*" wave; and one for which $E_z = 0$, called the *TE* or "*H*" wave.[5] These two waves can be excited separately or simultaneously, their relative amplitudes depending entirely on the manner of excitation. This is not the case for the helical waveguide. It is found that both the *TE* and *TM* waves are needed simultaneously to satisfy the boundary conditions at the helix, so that separation into *TE* and *TM* waves on the helix is of no advantage.

Since the electrons have been constrained to move only in the z direction, the electron beam does not couple to H_z. It may be said that the electron beam is not coupled to the "*H*" wave. This means

[5] W. L. Barrow and W. W. Mieher, "Natural Oscillations of Electrical Cavity Resonators" *Proc. I.R.E.*, Vol. 28, No. 4, pp. 184-191, April, 1940.

that the beam does not represent a boundary for H_z. Therefore, H_z, or more specifically, $Z_m(\eta r)$ need be defined only in the regions $r \leqq b$ and $r \geqq b$. However, E_z does couple to the beam so that $\mathfrak{z}_m(\eta r)$ must be defined in the three regions, $r \leqq a$, $b \geqq r \geqq a$, and $r \geqq b$. Writing out $Z_m(\eta r)$ and $\mathfrak{z}_m(\eta r)$ in terms of $J_m(\eta r)$, the Bessel function of the first kind, and $N_m(\eta r)$, the Bessel function of the second kind, we have

$$\mathfrak{z}_m(\eta r) = \begin{cases} A\, J_m(\eta r) & r \leqq a \\ B\, J_m(\eta r) + C\, N_m(\eta r) & a \leqq r \leqq b \\ D\, H_m^{(1)}(\eta r) & b \leqq r \end{cases} \tag{6}$$

and

$$Z_m(\eta r) = \begin{cases} F\, J_m(\eta r) & r \leqq b \\ G\, H_m^{(1)}(\eta r) & r \geqq b \end{cases} \tag{7}$$

where $H_m^{(1)}(\eta r)$ is the Hankel function of the first kind defined by

$$H_m^{(1)}(x) = J_m(x) + i\, N_m(x)\dagger$$

and A, B, C, D, F, and G are constants to be determined from the boundary conditions.

The boundary conditions needed at $r = a$ are, (1), tangential $\mathcal{E}$ is continuous through the boundary; and (2), the discontinuity in tangential $\mathcal{H}$ is equal to the surface current density normal to tangential $\mathcal{H}$. At the boundary $r = b$ in the surface containing the helix (1), the electric field is assumed to be normal to the helix wire and continuous through it; and (2), the magnetic field parallel to the helix wire is assumed to be continuous through the boundary. If I is the alternating current component of the convection current in the beam and is assumed to be of the form, $I = I_1 \exp i(\omega t - \Gamma z)$, the boundary conditions at $r = a$ become

$$E_z\,(a - o) = E_z\,(a + o) \tag{8a}$$

$$H_\varphi\,(a + o) - H_\varphi\,(a - o) = \frac{2I_1}{ac}$$

Referring to Figure 3 which shows a development of the helix and defines θ and p, the boundary conditions at $r = b$ become

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† Since the Hankel function of the first kind only is used in this analysis, the superscript is hereafter suppressed. The symbol for the Hankel function $H_m(\eta r)$ is distinguished from that for magnetic field components $H_z(r, \varphi)$ by noting the subscript.

$$\left.\begin{aligned}
&E_z(b-o) - E_\varphi(b-o)\tan\theta = 0\\
&E_z(b+o) - E_\varphi(b+o)\tan\theta = 0\\
&E_z(b-o)\tan\theta + E_\varphi(b-o) = E_z(b+o)\tan\theta + E_\varphi(b+o)\\
&H_z(b-o) - H_\varphi(b-o)\tan\theta = H_z(b+o) - H_\varphi(b+o)\tan\theta
\end{aligned}\right\} \quad (8b)$$

where $\tan\theta = \dfrac{2\pi b}{p}$.

Defining the quantities

$$k = 1 - \frac{2\pi m\Gamma}{\eta^2 p}\,;\; s = \frac{2\pi b}{p}\,\frac{\omega}{i\eta c} \qquad w = \left(\frac{2\pi b}{p}\right)^2 + \frac{2\pi m\Gamma}{\eta^2 p}$$

the boundary conditions at $r = a$ and b form the following array:

$$\left.\begin{array}{llllllll}
-AJ_m'(\eta a) & +CN_m'(\eta a)+ & BJ_m'(\eta a) & 0 & 0 & 0 & = \dfrac{2i\eta}{\omega a} I_1\\
-AJ_m(\eta a) & +CN_m(\eta a)+ & BJ_m(\eta a) & 0 & 0 & 0 & =0\\
0 & +CkN_m(\eta b)+ & BkJ_m(\eta b) & 0 & +FsJ_m'(\eta b) & 0 & =0\\
0 & 0 & 0 & +DkH_m(\eta b) & 0 & +GsH_m'(\eta b) & =0\\
0 & +CwN_m(\eta b)+ & BwJ_m(\eta b) & -DwH_m(\eta b) & -FsJ_m'(\eta b) & +GsH_m'(\eta b) & =0\\
0 & +CsN_m'(\eta b)+ & BsJ_m'(\eta b) & -DsH_m(\eta b) & -FkJ_m(\eta b) & +GkH_m(\eta b) & =0
\end{array}\right\} \quad (9)$$

All the constants may be evaluated in terms of I_1. Hence, $E_z(r=a)$ is known in terms of I_1 through Maxwell's equations.

In addition to the above relations, I_1 may be related to $E_z(r=a)$ through the force equation, and the equation of continuity which are written as follows:

$$\left.\begin{aligned}
&-\frac{e}{m}\mathcal{E}_z = \frac{dv}{dt} = \frac{\partial v}{\partial t} + v_0\frac{\partial v}{\partial z}\\
&\frac{\partial I}{\partial z} + \frac{\partial q}{\partial t} = 0
\end{aligned}\right\} \quad (10)$$

where v is the a-c component of the electron velocity
v_0 is the steady component of the electron velocity
q is the a-c component of the linear charge density.

Since this is a small signal analysis, we may write $I = \rho_0 v + \rho v_0$ (11)

where cross products of alternating-current terms are neglected and ρ_0 is the direct-current component of the linear charge density. Demanding that the alternating-current components of all the dynamic variables in the beam, v, q, vary as exp $[i(\omega t - \Gamma z)]$, Equations (10) lead to the relation

$$E_z = \frac{2i}{\beta_0}\frac{V_0}{I_0}(\beta_0 - \Gamma)^2 I_1 \tag{12}$$

where $I_0 = \rho_0 v_0$; $V_0 = \frac{1}{2}\frac{m}{e}v_0^2$; $\beta_0 = \frac{\omega}{v_0}$

Equation (11) represents a relation between the axial electric field at the beam and the beam current, which comes from electron dynamics, while Equations (9) represent this same relation which arises from field theory.

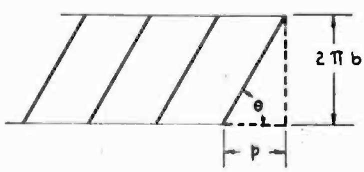

Fig. 3—The helix shown developed to define its dimensions.

Equation (11) combined with Equation (9) determines Γ from which the tube performance may be deduced. This process leads to hopelessly complicated expressions which can hardly be used for design information. The simplest approach is to assume that the introduction of the beam changes the propagation characteristics of the helical transmission system very little, so that the propagation constant of the system is the propagation constant of the undisturbed system (no beam) plus a small correction term, δ. Hence, one writes

$$\Gamma = \Gamma_0 + \delta; \quad (\delta << \Gamma_0) \tag{13}$$

where Γ_0 is the propagation constant with no beam, and δ is the correction term due to the introduction of the beam, and is evaluated in terms of the undisturbed system.

The first step, then, is to study the propagation characteristics of the undisturbed system. Γ_0 is determined from Equation (9) with I_1 and $C = 0$. The condition for the existence of a solution is that the determinant of the coefficients in the array enclosed by the dashed lines (Equation 9) be zero. That is,

$$\Delta\,(\eta_0 b) = \begin{vmatrix} k_0 J_m(\eta_0 b) & 0 & s_0 J_m{}'(\eta_0 b) & 0 \\ 0 & k_0 H_m(\eta_0 b) & 0 & s_0 H_m{}'(\eta_0 b) \\ w_0 J_m{}'(\eta_0 b) & -w_0 H_m(\eta_0 b) & -s_0 J_m{}'(\eta_0 b) & s_0 H_m{}'(\eta_0 b) \\ s_0 J_m{}'(\eta_0 b) & -s_0 H_m{}'(\eta_0 b) & -k_0 J_m(\eta_0 b) & k_0 H_m(\eta_0 b) \end{vmatrix} = 0 \tag{14}$$

where the subscript zero indicates no beam, so that

$$\eta_0{}^2 = \frac{\omega^2}{c^2} - \Gamma_0{}^2\,; \qquad s_0 = \frac{2\pi b}{p}\,\frac{\omega}{i\eta_0 c}$$

$$k_0 = 1 - \frac{2\pi m\Gamma_0}{\eta_0{}^2 p}\,; \quad w_0 = \left(\frac{2\pi b}{p}\right)^2 + \frac{2\pi m\Gamma_0{}^2}{\eta_0{}^2 p}.$$

Developing Equation (14), $\Delta\,(\eta_0 b) = 0$, gives

$$\frac{s_0{}^2}{k_0{}^2} = \frac{J_m(\eta_0 b)\; H_m(\eta_0 b)}{J_m{}'(\eta_0 b)\; H_m{}'(\eta_0 b)} \tag{15a}$$

as the condition on Γ_0.

Introducing the variable x such that $\eta_0 b = ix$, and defining the parameters $\nu = \dfrac{2\pi b}{p}$ and $\mu = \dfrac{2\pi b}{\lambda}\,\dfrac{2\pi b}{p}$, Equation (15a) becomes

$$\frac{1}{\mu}\left(x + m\sqrt{\nu^2 + \frac{\mu^2}{x^2}}\right) = \sqrt{\frac{J_m{}'(ix)\; H_m{}'(ix)}{J_m(ix)\; H_m(ix)}}. \tag{15b}$$

Solutions of Equation (15b) are sought which are associated with freely transmitting modes, that is, with Γ_0 real. Solutions for real x represent such freely transmitting modes, because real x implies real Γ_0, as seen from the relation

$$\Gamma_0 b = \sqrt{x^2 + \frac{\mu^2}{\nu^2}}.$$

Computations are, therefore, restricted here to real x.

Solutions of Equation (15b) are best obtained graphically. Figure 4 shows plots of the right-hand side of Equation (15b) vs. x for $m = 0, 1, 2$ These plots are included here for they are useful in estimating asymptotic behavior. Solutions for $m = 0$ are simple, for it is necessary only to draw straight lines of slope $1/\mu$ and note the intersection with the curve shown in Figure 4 for $m = 0$, thus obtaining x as a function of μ. For $m \neq 0$, solutions of (15b) are best obtained by inverting Equation (15b) and plotting the left and right-hand sides

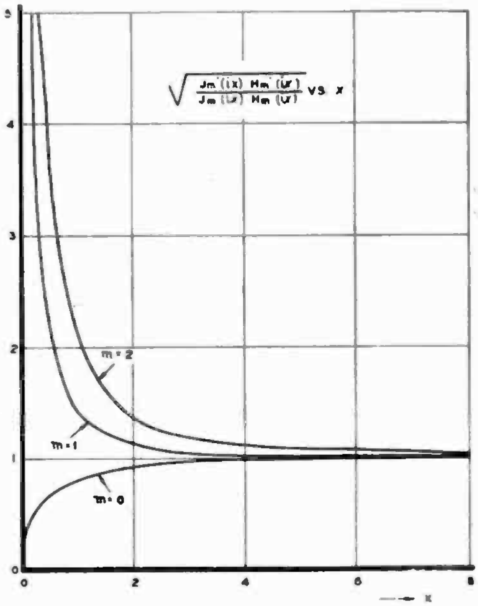

Fig. 4—The function $\sqrt{\dfrac{J_m{}'(ix)\ H_m{}'(ix)}{J_m\ (ix)\ H_m\ (ix)}}$ which is the right-hand side of Equation (15b) and is used to find graphical solutions of this equation.

of the inverted equation as functions of x for different values of μ and ν and noting the intersections.

Some of the solutions are summarized in Figure 5 which shows x vs. μ for $m = 0, 1, 2,$ and $\nu = 15$. It is seen that solutions exist for all values of μ indicating that there exists a set of freely transmitting modes which show no cut-off properties, in terms of which all helix fields might be described. This, of course, does not preclude the possibility of other solutions which may show cut-off properties.

Since Γ_0 is real for all μ, the propagation properties of the helix are best described in terms of a phase velocity which is defined as

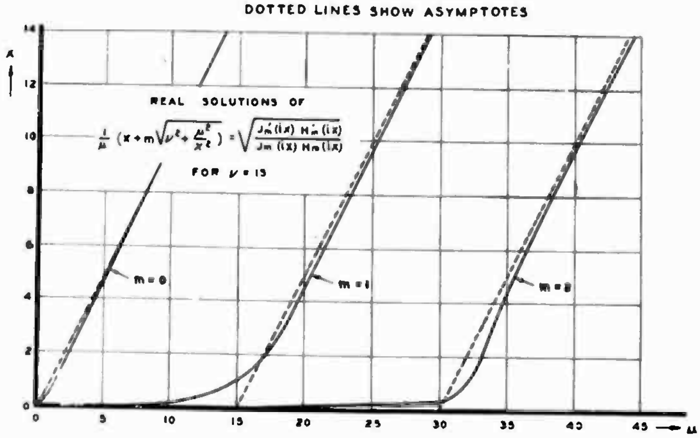

Fig. 5—Solutions of Equation (15b) which relates x, a function of the propagation constant of the undisturbed system, to μ which is a function of its geometry. ($\mu = 2\pi b/\lambda \cdot 2\pi b/p$)

$$v_p = \frac{\omega}{\Gamma_0} = \frac{c}{\sqrt{1 + \frac{x^2 \nu^2}{\mu^2}}} \tag{16}$$

$\frac{v_p}{c}$ is shown plotted in Figure 6 vs. μ for $\nu = 15$. Recalling that $\mu = \frac{2\pi b}{\lambda} \cdot \frac{2\pi b}{p}$ we see that Figure 6 is essentially a plot of phase velocity vs. frequency for a given helix. Since it is necessary that the phase velocity of the wave be near the beam velocity, it is clear from the curves that wide-band operation can be obtained for the higher modes herein described only for very high frequencies or very large helix diameters. It should be mentioned that, since it has been assumed that the field distribution in the surface containing the helix is the

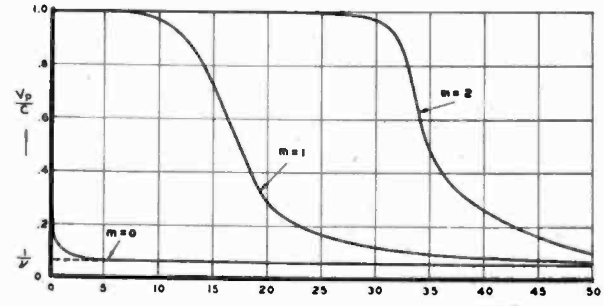

Fig. 6—Undisturbed axial phase velocity (v_p) of the electromagnetic waves along the helix as a function of μ for various modes with $\nu = 15$. ν is the ratio of pitch to circumference of the helix and μ is the product of ν and the number of wavelengths per helix turn.

same as that for infinitesimal pitch, the theory breaks down when the wavelength approaches the pitch of the helix, or say when λ is of the order of np, where n is between 5 and 10. Using this arbitrary criterion, we see that the theory applies only for

$$\mu \leq \frac{1}{n}\left(\frac{2\pi b}{p}\right)^2$$

Having determined Γ_0, the object is to obtain Γ in terms of Γ_0 with the assumption that the introduction of the beam changes the system very little, as previously mentioned. Combining Equations (9) and (12) we obtain a relation determining Γ which is

$$\frac{\pi\,\eta^2}{\omega}\left[N_m(\eta a) + \frac{B}{C} J_m(\eta a)\right] = \frac{2\ V_0}{\beta_0\ I_0}(\beta_0 - \Gamma)^2 \tag{17}$$

where
$$\frac{B}{C} = -i\left\{1 + \frac{\dfrac{H_m^2(\eta b)}{H_m'(\eta b)\ J_m'(\eta b)}}{\dfrac{s^2}{k^2} - \dfrac{H_m(\eta b)\ J_m(\eta b)}{H_m'(\eta b)\ J_m'(\eta b)}}\right\} \tag{18}$$

Equation (17)is rather unwieldly as it stands but is greatly simplified if the term in the Bessel function of the second kind can be neglected. This is indeed the case as the following indicates. It is known that C must vanish for zero beam current in a continuous manner, so that for very small beam currents one expects B/C to be large. More specifically, if the relation $\Gamma = \Gamma_0 + \delta$ is substituted into Equation (18) and the Bessel functions are expanded in the form

$$Z_m(\eta b) = Z_m(\eta_0 b) - \frac{\Gamma_0}{\eta_0}\delta b\, Z_m'(\eta_0 b)$$

where
$$\eta = \eta_0 - \frac{\Gamma_0}{\eta_0}\delta$$

one finds $\frac{B}{C}$ in terms of $\Gamma_0 + \delta$ to be $\frac{B}{C} = -\frac{\eta_0}{\Gamma_0 b}\frac{h_m(ix)}{\delta}$ (19)

where $h_m(ix) =$

$$\frac{-x\dfrac{H_m(ix)}{J_m(ix)}}{\dfrac{4}{1+m\nu\sqrt{1+\dfrac{\mu^2}{x^2\nu^2}}}+\dfrac{1}{1+\dfrac{\mu^2}{x^2\nu^2}}\left(1-\dfrac{2}{1+m\nu\sqrt{1+\dfrac{\mu^2}{x^2\nu^2}}}\right)}$$

$$+x\left[i\frac{J_m(ix)}{J_m'(ix)}+i\frac{H_m(ix)}{H_m'(ix)}\right]\left[1+\left(1+\frac{m^2}{x^2}\right)\left(\frac{\mu}{x+m\nu\sqrt{1+\dfrac{\mu^2}{x^2\nu^2}}}\right)^2\right] \tag{20}$$

Since δ is assumed to be very small with respect to Γ_0, $N_m(\eta a)$ of Equation (17) may be neglected with respect to $\dfrac{B}{C}J_m(\eta a)$ in a consistent manner, except possibly in the vicinity of $a=0$, where $-N_m(\eta a)$ increases without limit as a approaches zero. However, it can be shown that $N_m(\eta a)$ may be neglected even for very small a as follows. In the first place, only the case for $m=0$ need be considered in detail because with $m \neq 0$, $E_z(a=0)=0$ so that no tube could be operated with $a=0$. Therefore, it can be said immediately that for $m \neq 0$, the $N_m(\eta a)$ term may be neglected. Considering then the case for $m=0$, it is assumed that δ is a continuous function of I_0, more specifically, δ approaches zero continuously as the beam current goes to zero. Hence, one may say that, for small I_0, δ may be expressed by the first non-vanishing term of a Taylor expansion in I_0, say $\delta = K I_0^l$ where l is a positive integer. In any actual device with a shell beam, I_0 would be proportional to at least a if not a^2, so that B/C is proportional at worst to $1/a^l$ for small a. Now $N_0(\eta a) \rightarrow ln\, 1/a$ as a $\rightarrow 0$. It is clear then that $\dfrac{B}{C}J_0(\eta a) >> N_0(\eta a)$ as a $\rightarrow 0$. Consequently, $N_m(\eta a)$ is neglected for small a and Equation (17) is rewritten as follows:

$$\frac{-\pi\eta_0^3}{\omega\Gamma_0^b}\frac{h_m(ix)}{\delta}J_m^2(\eta_0 a)=\frac{2}{\beta_0}\frac{V_0}{I_0}(\beta_0-\Gamma)^2 \tag{21}$$

where the difference between $\eta^2 J_m^2(\eta a)$ and $\eta_0^2 J_m^2(\eta_0 a)$ has been neglected to be consistent with the approximations already made. Equation (21) is further simplified if one is restricted to solutions where the beam velocity is the same as the undisturbed velocity of the wave along the axis of the helix, that is, where $\beta_0 = \Gamma_0$. This leads to the cubic equation

$$\delta^3 = \frac{\lambda}{4b^4c}\,\frac{I_0}{V_0}\,x^3\,i\,h_m(ix)\,J_m^2\!\left(ix\frac{a}{b}\right); \qquad \left(\frac{a}{b} \leq 1\right) \tag{22}$$

where λ is the free space wavelength of the applied field. Equation (22) represents the final solution from which the gain of the tube may be deduced. Since $ih(ix)\,J_m^2\!\left(ix\frac{a}{b}\right)$ is a negative real function of x so that δ^3 is always negative and real, the three solutions of the cubic

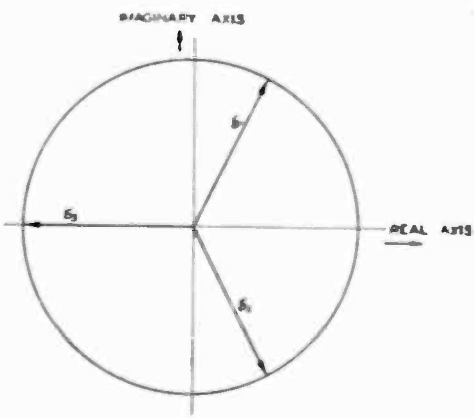

Fig. 7—Relation of the three solutions of Equation (22). δ is the increment to the propagation constant of the undisturbed system caused by introduction of the electron beam.

are related as shown in Figure 7, and correspond to the three forward waves described by J. R. Pierce.[2] (A fourth wave, moving in the direction opposite to motion of the electron beam, is not considered here.) The solution with a positive imaginary component of δ corresponds to the growing wave whose phase velocity is slightly less than the electron beam velocity.

If an external axial field is applied to the helix at $z = 0$, this field will in general excite many waves corresponding to modes which have an axial electric field component at the point of application. Those modes which are freely transmitting and whose phase velocities are close to the electric beam velocity will interact with the beam and show amplification. Each mode which interacts with the beam is somewhat perturbed and forms a complex wave which is most conveniently described in terms of three simple waves of the form

$$\exp\,[i\,(\omega t - \Gamma_0 z - \delta_j z)]\,;\;\; j = 1,2,3$$

corresponding to the three solutions of Equation (22). The general problem of calculating gain for a given mode is the determination of the amplitude of the growing component of the interacting mode in terms of the applied field. Under ordinary conditions only one mode interacts with the beam and all the rest are either not excited, not freely transmitted, or both.

Let an axial electric field be applied to the helix at $z = 0$ and assume that only one mode of transmission is excited and interacts with the beam. The propagation of the applied disturbance is to be described by the three component waves of the form

$$E_{zj} = A_j \exp\,[i\,(\omega t - \Gamma_0 z - \delta_j z)\,]$$

where E_{zj} is the axial electric field at the beam associated with one of the component waves. Associated with each of the three components is a current and velocity related to E_{zj} through Equations (10) and (12). From Equation (12), we have

$$I_{1j} = \frac{\beta_0}{2i}\,\frac{I_0}{V_0}\,\frac{E_{zj}}{\delta_j^{\,2}} \qquad (23)$$

and from Equations (10) the corresponding velocities are

$$v_j = i\,\frac{e}{m}\,\frac{1}{v_0}\,\frac{E_{zj}}{\delta_j} \qquad (24)$$

The three wave components add up to give the actual fields, currents and velocities existing anywhere along the tube. Assuming that at $z = 0$ the alternating-current and velocity in the beam are zero, then

$$\left.\begin{aligned} &A_1 + A_2 + A_3 = E_z \\ &\frac{A_1}{\delta_1} + \frac{A_2}{\delta_2} + \frac{A_3}{\delta_3} = 0 \\ &\frac{A_1}{\delta_1^{\,2}} + \frac{A_2}{\delta_2^{\,2}} + \frac{A_3}{\delta_3^{\,2}} = 0 \end{aligned}\right\} \qquad (25)$$

which are identical to Equations (28), (30), and (31) of Pierce's paper,[2] where E_z is the resultant axial electric field at the beam due to the field applied to the helix. The solution is

$$A_1 = A_2 = A_3 = E_z \frac{1}{\left(1 - \dfrac{\delta_3}{\delta_1}\right)\left(1 - \dfrac{\delta_2}{\delta_1}\right)} = \frac{E_z}{3} \tag{26}$$

which is Pierce's Equation (33)[2]. The voltage gain at some point along the tube is expressed as the ratio of the field at that point to the applied field. Since, at some distance along the tube, the growing component of the field is large compared to the others, the gain is immediately written as

$$\text{Gain} = \frac{E_{z1}}{E_z} = \frac{1}{3} \exp\left(\frac{\sqrt{3}}{2} |\delta|\, L\right) \tag{27}$$

where L is the distance along the axis of the helix between the input and output points, and $\dfrac{\sqrt{3}}{2}|\delta|$ is the imaginary part of δ_1.

A study of $|\delta|$ gives the gain characteristics of the tube. The design process involves adjustment of $|\delta|$ to obtain the best approximation to desired performance. The behavior of $|\delta|$ is most conveniently presented in terms of plots of $|\delta|$ vs. μ with ν, a/b and m as parameters, where it is recalled that

$$\mu = \frac{2\pi b}{\lambda} \cdot \frac{2\pi b}{p}, \text{ and } \nu = \frac{2\pi b}{p}.$$

V_0 can be eliminated from Equation (22) through the requirement that the beam velocity be the same as the undisturbed phase velocity of the helical wave. V_0 is then given by

$$V_0 = \frac{1}{2}\frac{m}{e} v_p^2 = \frac{1}{2\dfrac{e}{m}} \frac{c^2}{\left(1 + \dfrac{x^2 v^2}{\mu^2}\right)} \tag{28}$$

so that Equation (22) becomes

$$|\delta| = \left(\pi I_0 \frac{e}{m}\right)^{1/3} \frac{\lambda}{2\pi c b^2} \left| \left(1 + \frac{\mu^2}{x^2 \nu^2}\right)^{1/3} x^{5/3} h_m^{1/3}(ix)\, J_m^{2/3}\left(ix \frac{a}{b}\right) \right|$$

$$\left(\frac{a}{b} \leq 1\right) \quad (29)$$

Calculations have been made only for the case $m = 0$ which is the cylindrically symmetric mode and the simplest. This is the mode which is used in all the work published to date.[1,2]

$$y_0(ix) = \frac{\omega b^2}{\left(\pi I_0 \frac{e}{m}\right)^{1/3}} |\delta|_{m=0}$$

is shown plotted in Figure 8 for constant I_0, λ, and b as a function of μ for various values of a/b. This is essentially a plot of gain vs. $1/p$. It is seen that for each value of a/b, there exists a particular value of μ which yields maximum gain. It is further clear that the gain increases as a/b approaches unity. Hence, to design for maximum gain it is necessary to place the beam as close to the helix wire as possible and to operate at a value of μ giving maximum $|\delta|$ for a given beam dimension. The optimum value of μ is not critical as can be seen from the curves. Furthermore, it appears from Equation (29), that for a given wavelength and beam current it is desirable to reduce b as much as possible at the same time reducing p in such a way that

$$\frac{b}{p} \cdot \frac{b}{\lambda} = \text{constant.}$$

It is rather difficult to apply these conclusions to the case of a solid-beam traveling-wave tube. One might argue that, in an actual solid-beam tube, the potential distribution in the beam is such that each elementary cylindrical shell section of the beam is moving with a different velocity so that only one element is in synchronism with the traveling wave along the helix. Furthermore, if the helix voltage is adjusted for maximum gain presumably the outer section of the beam is in synchronism with the wave on the helix, for it is the outer electrons that give maximum gain. This idea makes even more sense if one notes that the outer electrons introduce a radio-frequency shielding effect which tends to reduce the coupling between the helix and the electrons inside the beam. Hence, one might guess that in a solid-beam tube, when the tube adjusted for maximum gain, one really establishes synchronism between the helix wave and that cylindrical shell of electrons which is closest to the helix and which makes a nearly complete transit of the tube. This effect will depend on the beam current, magnetic field, focusing, and electron optics in general. These considerations lead one to guess, therefore, that the design considerations described above for the shell beam case also apply to solid beam tubes

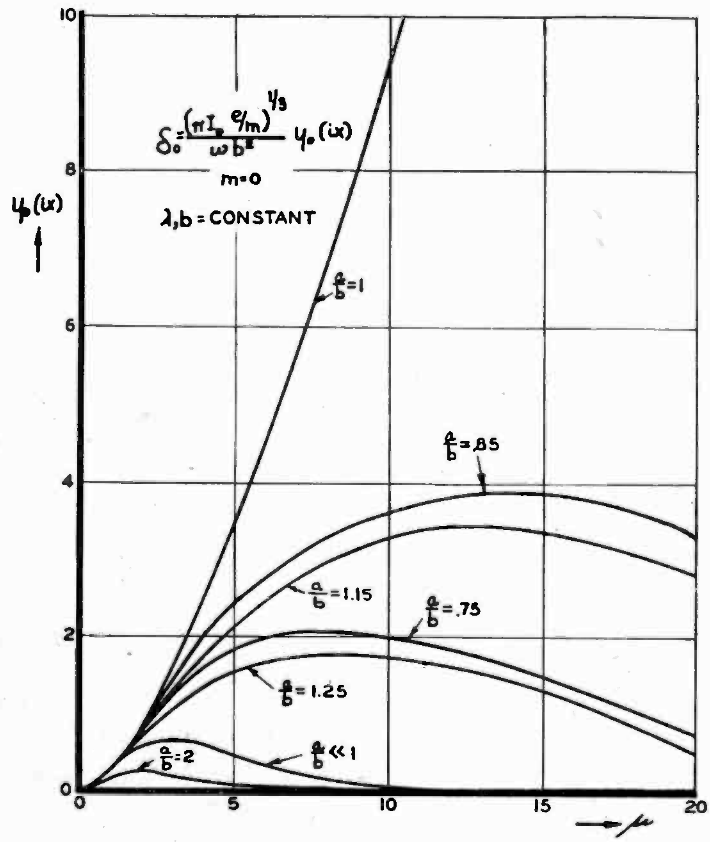

Fig. 8—Y_o as a function of μ with $\frac{a}{b}$ (ratio of beam radius to helix radius) as a parameter. Y_o is proportional to gain per unit length in *db*. ($\mu = 2\pi b/\lambda \cdot 2\pi b/p$)

where a/b is less than, say 0.75. For solid beams which fill the helix, that is $a/b = 1$, we would guess that the effective a/b is between 0.75 and 0.85, so that for maximum gain in this case μ should be between 5 and 15.

For values of $\mu \geqq 5$, asymptotic expansions for the Bessel functions may be used with sufficient accuracy so that Equation (29) becomes

$$|\delta|_{m=0} = \left(\frac{\pi}{2} I_0 \frac{e}{m}\right)^{1/3} \frac{\lambda}{2\pi b^2 c} \mu^{5/3} e^{-\frac{2\mu}{3}} J_0^{2/3}\left(i\mu\frac{a}{b}\right) \qquad (30)$$

$$(\mu \geqq 5) \qquad \left(\frac{a}{b} \leqq 1\right)$$

which for values of $\frac{a}{b} \geqq 0.75$ becomes

$$|\delta|_{m=0} = \left(\frac{I_0}{4}\frac{e}{m}\right)^{1/3} \frac{\lambda}{2\pi b^2 c} \frac{1}{\left(\frac{a}{b}\right)^{1/3}} \mu^{4/3} e^{-\frac{2}{3}\mu\left(1-\frac{a}{b}\right)} \qquad (31)$$

$$\mu \geqq 5$$

$$1 \geqq \frac{a}{b} \geqq 0.75$$

If positive ion space charge neutralization were in effect, so that the whole electron beam were moving at uniform velocity, the foregoing arguments would still be applicable because it is the outer electrons which are most effective in giving large gain and the μ would be chosen to optimize an outer shell.

It can be seen from Figure 8 that values of μ greater than 5 are indicated. From Figure 5 we see that for $m = 0$, $x \simeq \mu$ for values of $\mu \geqq 5$ so that one may write for V_0

$$V_0 = \frac{1}{2\frac{e}{m}} \frac{c^2}{\nu^2} = \left(\frac{\lambda}{2\pi b}\right)^2 \frac{1}{2\frac{e}{m}} \frac{c^2}{\mu^2} \qquad (28a)$$

where it is assumed that $\nu^2 >> 1$. Hence it is clear that optimum beam voltages exist which give maximum gain for a given beam current, wavelength and helix diameter. The closer the beam approaches the helix, the higher the gain becomes and the lower the optimum beam voltage becomes. This neglects direct-current space-charge effects which would, of course, limit the minimum beam voltage for a given beam current.

Having determined the conditions for best gain, it is of importance to repeat the whole process from the point of view of optimum noise factor which is here defined as the ratio of the total noise power to the antenna noise power. The noise factor is estimated in the manner outlined by J. R. Pierce[2] as follows. The noise power introduced by the tube itself is assumed to come from the shot noise in the beam. The amplitude of the growing component of the traveling wave due to the shot noise excitation is related to the shot noise current through relations of the type shown in Equation (25). The boundary conditions assumed at $z = 0$ are that the axial electric field due to shot effect is zero, the alternating-current component of beam velocity due to shot effect is zero, and the root-mean-square convection current in the beam due to shot effect is given by

$$\overline{I_n^2} = \alpha^2\, 2eI_0 \,\Delta f \tag{32}$$

where α^2 is a factor depending on space charge and interaction effects. Hence, at $z = 0$ we get

$$\left.\begin{aligned} &A_{n1} + A_{n2} + A_{n3} = 0 \\ &\frac{A_{n1}}{\delta_1} + \frac{A_{n2}}{\delta_2} + \frac{A_{n3}}{\delta_3} = 0 \\ &\frac{A_{n1}}{\delta_1^2} + \frac{A_{n2}}{\delta_2^2} + \frac{A_{n3}}{\delta_3^2} = \sqrt{2}\,\frac{2i}{\beta_0}\,\frac{V_0}{I_0}\sqrt{\overline{I_n^2}} \end{aligned}\right\} \tag{33}$$

(The factor $\sqrt{2}$ occurs because the A's are peak values)

where the axial electric field at the beam associated with each component waves due to shot noise excitation is written

$$E_{znj} = A_{nj} \exp\left[i\left(\omega t - \Gamma_0 z - \delta_j z\right)\right];\ j = 1, 2, 3.$$

Solution of Equation (33) gives

$$A_{n1} = \frac{2\sqrt{2i}}{\beta_0}\,\frac{V_0}{I_0}\sqrt{\overline{I_n^2}}\,\frac{\delta_2\delta_1}{\left(1-\dfrac{\delta_2}{\delta_1}\right)\left(1-\dfrac{\delta_3}{\delta_1}\right)} \tag{34}$$

$$= \frac{2\sqrt{2i}}{\beta_0}\,\frac{V_0}{I_0}\sqrt{\overline{I_n^2}}\,\frac{\delta_1\delta_2}{3} \tag{34a}$$

so

$$\overline{A_{n1}^2} = \frac{8}{\beta_0^2}\left(\frac{V_0}{I_0}\right)^2 \frac{|\,\delta\,|^4}{9}\,\alpha^2 eI_0\,\Delta f \tag{35}$$

where A_{n_1} is the amplitude of the growing component due to shot excitation.

To calculate the amplitude of the growing wave component due to thermal noise excitation from the antenna, the available noise power from the antenna is equated to the total power crossing the plane normal to the helix at $z = 0$. This total power is obtained by integrating the axial component of the Poynting vector over the entire r, φ plane and can be written in terms of the amplitude of the axial electric field at any arbitrary radius, say at the beam. The ratio of the square of the axial field at the beam to the total power transmitted along the helix is essentially a property of the system geometry and has the dimensions of an impedance density.

The average power transmitted per unit area is

$$\overline{S}_z = \frac{c}{8\pi} (E_r H_\varphi^* - E_\varphi H_r^*) \tag{36}$$

and the average power transmitted is

$$\overline{P} = \int_0^{2\pi} \int_0^{\infty} \overline{S}_z \, r d\varphi dr \tag{37}$$

Since it is assumed that the introduction of the beam changes the field distribution very little, it is sufficient for all practical purposes to use the unperturbed field distribution for the calculation of $\overline{P}$. An impedance-like factor $R_m(a)$ is defined through the equation

$$\overline{P} = \frac{E_z^2(a)}{2\, R_m(a)} \tag{38}$$

where the "a" refers to the radius at which R_m is defined. For instance, $R_m(a)$ is the impedance density at the beam, while $R_m(b)$ is the impedance density at the helix and

$$\frac{R_m(a)}{R_m(b)} = \frac{J_m^2(\eta_0 a)}{J_m^2(\eta_0 b)} = \frac{J_m^2\left(ix\dfrac{a}{b}\right)}{J_m^2(ix)}$$

is an impedance transformation ratio.

Carrying out the indicated integration gives

$$R_m(a) = \frac{2\, J_m^2\left(ix\dfrac{a}{b}\right)}{c\, p_m(ix)}; \qquad \left(\frac{a}{b} \leq 1\right) \tag{39}$$

where

$$(-1)^m\, p_m(ix) = \frac{\Gamma_0}{x^2}\,\frac{\pi b^2}{\lambda}\left\{\left[1 - \frac{k_0^2}{s_0^2}\,\frac{J_m^2(ix)}{J_m'^2(ix)}\right]\right.$$

$$\left[J_{m+1}^2(ix) - J_m(ix)\, J_{m+2}(ix)\right] + \left[\frac{k_0^2}{s_0^2}\,\frac{J_m^2(ix)}{H_m'^2(ix)} - \frac{J_m^2(ix)}{H_m^2(ix)}\right]$$

$$\left.\left[H_{m+1}^2(ix) - H_m(ix)\, H_{m+2}(ix)\right]\right\} + im\,\frac{k_0\,\Gamma_0^2}{s_0 x^4}\left(1 + \frac{\omega^2}{\Gamma_0^2 c^2}\right)$$

$$\left\{J_m^2(ix)\,\frac{H_m(ix)}{H_m'(ix)} - \frac{J_m(ix)}{J_m'(ix)}\left[J_m^2(ix) - J_m^2(o)\right]\right\}$$

$$+\, 2m\Gamma_0\,\frac{\pi}{\lambda x^4}\left[1 - \frac{k_0^2}{s_0^2}\,\frac{J_m^2(ix)}{J_m'^2(ix)}\right]\left\{\left[J_m^2(ix) - J_m^2(o)\right]\right.$$

$$\left. + H_m^2(ix)\left[\frac{k_0^2}{s_0^2}\,\frac{J_m^2(ix)}{H_m'^2(ix)} - \frac{J_m^2(ix)}{H_m^2(ix)}\right]\right\} \tag{40}$$

Having related the power transmitted through the helix to the axial electric field at the helix, it is now possible to estimate the noise factor of the tube. The thermal noise power delivered to the helix waveguide by a perfectly matched antenna is

$$\overline{P_t} = kT \triangle f \tag{41}$$

where k is Boltzmann's constant, T is the absolute temperature of the antenna, $\triangle f$ is the effective noise bandwidth of the system.

Equating Equations (38) and (41) and inserting (39), we get the axial electric field at the beam due to antenna noise excitation which is written

$$\overline{E_{zt}^2} = \frac{4kt \, \Delta f}{c} \; \frac{J_m^2 \left(ix \dfrac{a}{b}\right)}{p_m(ix)}; \qquad \left(\frac{a}{b} \leq 1\right) \tag{42}$$

The three component waves associated with the antenna noise excitation are of the form

$$E_{zt_j} = A_{t_j} \exp \left[i \left(\omega t - \Gamma_0 z - \delta_j z \right) \right]; \; j = 1, 2, 3$$

and we get $A_{t_1} = \dfrac{E_{zt}}{3}$ for the growing component just as in the case for the applied axial field.

The noise factor* becomes $F = \dfrac{\overline{A_{t_1}^2} + \overline{A_{n_1}^2}}{\overline{A_{t_1}^2}}$ (43)

Substituting Equation (35), (22), and (42) into Equation (43), gives

$$F = 1 + \frac{\alpha^2 \, ec \, \pi^{4/3} \, I_0^{1/3}}{2kT \left(\dfrac{e}{m}\right)^{4/3}} \; \frac{1}{b^2} \; \frac{x^4}{\left(x^2 + \dfrac{\mu^2}{\nu^2}\right)^{5/3}} \left| p_m(ix) \, J_m^{2/3} \left(ix \frac{a}{b}\right) h_m^{4/3}(ix) \right|$$

$$\left(\frac{a}{b} \leq 1\right) \tag{44}$$

where the voltage relation, Equation (28), has been inserted.

Optimum conditions must be obtained from a study of Equation (44) which is highly complicated. However, for $m = 0$, the relations become rather simple and have been calculated. For $m = 0$, one defines a function $Q_0 \left(ix, \dfrac{a}{b}\right)$ such that

$$(F)_{m=0} = 1 + \frac{\alpha^2 \, ec\lambda}{8kTb} \; \frac{(\pi I_0)^{1/3}}{\left(\dfrac{e}{m}\right)^{2/3}} \, Q_0 \left(ix, \frac{a}{b}\right); \qquad \left(\frac{a}{b} \leq 1\right) \tag{45}$$

---

* The definition used for noise factor is that of the tube by itself, i.e., no noise is assumed for the output system. The distinction from an over-all noise factor is of particular importance because, in the case of a match of the output to a room temperature resistance load, with a loss-free helix, the output load noise is freely transmitted back to the input thereby giving a minimum over-all noise factor of 2 even when no tube noise is present.

Figure 9 shows $Q_0(ix, \frac{a}{b})$ plotted vs. μ for various values of a/b with I_0, λ, and b held constant. It is clear from the curves that no finite value of μ exists which gives a minimum F. F decreases continuously with increasing μ, and the smaller a/b, the more rapidly F decreases. It is also noted that the beam current appears in the numerator so that the noise factor improves as the beam current is reduced.

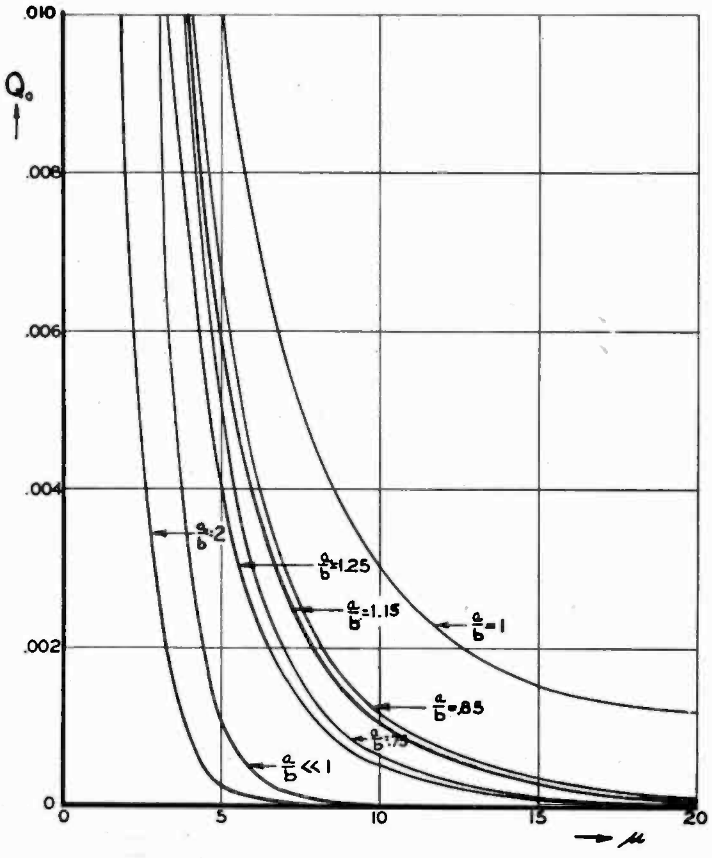

Fig. 9—Q_o as a function of μ with $\frac{a}{b}$ (ratio of beam radius to helix radius) as a parameter. Q_o is proportional to the noise factor minus one. $(\mu = 2\pi b/\lambda \cdot 2\pi b/p)$

Since large values of μ are indicated, asymptotic expressions for the Bessel functions may be used so that the noise factor may be written

$$(F)_{m=0} = 1 + \frac{\alpha^2 e c \lambda}{32kTb} \frac{(\frac{1}{2} I_0)^{1/3}}{\left(\pi \frac{e}{m}\right)^{2/3}} \frac{e^{-\frac{2}{3}\mu}}{\mu^{4/3}} J_0^{2/3}\left(i\mu\frac{a}{b}\right) \qquad (46)$$

$$\mu \geq 5 \qquad \left(\frac{a}{b} \leq 1\right)$$

As the noise factor is reduced by decreasing I_0 and a/b while at the same time increasing μ, one soon reaches a point where the noise factor of the stage following the traveling-wave tube begins to be significant, for the gain of the device falls off as I_0 and a/b decrease. Hence, for a given helix and following stage, there will be a definite beam current for minimum over-all noise factor. It is clear then that, if one increases the length of the helix while reducing beam current, the gain can be maintained and over-all noise factor reduced. Hence, for inside beam tubes, design for low-noise factor would tend toward very long helices operating with very small beam currents near the center of the helix.

Since for low-noise factors large values of μ are indicated as shown in Figure 9, the beam voltage may again be expressed as in Equation (28a) so that one may say that the lower the beam voltage the lower noise factor for all values of a/b.

Solution with Beam Outside the Helix

The solution for the beam outside the helix follows in exactly the same manner outlined for the solution with the beam inside. The only difference occurs in the application of the boundary conditions. It is readily shown that the expression for δ with the beam outside is given by

$$\delta^3 = \frac{\lambda}{4b^4c} \frac{I_0}{V_0} x^3 h_m(ix) J_m{}^2(ix) \frac{H_m{}^2 (ix \frac{a}{b})}{H_m{}^2(ix)} \qquad (47)$$

$$\left(\frac{a}{b} \geqq 1\right)$$

which we may write more simply as follows:

$$\delta \left(\frac{a}{b} \geqq 1\right) = \left[\frac{H_m (ix \frac{a}{b})}{H_m(ix)}\right]^{2/3} \cdot \delta \left(\frac{a}{b} = 1\right) \qquad (48)$$

Figure 8 shows plots of $\delta \left(\frac{a}{b} \geqq 1\right)$ vs. μ for various values of $\frac{a}{b}$ where the same sort of behavior as was seen for $\frac{a}{b} \leqq 1$ is apparent.

To design for maximum gain, it is again necessary to place the beam as close to the helix as possible and to operate at a value of μ which gives a maximum value of $|\delta|$. The limiting factors determining maximum gain, which are based on how close one can place the beam to the helix, are about the same just inside and outside the helix, so that for maximum gain, μ should be between 5 and 15 as was estimated for the inside beam case.

The expression for noise factor is given by the relation

$$F = 1 + \frac{\alpha^2 ec\lambda}{8kTb} \frac{(\pi I_0)^{1/3}}{\left(\dfrac{e}{m}\right)^{1/3}} Q_0 \left(ix, \frac{a}{b} \geqq 1\right) \tag{49}$$

where

$$Q_0 \left(ix, \frac{a}{b} \geqq 1\right) = Q_0 (ix, 1) \left\{ \frac{H_m \left(ix \dfrac{a}{b}\right)}{H_m (ix)} \right\}^{2/3} \tag{50}$$

$Q_0 \left(ix, \frac{a}{b} \geqq 1\right)$ is shown plotted in Figure 9 vs. μ for various values of $\frac{a}{b}$. Again it is clear that no finite value μ exists which gives a minimum value of F. F decreases continuously with μ, and the larger a/b, the more rapidly F decreases. The asymptotic expression for large μ, i.e. $(\mu \geqq 5)$, gives

$$(F)_{m=0} = 1 + \frac{\alpha^2 ec\lambda}{32kTb} \frac{I_0^{1/3}}{\left(2\dfrac{e}{m}\right)^{2/3}} \frac{1}{\pi\left(\dfrac{a}{b}\right)^{1/3}} \frac{e^{-\frac{2}{3}\left(\frac{a}{b}-1\right)}}{\mu^{5/3}} \tag{51}$$

$$\frac{a}{b} \geqq 1$$

Very low-noise factors may be obtained with large value of a/b, of course, subject to the limitation introduced by the noise factor of the following stage. However, it should be emphasized that these results apply only to the case with no outer shield, and must be applied judiciously to cases where a shield is present. The range of applicability is best estimated by referring to field plots of E_z vs. r/b as shown in Figure 10. The introduction of a cylindrical shield at a value of r/b where the axial field is small would change the field distribution little, so that in such a case one might expect Equations (48) and (50) to hold as long as the beam is not close to the shield. From Figure 10 it is clear that the larger μ, the larger the range of applicability of Equations (48) and (50). Since introduction of a shield causes the field to fall off more rapidly, one can certainly say that as a/b increases, both noise factor and gain per unit length drop off more rapidly with a shield than without a shield.

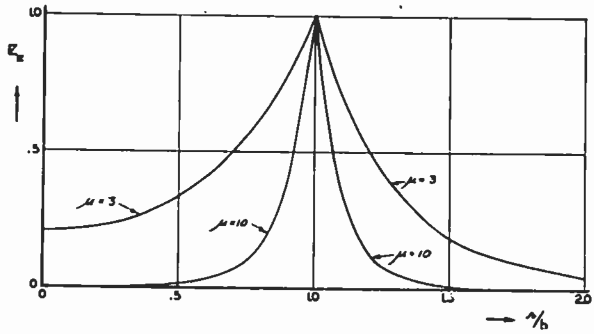

Fig. 10—Axial electric field in the undisturbed system as a function of the ratio of distance from the axis to radius of the helix for $\mu = 3$ and 10. ($\mu = 2\pi b/\lambda \cdot 2\pi b/p$)

CONCLUSIONS

It is found that for a given beam current and wavelength, to design for maximum gain, one should use as small a helix diameter as possible with the beam as close to the helix as possible either inside or out. Under these conditions it is estimated that the parameter $\mu = \frac{2\pi b}{p} \cdot \frac{2\mu b}{\lambda}$ should be between 5 and 15 with the optimum beam voltage being given by the relation

$$V_0 = \left(\frac{\lambda}{2\pi b}\right)^2 \frac{1}{2\frac{e}{m}} \frac{c^2}{\mu^2}$$

To design for minimum noise factor, no finite optimum value of μ is found. μ should be as large as possible, which means as low a value of V_0 as possible. The beam should be as far from the helix as possible, that is, close to the center for an inside beam tube, and large beam diameters for outside beam tubes. For an outside beam tube with an external cylindrical shield, the beam should be placed close to the shield for low noise. Since the gain becomes small as the noise factor is reduced, the over-all helix length must be increased to maintain sufficient gain. Hence, design for low-noise factor would tend toward very long helices operating with small beam currents at the center of the helix or with large cylindrical shield beams outside the helix, or even possibly a line beam outside the helix.

Acknowledgment

The authors wish to express their gratitude to L. S. Nergaard, D. O. North and E. W. Herold for their coöperative discussions and suggestions in the preparation of this paper.

TELERAN*

Part II — First Experimental Installation†

BY

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NOTE: Part I of this paper entitled "TELERAN—Air Navigation and Traffic Control By Means of Television and Radar" appeared in *RCA REVIEW* in December, 1946.[1] The Summary of this first paper is reprinted herewith for reference purposes. A limited number of copies of the December 1946 issue is still available for those who desire a complete file on the TELERAN system. It is planned that Part III concerning experimental tests and operational results will appear in *RCA REVIEW* sometime in 1948.

Part I — Air Navigation and Traffic Control by Means of Television and Radar

(Reprinted from RCA REVIEW, December, 1946)

Summary*—Wartime development of radar techniques offer a new approach to the problem of improving air navigation and traffic control, two fields in which existing equipment is obsolescent. Since no one military equipment appears to be ideally suited for solving the many problems, a re-investigation of the requirements necessary to handle very heavy traffic is first made.*

After analyzing the requirements, a system is described which appears to fulfill these requirements and offers unique advantages. In this system (Teleran), aircraft position information is presented to ground observers and controllers on a series of plan position indicators. One indicator is used for each altitude layer. Information on the position of aircraft in a given altitude is superimposed on a map of the region covered by the ground radar; this, together with weather, traffic control, and other desired information, is transmitted by television to each aircraft in the region. Each cooperating aircraft is equipped with a transponder beacon which serves not only to reinforce the radar echo but also to provide an altitude-dependent reply which allows the ground station to differentiate among aircraft by altitude.

The application of this system to concrete problems of navigation and traffic control is also discussed. Included are the problems of enroute navigation, approach and landing procedures, automatic flight and landing,

* Decimal Classification: 629.132.5 x R583 x R537.

† Presented, Philadelphia Section, Institute of Radio Engineers, November 6, 1947.

collision prevention, personal identification and communication, and air traffic control.

* * *

There follows the second paper in this series: Part II—First Experimental Installation.

The Manager, RCA REVIEW

Part II — First Experimental Installation

Summary—A set of equipment designed to demonstrate the principles of Teleran as outlined in a previous paper is described.[1] *Liberal use of conventional circuits and practices has been made and only those new circuits designed specifically for Teleran are described here. Pictures of the airborne equipment are included. The paper covers in detail the choice of television parameters, synchronizing equipment, the television transmitter, altitude coding, cameras and mixing, the landing display, the airborne beacon, and the airborne television receiver.*

INTRODUCTION

An EARLIER paper[1] outlined the services of Teleran, an air navigation and traffic control system employing television and radar. The present paper describes a set of equipment providing those services, which equipment has been developed under contract No. W28-099ac-107 with the Watson Laboratories, Air Materiel Command, U. S. Air Force. The apparatus described here has been constructed in order to demonstrate the general principles of Teleran and does not provide all of the facilities which would be provided in a comprehensive installation. The automatic en route flight, automatic landing and "Command" features have been omitted, although the transmitter and airborne receiver portions of the present equipment have been designed in such a way that these features can be added at any time. Likewise only three altitude channels have been provided in the first installation. Although the method used for altitude separation is capable of providing more channels, three were chosen as the irreducible minimum necessary to demonstrate the principles involved.

The first experimental installation includes an airway search radar, an airport search radar, and a final approach radar together with the television equipment associated therewith. The airway radar is a modified AN/CPS-1, the airport radar another surveillance set, and the final

[1] D. H. Ewing and R. W. K. Smith, "TELERAN—Air Navigation and Traffic Control by Means of Television and Radar", *RCA REVIEW*, Vol. VII, No. 4, pp. 601-621, December, 1946. (Referred to hereafter as Part I.)

approach set a modified AN/MPN-1 (GCA). Since the differences between the Teleran attachments for the airway search and airport search radars are minor, only the former will be described, together with a description of the Teleran attachments for the final approach radar.

The television techniques used in the first Teleran model are conventional wherever possible. This paper describes only those variations from entertainment television practice which are peculiar to Teleran and assumes reader familiarity with entertainment television practice. Similarly, the radar equipment with which the Teleran gear is associated is not described in detail except for departures from conventional design practice.

Choice of Television Parameters

In order to make use of existing design data and components it was desirable to stay as closely as possible to RMA entertainment television standards. However, two requirements not found in entertainment television necessitated certain changes in components. First, as explained in Part I, many television pictures are transmitted from one ground transmitter. If each of these pictures were to be transmitted on a separate radio frequency a large portion of the spectrum would be utilized.

However, the motion of the spots representing aircraft on a radar display is relatively slow and a high frame rate is not required, as it is in entertainment television, to "stop" the motion. Accordingly, it was decided to use time sharing (time sequencing) methods and to transmit the separate television pictures from each ground station sequentially, identifying each frame so as to allow an aircraft receiving station to select the desired picture easily. Second, certain changes in the synchronization techniques were necessary in order to display the self-identification line whose characteristics are discussed on pages 605-606 of Part I. This line appears uniquely in each aircraft picture passing from the center of the picture through the pip corresponding to the aircraft in which the picture appears. The self-identification line is produced in the following way: a bright line rotating in synchronism with the ground antenna is picked up by a television pick-up device on the ground; the video signal of the line is transmitted aloft as a part of every television frame at a different amplitude from the remainder of the picture video. This self-identification video is separated from the picture video proper in the airborne receiver. At the instant at which an airborne beacon is interrogated by a ground radar the beacon transmits a pulse to the television receiver over a co-axial line. This pulse actuates a precedence circuit which serves to remove the normal picture from the kinescope for the next frame and substitutes therefor a picture of the rotating line. Inasmuch as the rotating line is in synchonism with the ground antenna which is interrogating the airborne beacon, the line on the kinescope appears to pass through the proper spot on the kinescope screen. In order that self-identification lines will not be observed from all ground radars in the vicinity which might interrogate the airborne beacon, the radar pulse repetition rate is made synchronous with the television horizontal synchronization rate, and coincidence between a horizontal synchronizing pulse and a radar interrogation is required in order to actuate the self-identification precedence circuit. Thus, if there is some slight spread among the radar repetition rates in the various ground radars the self-identification line will be seen only from that ground radar to whose associated television transmitter the airborne receiver is tuned. Inasmuch as many ground radars, particularly those with spark-gap modulators or delay line moving target indicator, must operate self-synchronously, it is necessary that the television line frequency be derived from the radar repetition frequency.

The television line frequency was chosen as 16.2 kilocycles (reasonably close to the 15.8 kilocycles used in entertainment television) since this is a multiple of the 900 cycle repetition rate which is employed by the ground radars with which installation was to be made. With the phosphors available for airborne kinescopes at the time decision on the value of television parameters was necessary, it appeared that a screen brightness of 50 foot lamberts could be obtained with a frame rate of 10 per second on any individual frame in the time-shared sequence, but that a lower frame rate would not allow the desirable screen brightness. Since a maximum of four pictures (3 altitude channels and 1 weather map) was to be transmitted in the experimental installation on a single radio frequency, it was therefore decided that the frame frequency should be at least 40. Actually 45 was chosen. In order to minimize flicker it was decided to use single interlacing so that the composite field frequency is 90 per second. In achieving the interlacing required it was again necessary to depart from standard practice. It is desirable for visibility reasons to scan the odd lines (i.e. 1st, 3rd, 5th, etc.) of the first picture in a time shared sequence, then the odd lines of the second, third and fourth, then to repeat for the entire sequence scanning the even lines. The odd line interlacing used in entertainment television can only accommodate an odd-number of time shared pictures. Further, in an odd-line interlace system the blanking time is used for transmitting the equalizing pulses. In a time-shared system, it is necessary to use the blanking time for frame identification pulses and no time is available for the equalizing pulses. Accordingly, it was decided to use even-line interlacing, a direct current component being injected into the vertical deflection of both the camera and display tubes, on alternate fields, in order to accomplish the required vertical displacement.

SYNCHRONIZING EQUIPMENT

The synchronizing equipment has two broad functions: (1) to furnish pick-up devices and monitors with horizontal drive pulses, vertical drive pulses, interlace locking pulses, blanking pulses, and time sequence selecting pulses, which control the order of transmission of pictures in a time-shared sequence; (2) to furnish the transmitter with a mixture of horizontal and vertical synchronizing pulses, interlace coding pulses and frame identification pulses which establish the identity of any picture in the time-shared sequence. The synchronizing generator is modeled after the studio Synchronizing Generator built for entertainment television purposes, and circuits employed in that equipment have been used liberally.

As mentioned previously, the pulses transmitted from the ground radar must be coincident with horizontal synchronizing pulses in order to prevent the appearance of false self-identification lines on the airborne display. Both the transmitted radar pulses and the horizontal synchronizing pulses are two microseconds in width, so that the circuits maintaining their coincidence must have a precision better than 2 microseconds. Synchronism between the transmitted radar pulses and horizontal synchronizing pulses is assured in the following way: The output of the horizontal oscillator (16.2 kilocycles) is peaked and the pulses are counted down eighteen-fold to a rate of 900 per second. These counted-down pulses are used to trigger a gate-forming multivibrator which is used to allow horizontal synchronizing pulses to pass through at the 900-cycle rate. Pulses derived from the radar modulator are used to form saw-tooth-shaped wave forms which are mixed with the gated synchronizing pulses. This mixture is then acted on by a discriminator whose output voltage controls a reactance tube which is coupled to the horizontal oscillator. Laboratory measurements have shown that the desired coincidence precision of two microseconds can be maintained with repetition rate variations of about 1 per cent.

The four-fold time-sharing generator consists essentially of 4 thyratrons which are made to conduct sequentially on successive vertical synchronizing pulses. Currents flowing through resistors in the cathode circuits of these thyratrons produce voltages which are used as selecting pulses to the various pickup devices and monitors.

The output of the various pick-up devices must be identified in such a way as to facilitate choice in an airborne receiver of the proper picture in the sequence. The frame identification coding used consists of pairs of pulses separated by time intervals which are multiples of the horizontal synchronizing period. As mentioned previously these coded pairs are transmitted during the vertical blanking time. The coded pairs are generated in the following way: The vertical blanking pulse is differentiated producing a narrow pulse at the beginning of the vertical blanking interval. This pulse is used to trigger a phantastron whose "on" time is controlled by the proper thyratron in the time sharing generator. The trailing edge of the output wave of the phantastron is then used to form a narrow gate which allows one horizontal synchronizing pulse to pass through. Thus, if the horizontal synchronizing pulse occurring at the time of the onset of vertical blanking is numbered 1, successive frames on one radio-frequency channel are identified by pulse pairs consisting of the first and third horizontal synchronizing pulses, the first and the fourth, the first and the fifth, the first and the sixth.

As mentioned previously it was necessary to depart from the odd-line interlacing system used in entertainment television and to adopt an even-line interlacing system. In this system each time-shared channel is transmitted once with the scanning raster in its normal position; and the second time a direct current is injected into the deflection coils of both camera and receiver, shifting the entire raster a distance equal to one-half the spacing between successive lines. This arrangement demands good vertical linearity since a departure from vertical linearity will result in poor pairing between the two fields in a frame. However, good linearity is a prime requisite in the Teleran television system in any case, and the linearity demands for even line interlacing are no worse than those dictated by operational requirements. An interlace trigger pulse is obtained from the plate circuit of the thyratron selecting the fourth channel in the time-shared sequence. This trigger is used to fire a two-stage ring oscillator. The ring oscillator is thus fired every fourth field and there is derived from it a symmetrical square wave of four fields duration. Application of this square wave to pick-up devices and monitors in the ground equipment produces the desired interlacing. The square wave of four fields duration is too long for transmission over the radio frequency link and the interlace trigger is transmitted instead, each receiver being supplied with a circuit similar in function to the two-stage ring oscillator just described.

TELEVISION TRANSMITTER

The television transmitter associated with the airways radar operates on a carrier frequency of 300 megacycles. This frequency is not contained in a band allocated to air navigation purposes and thus can be used only for an interim period. The frequency was chosen in order to obtain rapid systems tests earlier than would be possible at the higher frequencies for which air navigation allocations are available. Considerable operational experience in the use of 300 Mc for ground-to-air transmission was already on hand because of the very extensive tests during the war with Block III equipment.[2] In type of modulation it is desirable again to depart from conventional television practice. The Teleran picture, consisting of white spots and lines on a black background, contains no half tones. The fact that the black level is transmitted for approximately 90 per cent of the time makes possible a material saving in transmitter power by assigning the black signal a low level of radio-frequency energy and modulating so that maximum energy corresponds to a white picture spot. Since all of the spots and lines are of equal brightness, they all modulate the transmitter to an equal extent. As explained previously, in order to produce the self-identification line on the airborne display a separate picture of a rotating white line is mixed in with the picture video proper. The fact that aircraft spots and map signals are all of equal amplitude makes it possible to transmit the self-identification video as modulation in the same direction as the picture video, but at a reduced level, and to effect separation in the airborne receiver by amplitude discrimination. It was, therefore, decided to modulate the transmitter upward for both picture and self-identification video, but to make the percentage modulation used for self-identification one-half that used for the picture. The modulation for a typical Teleran line is shown in Figure 1.

The radio-frequency source is a master oscillator whose signal is amplified by a buffer amplifier and further amplified to a controlled extent by a modulated amplifier. Each of these stages comprises two triode tubes operated in a grounded-grid circuit, using parallel wire transmission lines for cathode and anode tuning, and grid modulation in the modulated stage. In order to suppress frequency modulation, both the buffer and the modulated amplified are neutralized. The high level assembly comprises sufficient radio-frequency amplification to bring the output of the exciter to a suitable level for transmission. Two stages of amplification are used, each consisting of one 5588 vacuum tube operated in a grounded grid circuit, the cathode and anode tuned

[2] M. A. Trainer and W. J. Poch, "Television Equipment for Aircraft", *RCA REVIEW*, Vol. VII, No. 4, pp. 469-502, December, 1946.

circuits being resonant coaxial transmission lines. Associated with the final stage is a small rectifier-monitor and cathode follower, allowing the take-off of d-c and a-c components of the radio-frequency envelope for monitoring purposes. The transmitter has a power output of 35 watts unmodulated and a peak power of 600 watts for a 10 per cent duty cycle.

The entire transmitter, except for the antenna, is housed in four racks. One contains power supplies for the entire unit; the second, the master oscillator, buffer and modulated amplifier and the two high level assemblies; the third contains the monitor and cooling fan for the entire transmitter, while the fourth contains a mixer panel for receiving the outputs of the synchronizing generator and the various pickup devices, clipping and limiting them, mixing them and transmitting them to the modulated amplifier.

The antenna consists of a four section pylon, the elements of which are fed so as to reduce the amount of energy radiated at angles near the zenith and increase the gain at low elevation angles.

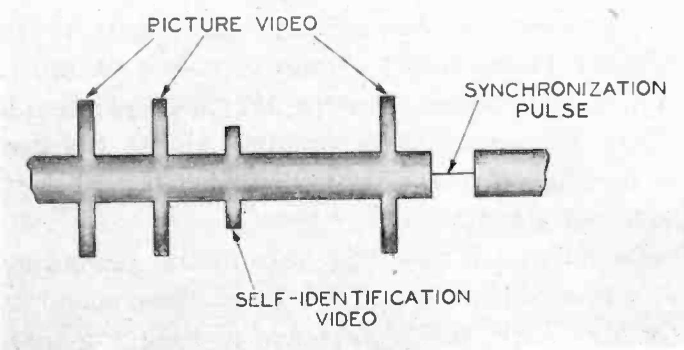

Fig. 1—Modulation for a typical line.

Altitude Coding

Separation of aircraft by altitude layers is achieved by having an airborne transponder reply to interrogation from the ground with an altitude dependent code and subsequent sorting of these coded replies at the ground station. There is a large variety of possible methods of achieving this coding and the choice of any one is dependent on layer thickness desired, accuracy obtainable, simplicity of coding and decoding devices and possible ambiguities.[3] A complete discussion of the possible coding methods is outside the scope of the present paper; for

[3] An excellent discussion of the ambiguities for cases in which aircraft can be assumed to be in random flight is contained in Airborne Instruments Laboratory Report No. 506-2, "Spurious Responses in Spaced-Pulse Coding of Airborne Transponders for Air Traffic Control", September, 1946.

the present it may be said that the choice was a three pulse code, the separations between the first and second and between the second and third being equal and a function of altitude.

Although it seems probable that ultimately a larger number will be necessary, the present system has been designed for eight altitude layers. (Studies have been made on methods of extending this and other coding systems to a larger number of layers.) Three sets of ground displays are provided, set for 0-2000, 2000-4000, and 4000-6000 feet. Adjustments are provided to place these displays at other altitude levels if desired.

The means employed for encoding the beacon response will be described in a subsequent section together with a description of the other beacon features. Coded beacon replies are received on the ground, and, after detection, are fed to a video distribution chassis, which shapes the pulses and distinguishes the first beacon pulse of each triplet by its coincidence in time with the radar pulse reflected from the aircraft. This chassis also supplies an output to each decoder channel, which in turn, sorts out groups of pulses within the required range of spacing, and feeds them to the appropriate plan position indicator (PPI). The decoding technique used is the comparison in a coincidence circuit, of signals delayed through an artifical delay line, with undelayed signals.

Figure 2 is a functional block diagram of the ground decoder. The beacon signal passes through a pulse stretching circuit which adds to the pulse width. This stretched pulse is used for three purposes. First, it is fed to a pulse width limiter which cuts the pulse to a determined size independent of the input pulse over a range of about 2:1, including pulses both wider and narrower than the output pulse. This is the beacon signal for the decoders. Second, the output of the stretcher is fed along with the radar pulse to a coincidence tube, and the output of this tube provides an instigating pulse for the decoders. Third, the stretched pulse is fed through an anti-coincidence circuit which selects those radar replies which do not coincide with beacon responses. These pulses can then be used to provide a display of aircraft unequipped with Teleran. The decoder section of the block diagram represents only one of the several decoder chassis used.

With reference to the action of the decoder, the instigating pulse is fed through the delay line and is reshaped and amplified after passing through the line. This delayed pulse and the shaped beacon pulses are then fed through a coincidence tube, the output of which goes through an identical process of delay and reshaping, and is again matched against a beacon pulse in a second coincidence tube. The single output pulse from these two circuits will be delayed with respect to the first pulse by an amount depending on the code channel, and after the final coincidence, it is fed through an additional delay line in order to bring the overall delay up to a standard of about 32 microseconds. Thus, the delay in each decoder channel is the same and all altitude displays may use the same trigger pulse for their initiation, this trigger pulse occurring, of course, about 32 microseconds after the trigger pulse for the radar transmitter.

Cameras and Mixing

As was stated previously, the airways station in this experimental installation transmits three altitude pictures. In addition, there is provided, on a separate time-shared channel, a weather map which can be viewed on any airborne equipment at any time. Thus, in the

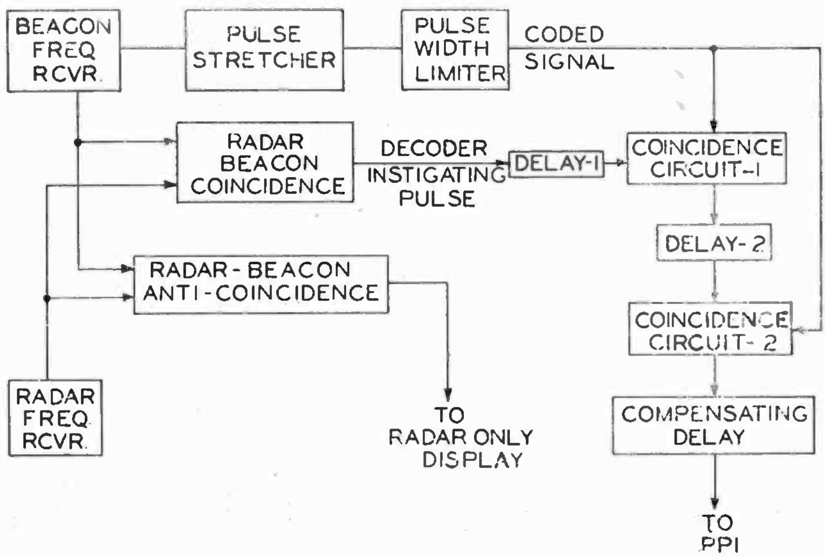

Fig. 2—Block diagram of ground decoder.

airway stations there are five televsion pickup devices. Three of these pick up the aircraft echoes from PPI displays and mix therewith map and control information. One picks up the bright rotating line to produce the self-identification line, and the fifth picks up the weather map. These different pick up devices will now be described in some detail.

One of the prime requisites for a Teleran system is a means for changing radar information into television form. Narrow beam high resolution ground radars, because of the size of antennas used, are necessarily limited in azimuthal scanning speed to a few revolutions per minute; whereas, for good visibility and high brightness, television pictures must be transmitted at many frames per second. The ground radars with which the first Teleran system is connected rotate at 12 revolutions per minute. In television terminology this is a radar frame rate of 1/5 per second. In order to change the radar information into television form, it is necessary to have some kind of storage device which will receive the radar information and retain it for one, or more, radar scans. There seemed to be two general methods of approach to the solution of this problem. One was the storage of the radar information on a long persistence cathode ray tube and the television pick-up of this tube, employing the most sensitive pick up tube available. The second possibility was the storage of the information in the television pick-up tube itself. Actually, both methods were tried out, but the second proved considerably better. The difficulty of phosphor storage seems to be two-fold; the persistence of the cathode-ray tubes available seems to be somewhat less than adequate and the range of brightness of the cathode-ray tube ouput, from very bright flash to very dim afterglow after a few seconds, is too great. An image orthicon used at sensitivities sufficiently high to pick up the dim afterglow charged up very severely on the initial bright flash. An orthicon pick-up tube having a very thin (and consequently high capacity) target, employing a very small beam current so that many scans are necessary to erase a given picture charge, has proven to be an eminently successful means for obtaining the acquired storage. With the exception of the target, this tube (the storage orthicon[4]) is in all respects similar to the 4½ inch orthicon used for entertainment television pickup purposes.

With respect to the mixing of radar and graphical information it was possible to do this by either optical or electronic means. At the time it was necessary to make a decision between these two, it did not seem to be possible to insure the required high degree of linearity and the accuracy of register with electronic methods. Accordingly, it was decided to mix optically.

A double relay rack assembly pictured in Figure 3 contains all of the pickup equipment for one altitude layer. Light from the PPI tube is transmitted through a mirror (having neutral color characteristics and 70 per cent transmission, 30 per cent reflection) placed directly in front of it and is reflected from a dichroic mirror (reflecting blue light and transmitting red light) into the storage orthicon camera tube. Light from the map is reflected from the neutral mirror, and again reflected from the dichroic mirror, and is optically superposed with the image of the PPI on the storage orthicon target. The map is made on plate glass and is illuminated from below by a set of

[4] S. V. Forgue, "Storage Orthicon and Its Application to Teleran", *RCA REVIEW*, Vol. VIII No. 4, pp. 633-650 (this issue), December, 1947.

fluorescent lamps. The monitor on the output television picture is housed in a desk-high attachment to the relay rack at which an operator can be comfortably seated. Within the operator's reach there is placed a small pen-like device which he can move over the surface of the monitor tube. This is connected, by a pantograph arrangement, to a

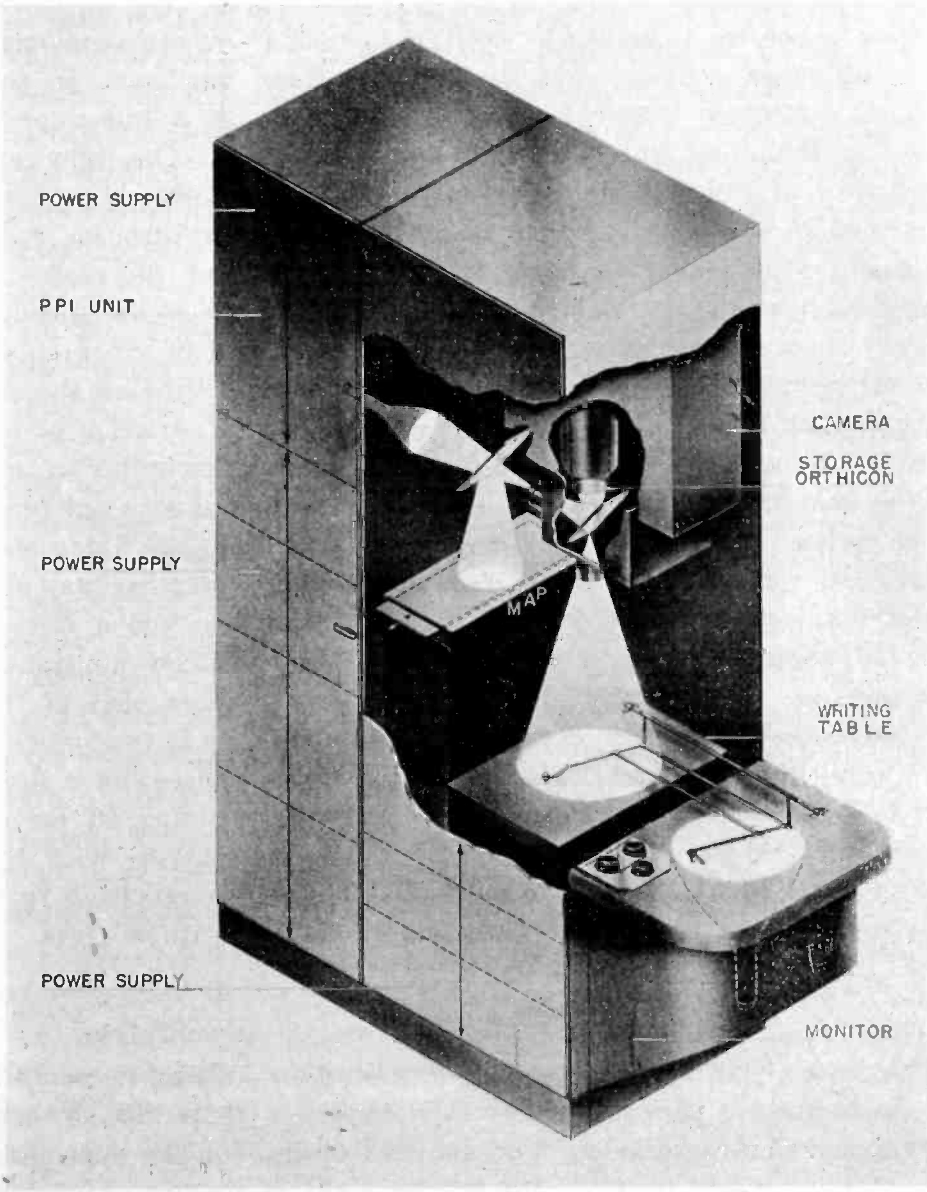

Fig. 3—Altitude console: (This unit contains all pickup and ground display equipment for one altitude layer).

sharp stylus which scratches a blackened surface on the writing table, its motion duplicating the motion of the pen in the hand of the observer. Fluorescent lamps inside the writing table illuminate the inscriptions on the surface of the writing table. The light which passes through the inscriptions is focused on the target of the storage orthicon camera and is thus optically superposed with images of the map and the PPI. The camera output is fed to the monitor and to the television transmitter.

For pick up of the self-identification line and weather maps it is unnecessary to provide storage in the television pick up device. Instead of using a camera tube it was decided to take the more economical course of using a flying spot scanner[5] for these purposes. In this system a television raster formed on the screen of a kinescope is imaged on the object to be viewed by means of a lens. The light rays transmitted or reflected by the object are condensed onto the photocathode of an electron multiplier tube. In the self-identification rack, the flying spot kinescope is mounted vertically and its raster is focused by means of a standard camera lens onto a black disc through which is cut a narrow radial slit. This disc is rotated in synchronism with the ground radar antenna and the light passing through the slit is focused by another lens onto the photocathode of a 931A photomultiplier tube. The weather map pickup is somewhat similar in that a flying spot scanner is used but is different in that it was not practicable to use a transparent weather map. Light from the flying spot kinescope is reflected from the surface of a standard weather map and is picked up by a special phototube. This phototube has a 5 inch diameter photocathode and contains a 7-stage electron multiplier. This tube has a sensitivity approximately twenty times that of the type 931A. Since the noise developed in the phototube is proportional to the total light on its cathode, the equipment is enclosed in a light tight box. Further, since the phototube must be mounted off center with respect to the map, reflection from the map varies with the position of the spot. In order to minimize the shading resulting from this effect, the box containing the assembly is lined with mirrors.

Landing Display

The instrument approach display employed in Teleran presents a spot on the face of the cathode ray tube displaced to the left or right of a vertical line representing an extension of the runway depending on whether the aircraft is to the left or right of its proper ground track.[6] In addition, a horizontal line shows the plane's position with

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[5] The flying spot scanner is described in: "Simultaneous All-Electronic Color Television", *RCA REVIEW*, Vol. VII, No. 4, pp. 459-468, December, 1946; see also G. C. Sziklai, R. C. Ballard, and A. C. Schroeder, "An Experimental Simultaneous Color Television System — Part II: Pickup Equipment," *Proc. I.R.E.*, Vol. 35, No. 9, pp. 862-870, September, 1947.

[6] See, for example, Figure 11 of Part I.

respect to the proper glide path, the line passing through a pip when the corresponding plane is at the correct altitude.

An AN/MPN-1 (GCA) radar is used for Teleran instrument approach. The GCA uses separate antennas for azimuth and elevation, each of them making a complete scan in two seconds. At the end of the swing of the antenna in each direction there is an interval of time which is useless for display purpose, since the antenna is changing direction. By moving the azimuth and elevation antenna, 90 degrees out of phase with each other it is possible to alternate the two portions of the display. Accordingly, an azimuth display is written on as the azimuth antenna swings from left to right; a vertical display as the vertical antenna sweeps up, an azimuth display as the azimuth swings right to left and a vertical display as the vertical antenna sweeps down, thus completing the cycle. The azimuth portion of the Teleran display is formed in exactly the same way as is that of the GCA as normally used. The elevation portion of the display is applied to the same tube alternately with the azimuthal portion. The horizontal line which indicates the elevation of each aircraft is drawn on during the elevation portion of the cycle and is formed by having the signals received from the elevation antenna trigger a high-frequency oscillator which applies a deflection field to the kinescope. In addition, the elevation signals intensity modulate the cathode-ray tube. The vertical displacement of this horizontal line from the spot is achieved by varying the point at which the vertical kinescope sweep starts according to the instantaneous elevation of the elevation antenna. The vertical sweep of the elevation cycle starts at the same point as the vertical sweep of the azimuthal cycle when the vertical antenna is at the elevation of the desired glide path. When the vertical antenna has an elevation above that of the desired glide path, the vertical sweep on the elevation cycle is started below the start of the vertical sweep on the azimuth cycle.

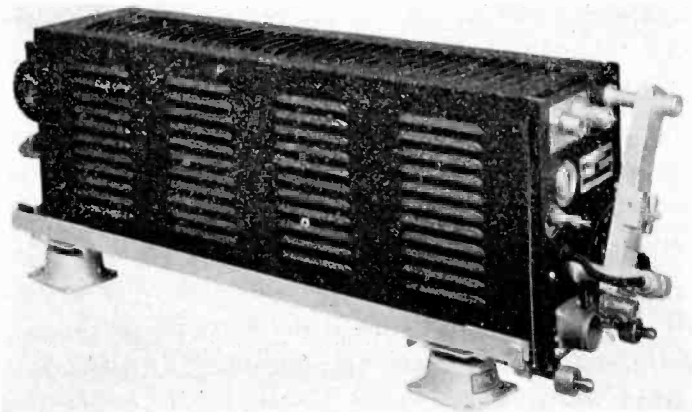
A correction in the landing display is necessary due to the fact that the precision radar equipment may be displaced some thousands of feet from the touchdown point. Under these circumstances, the elevation angle subtended at the elevation antenna by an aircraft which is maintaining the proper glide angle will become smaller and smaller as the aircraft nears touchdown. Correction for this is accomplished by a retardation in the vertical sweep of the landing display for short ranges.

Airborne Beacon

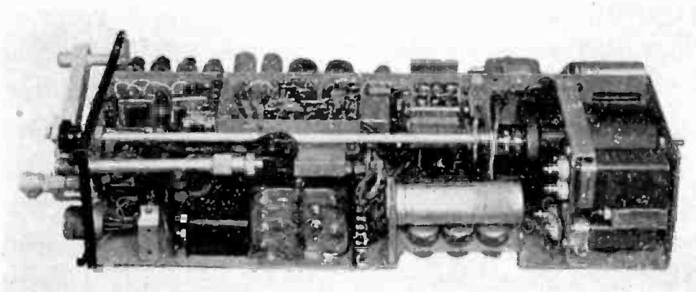
The airborne beacon must receive, omni-directionally, interrogation pulses from ground radars and re-transmit, again omni-directionally, a set of three pulses whose separation varies with the altitude of the aircraft carrying the beacon. It must send an indication of the receipt of an interrogating pulse to the television receiver for self-identification purposes and it should have circuits permitting it to respond only to interrogation from the main lobe of a ground radar equipment, rejecting side lobe interrogation. The AN/APN-19 beacon served as a model for the Teleran beacon, the receiver and transmitter circuits of the latter having been copied directly. A number of modifications and additions were required, however, for purposes peculiar to Teleran.

The beacon has two radio-frequency heads: one for the receipt of interrogations from the airway or airport radar; the other for receipt of interrogation from the final approach radar. Crystal detectors are used for both frequencies. The output of one or the other (depending on whether an approach or a navigational picture is desired) is fed to a video amplifier. Output video pulses are fed to the television receiver to initiate the self-identification circuits and are also used to initiate the coding process by triggering a ringing circuit. The frequency of this ringing circuit is controlled by an air dielectric capacitor which is mechanically connected to a Kollsman Altitude Variation Indicator. This latter unit contains an aneroid capsule and a small servo-mechanism, and produces a shaft rotation which is linear function of altitude. The output of the ringing circuit is fed to a blocking oscillator which forms a series of short duration square pulses separated by precise intervals of time determined by the period of the ringing circuit. These pulses are fed back to a counter tube which counts the number of pulses out of the blocking oscillator circuit and cuts the ringing circuit off after a total of three pulses is produced. Measurements on this coding circuit indicates that the code spacings (variable from 8 to 15 microseconds) can be held to about 1/10 microsecond under all probable variations of pressure and temperature.

One of the operational difficulties with airborne transponders is the fact that at small ranges the transponder is interrogated not only by the principal lobe of the radar set but also by the side lobes. Thus, if the response from an airborne beacon is followed on a PPI from long range to short range the spot will be very narrow at long range, its angular width being the same as the beam width of the interrogating radar. As the range closes, interrogation from the side lobes begins and the spots enlarge into arcs and at very close ranges may even become entire circles. The Teleran beacon includes a circuit which sets the gain of the beacon receiver after the receipt of a set of interrogating pulses to a level such that interrogation at a power level appreciably lower than that of the principal lobe of the radar antenna will not elicit a response from the beacon. An engineering model of the Teleran beacon containing the automatic gain control circuit was flown against an AN/CPS-1 radar in early tests. No interrogation by side lobes was observed between a maximum range of 75 miles and a minimum range of one-half mile. Flight tests of



(a) Complete with dust cover and mounting tray.



(b) Bottom view, dust cover removed.

Fig. 4—Airborne beacon.

an AN/APN-19 beacon which contained no automatic gain control, on the same day indicated very serious (about six-fold) broadening of the response due to side lobe interrogation at ranges less than 40 miles.

The beacon, which is pictured in Figure 4, is housed in a size one-half standard ATR rack (5 inches x 6¾ inches x 20 inches) and weighs 25 pounds complete with dust cover and shock mounts. Figure 4(a) shows the beacon complete with dust cover. Two antennas are used for receipt of interrogation (one from the airway or airport radar; the other from the final approach radar). One antenna is used for transmission. Figure 4(b) shows the beacon with the dust cover removed. The Altitude Variation Indicator can be seen on the extreme right.

AIRBORNE TELEVISION RECEIVER

In addition to the normal functions of a television receiver the Teleran receiver unit must perform the following: the selection of the picture for the proper altitude out of the time-shared sequence of pictures, the display of the self-identification line, the selection, from the composite television signal, of pulses carrying information for automatic flight, and presentation in front of the television kinescope of the instantaneous heading of the aircraft. The operation of the self-identification circuit has been explained previously. As was indicated earlier in the paper the first experimental installation will not contain provision for automatic flight although the transmitting and receiving equipment has been designed in such a way that automatic flight features can be added without any change in these components. The other two unusual functions, the decoding of altitude information and presentation of aircraft heading, are discussed below in some detail.

The frame identification coding consists of pairs of pulses separated by an integral multiple of the horizontal synchronization period. These pulses are decoded in the following way. A phantastron circuit, whose output pulse width is controlled by means of the altitude channel selector switch on the indicator panel, is triggered into operation at the beginning of every field by the vertical synchronization pulse. The trailing edge of the phantastron output is used to trigger a gate generating circuit which puts out a positive pulse approximately one line (62 micro-seconds) in width. This gate pulse is applied to the suppressor grid of a coincidence tube to whose control grid the frame identification pulses are sent. If the second of a pair of frame identification pulses occurs during the time that the positive gate is applied to the suppressor grid, a triggered pulse is produced at the plate of the coincidence tube and applied to a video enabler flip-flop circuit. In this way the video selector tube puts out a gating pulse only when a frame identification code pulse is present. In the interval chosen by the altitude channel selector, the square, positive going pulse from the selector tube which is exactly one field in duration, is applied to the suppressor grid of a gated video tube. This tube then passes a video signal during the frame which it is desired to display and no video signal is allowed to pass during the time that unwanted frames are being received.

The airborne kinescope screen is viewed through a transparent disk, called a heading disk, which carries a set of parallel lines and which is linked to a Gyrosyn or Flux Gate compass through a servo so that it is essentially a remote compass indicator. The center line on the heading disk can be read against an angular scale in order to obtain the heading of the aircraft. All Teleran navigation maps, i.e., those transmitted from the airport and airways station, are transmitted with North at the top of the picture. However, the Teleran landing display is transmitted in such a way that the runway in use appears as a centered vertical line on the display. In order to make the heading disk read properly for final approach an angular displacement must be introduced between the compass and the heading disk servo. This is done by means of a differential synchro, one side of which is fed by the compass, the other side of which is fed from a pilot-operated synchro mounted near the display tube. The heading disk performs the function of a color filter, being made of a colored plastic material having transmission characteristics similar to that of a Wratten No. 21 filter. This color matches the spectral characteristics of the cathode ray tube screen quite closely and the addition of this filter increases contrast considerably.

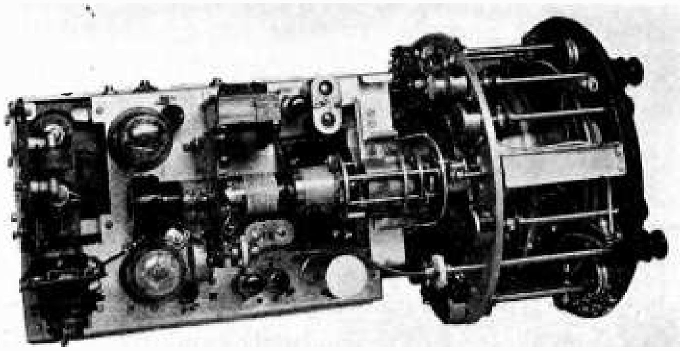
The airborne television receiver is packaged in two units. One is a cylindrical pressurized can 11 inches in diameter and 19 inches long containing the seven-inch kinescope, high-voltage power supply, deflecting circuits and final video amplifiers. On the front, around the face of the kinescope are the six controls, the altitude channel selector switch, focus control, brightness control, station selector switch, horizontal hold control and the angular scale adjusting knob for use just prior to final approach. This unit, called the indicator, weighs 29 pounds and is built for instrument panel installation. A front view of the assembled indicator is shown in Figure 5(a) and a top view of the chassis in Figure 5(b).

The second component of the receiver contains the power supply and all receiver circuits and is housed in a box 10⅛ inches wide, 10⅝ inches high and 19-9/16 inches deep. Figure 6 shows the internal construction of the unit: (a) shows the complete unit assembled, (b) shows the unit with its dust cover removed, and (c) shows the power supply and receiver units separated. The receiver chassis is built as two separate units which are mounted back to back and which are hinged so that the chassis can be opened for servicing. The complete receiver unit, including power supply, weighs 46 pounds.

The prime consideration in the design of the airborne equipment was the provision of equipment which would demonstrate the Teleran system. Considerable effort was expended to make this equipment as light and as compact as possible consistent with the original objective and it is felt that some progress in the direction of lightness was made. However, present knowledge will allow considerable saving in size and weight in the next model.

The airborne television antenna consists of a pair of horizontal dipoles, mounted at right angles to each other in a horizontal plane, their centers being spaced by a half wave length. These dipoles are coupled into a common line through a T-junction, the distance from

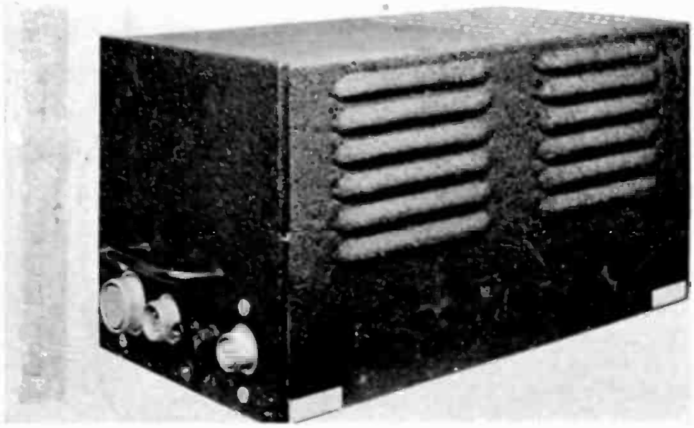
(a) Front view.



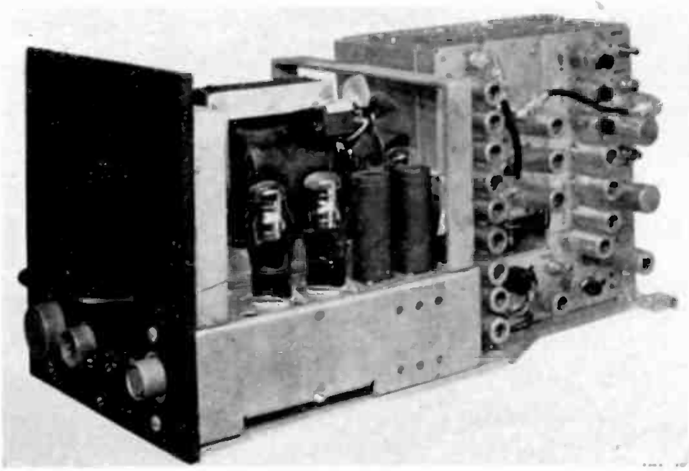
(b) Top view of chassis with pressurized case removed.

Fig. 5—Airborne display.

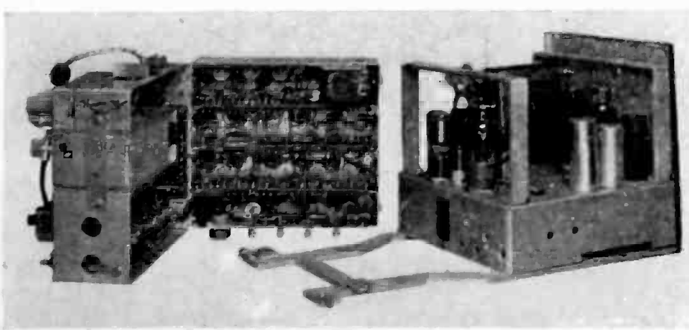
the T-junction to the center of the dipoles being the same in each case. This arrangement produces a horizontal antenna pattern which is very nearly omni-directional and tests taken over a flat plane showed differences of only 1.6 decibels between maximum and minimum points in the horizontal pattern. In actual installation the dipoles are mounted approximately six-tenths wavelength from the skin of the aircraft.



(a) Assembled.



(b) Dust cover removed.



(c) Receiver unit and power supply separated. (Receiver unit opened, showing suitcase design.)

Fig. 6—Airborne television receiver.

Acknowledgments

It is a pleasure to express appreciation to the Research Department of RCA Laboratories Division for the development of the storage orthicon, the airborne kinescope and mixing methods, to the RCA Victor Division Tube Department for its splendid cooperation in manufacturing the many special tubes required, and to colleagues of the Engineering Products Department for much valuable assistance and many helpful suggestions. Further, let there be no doubt that credit for the work herein reported is due all members of the Teleran Engineering Section.

THE STORAGE ORTHICON AND ITS APPLICATION TO TELERAN*†

BY

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Summary—An orthicon type of pickup tube, having a very high capacity target, and operating with a low beam current, has been used successfully to pick up a radar PPI presentation for television reproduction. By virtue of its large storage capacity the tube can reproduce for hundreds or even thousands of television scans information presented but once on the PPI screen.

INTRODUCTION

FOR commercial television, time lag effects have been looked upon solely as a defect to be eliminated. For example, in some of the orthicon-type of pickup tubes a moving white object might leave behind it a white trail, whose length was a measure of the inability of the scanning beam to discharge rapidly the picture charge on the target.

More recently there have arisen several applications for a pickup tube that could "remember" what it had seen for several seconds or even minutes. One such application arises in the Teleran system of aerial traffic control.[1] Here information from a radar PPI (plan position indicator) scope is picked up on the ground and rebroadcast by television to airplanes. Another application is one of projecting on a large screen a television picture of a radar presentation for mass viewing. Still another use is in the production of a bright kinescope picture of the relatively dim PPI scope information. This paper is primarily concerned with the first application.

In conventional television practice the frame time is quite short, being only 1/30th of a second. Here persistence of vision can blend the individual pieces of information laid down by the scanning beam into a complete picture. In radar the repetition rate may be of the order of several seconds, as set by the relatively slow rotation rate of

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* Decimal Classification: R339 × 629.132.5 × R583 × R537.

† Part of the work described in this paper was carried out under Contract W28-099ac-107 between the Watson Laboratories, Air Matériel Command, U. S. Air Force and Radio Corporation of America.

[1] D. H. Ewing and R. W. K. Smith, "TELERAN: Part I—Air Navigation and Traffic Control by Means of Television and Radar", *RCA REVIEW*, Vol. VII, No. 4, pp. 601-621, December, 1946.

the scanning antenna. In this case use has been made of the long afterglow properties of certain phosphors to produce a composite picture on the face of the PPI tube.

Such a picture might be picked up by a very sensitive pickup tube responding to this relatively dim phosphor afterglow. This actually has been done in the radar projection application mentioned above through the use of an image orthicon equipped with a Schmidt optical system.

Instead of using the time lag of the phosphor itself to produce the required storage effect, one might employ a pickup with a "memory" to look at the initial bright flash of the PPI scope. An orthicon with a very high capacity target, and employing such a small beam current that many scans are necessary to erase a given picture charge, has proven to be a means for attaining the required memory or storage.

This "memory" tube method of picture storage has appeared to have distinct advantages over the phosphor afterglow method, which advantages have recommended it for the Teleran application.

General Description of the Storage Tube and Its Operation

The name "storage orthicon" has been used to differentiate between the present tube and the orthicon used for conventional television. Aside from the very high capacity target, the tube construction is the same as that which might be used were it to perform as a conventional multiplier orthicon pickup tube. Except for the target design and lack of an image section it is also very similar in construction to the "mimo" tube (miniature image orthicon, 2P22) which has been enlarged to use, as far as possible, standard size image orthicon (2P23) parts. Figure 1 shows the external appearance of the storage orthicon.

The operation of the orthicon has been described in the literature.[2] In addition several papers have appeared dealing with the operation of the image orthicon.[3,4] The latter is almost identical with the orthicon as far as the electron gun, scanning, and multiplier sections are concerned. Therefore, only a somewhat general description of the orthicon tube and its operation will be given here, while later in the paper a more detailed discussion of some of the structural and operational characteristics peculiar to the storage orthicon will be given.

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[2] A. Rose and H. A. Iams, "The Orthicon", *RCA REVIEW*, Vol. IV, No. 2, pp. 186-199, October, 1939.

[3] A. Rose, P. K. Weimer and H. B. Law, "The Image Orthicon—A Sensitive Television Pick-Up Tube", *Proc. I.R.E.*, Vol. 34, No. 7, pp. 424-432, July, 1946.

[4] P. K. Weimer, H. B. Law, and S. V. Forgue, "MIMO-Miniature Image Orthicon", *RCA REVIEW*, Vol. VII, No. 3, pp. 358-366, September, 1946.

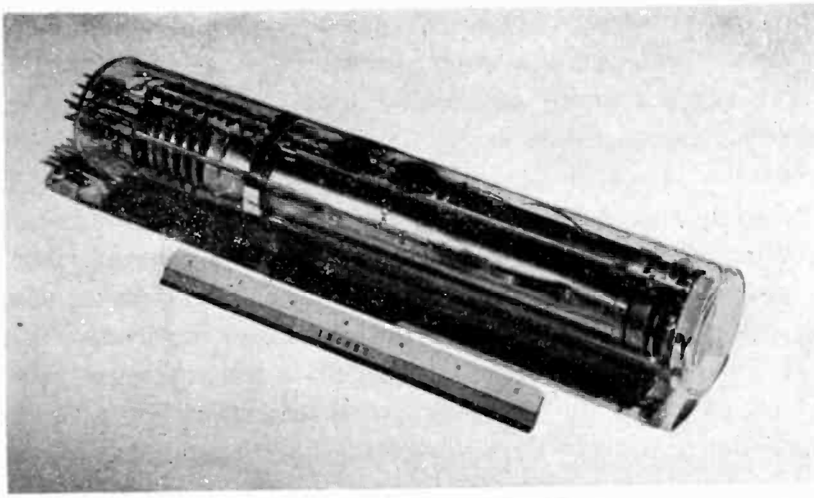

Fig. 1—The Storage Orthicon.

Figure 2 shows a diagrammatic cross section of the tube in a typical operational setup. In operation, a beam of electrons leaves the small (.002-inch) defining aperture in the end of the electron gun at about 200 volts velocity, and passes through the adjustable direct-current field of a rotatable alignment coil which accurately aligns the beam with the magnetic field of the focusing coil. It is then magnetically deflected horizontally and vertically so as to scan the target in a rectilinear pattern. The two mutually perpendicular sets of deflection coils have their fields perpendicular to the tube axis, and, in conjunction with the axial focusing field, give rise to a net warped field along the lines of which the electrons travel, to a first approximation. The beam leaves the deflection field almost parallel to the

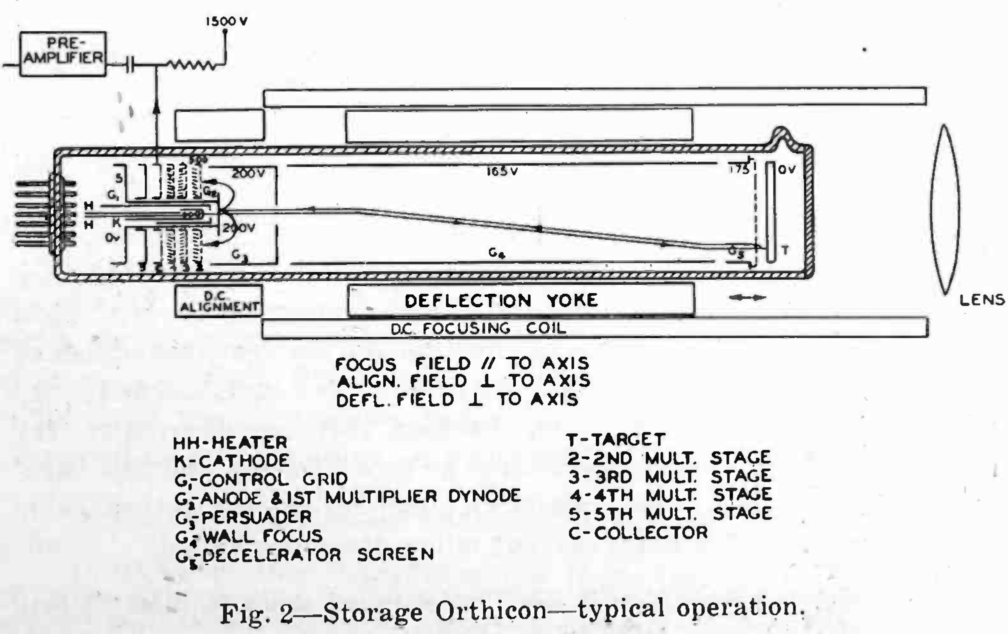

Fig. 2—Storage Orthicon—typical operation.

axis except for a small amount of helical motion imparted by the scanning field. Such helical motion means that the beam electrons have a transverse velocity component gained at the expense of the axial velocity. Because it is desirable that the beam reach the target perpendicularly, operating conditions are so chosen that this helical motion is as nearly as possible compensated for by an equal and opposite effect introduced by the electron lens in front of the target.

Principally in the region between the decelerator screen and target, the beam velocity is reduced to almost zero. As electrons striking a surface at low velocities knock out less than one secondary electron* per incident primary, on the average, the beam tends to discharge the surface down to cathode potential (or a volt or two negative if account is taken of the initial thermal emission velocities). When the insulated target surface is charged down to this potential, the beam no longer lands and cannot charge the target surface more negatively. Thus, the cathode is at a stable potential and this is the potential assumed in the absence of light and presence of the beam. In this condition the beam closely approaches the target, is reflected back approximately upon itself, and returns to the multiplier end of the tube. This is the condition for maximum direct-current output and zero signal current.

If part of the target surface is illuminated by light, (passing through the semi-transparent backing plate and dielectric) photoelectrons will leave the photosensitive (scanned) side, driving that part of the target positive. The field in front of the target is sufficiently strong to saturate this photoemission, the electrons of which are collected by the decelerator screen, wall, persuader and multiplier. When the scanning beam passes over the lighted parts, enough electrons in the beam will land each scan in these areas to drive them back down to cathode potential if the beam current is sufficiently large. In normal television operation, the current is adjusted so that this obtains.

The return beam is then modulated by subtraction according to the charge pattern left on the target by the escape of photoemission. It is a maximum for regions of no light and a minimum for high light areas. If the light were capable of effecting 100 per cent modulation of the beam, no current would return for areas of full light intensity. An amplifier connected to the target backing plate would receive (by capacitive coupling between individual picture elements and this backing plate) the video signal, while an amplifier on the return beam collector would see the beam current minus the video signal.

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* Secondary electron here is used in its broad sense to refer to both true secondaries and also reflected electrons.

In normal operation the potential swing of the target is usually not over two or three volts. A higher voltage swing would give rise to beam bending† at the target with resulting loss in resolution. In the event that the beam is insufficient to hold down the potential of a brightly illuminated area, this area can charge positively to a point which, in the storage orthicon, is determined almost entirely by the decelerator screen voltage. Usually this potential is high enough that the electrons strike at a velocity for which the secondary emission ratio is above unity and tend to maintain the high potential. To return the target to cathode potential it is necessary to lower the collector potential long enough to discharge the target and either to increase the beam or decrease the light if further charging is to be avoided.

The modulated beam returns to the first multiplier stage which also serves as the end of the electron gun. Here secondary electrons are released in a potential field configuration tending to pull them to the second stage which is struck at a velocity for which the secondary emission ratio is above unity. The 2nd, 3rd and 4th stages are "pin-wheel" type multipliers. Each stage presents practically an opaque surface to impinging electrons, while the slots between the vanes permit the collecting field of the succeeding stage to leak through for the saturation of the secondary emission. The coarse mesh screen in front of each stage prevents the retarding field of the preceding stage from hindering this collection. The secondary electrons from the last stage are collected by a collector screen which runs somewhat more positive. The current to this collector constitutes the video signal.

The Teleran Application of a Storage Pickup Tube

It is helpful to have in mind a picture of the Teleran system as a whole to understand the characteristics that a pickup tube must possess for successful use in this system. Figure 3 is a functional diagram of the basic teleran system as planned for initial installations.[5]

In brief, a ground search radar explores the air space of interest and displays the information received on several PPI scopes, each of which corresponds to a certain altitude layer into which the vertical airspace is divided. This division prevents the confusion that would occur if all the information picked up by the scanning radar were presented on a single scope. A plane equipped for the Teleran system

† Action of potential at one point affecting the beam landing at adjacent points.

[5] D. H. Ewing, H. J. Schrader, and R. W. K. Smith, "TELERAN: Part II—First Experimental Installation", *RCA REVIEW*, Vol. VIII, No. 4, pp. 612-632 (this issue), December, 1947.

carries a radar transponder which automatically replies to interrogation with a coded signal corresponding to the altitude of the plane. Thus, radar information is channeled to the appropriate scopes. How-

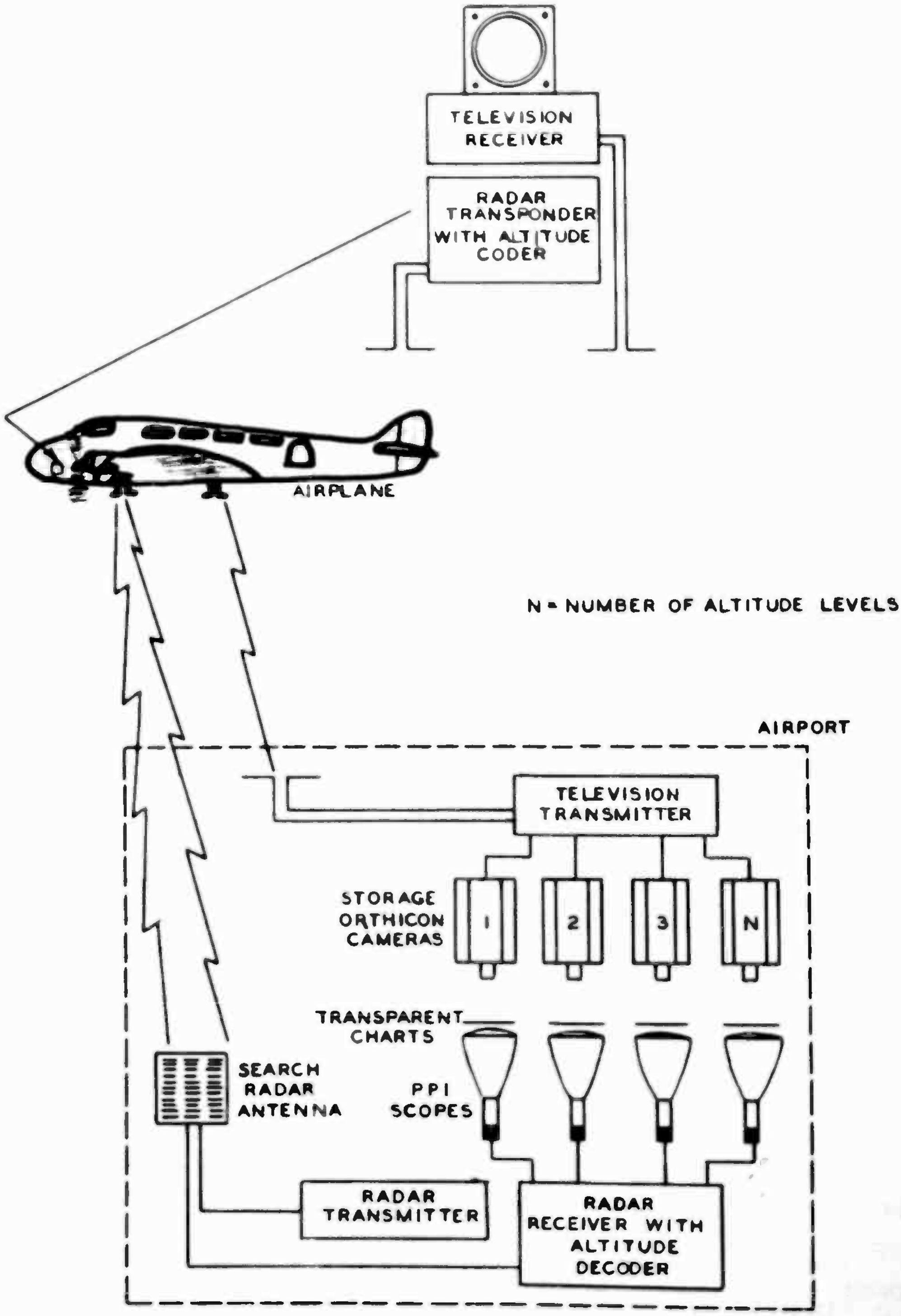


Fig. 3—Basic Teleran system.

ever, the television pictures sent to all planes are transmitted from one transmitter, but are segregated by a "time sharing" arrangement, where for n altitude levels, information about any one is broadcast approximately $1/n$th of the total time. Normally a pilot is primarily concerned with aircraft in or near his own altitude. He is able here, in addition, to look at information from other altitude levels, if he so desires.

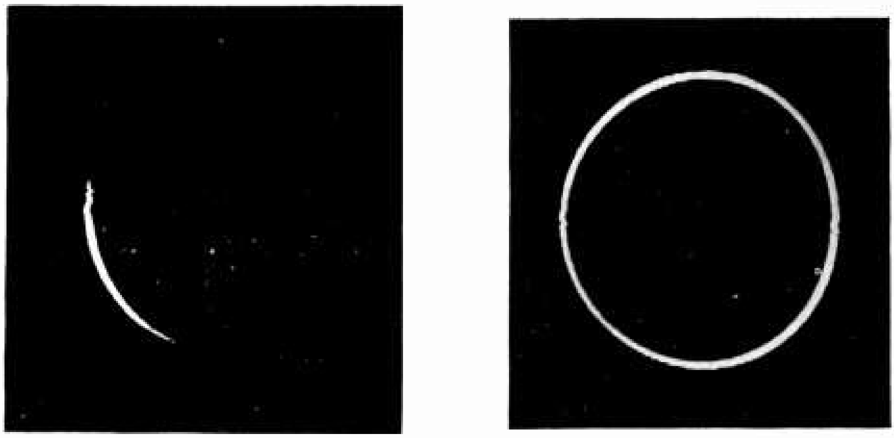
Airplanes show up on the television screen as bright spots (radar pips), with the spot corresponding to the pilot's own plane characterized by an individual bright radial line passing through it. Because of the storage elements in the system, several pips from a given plane may be seen simultaneously corresponding to several scans of the radar antenna. This permits one to determine the direction of motion of the plane with the weaker pips indicating the plane position on previous scans, and the brightest pip that on the last scan.

To indicate satisfactorily the direction of motion of a plane, it was felt that a minimum of three pips from three successive radar scans must be visible simultaneously. For a slow antenna rotational rate of say 6 revolutions per minute, and no time sharing, this means that at the end of 30 seconds (900 television scan times) an easily observable fraction of a signal pip must still be present.

With time sharing, however, as only one of the n cameras is furnishing information for transmission to a plane at any one time, each storage orthicon can operate with its beam current cut-off about $(n\text{-}1)/n$ths of the time. Thus the target signal on each tube is discharged only when signal is needed from that tube. Therefore, if a given storage orthicon has a useful time lag of t seconds under continuous operation it will have a useful time lag of nt seconds under time sharing operation, provided of course that the target electrical leakage is low enough so that the signal charge doesn't appreciably leak away in this time. (The latter condition has been easily met in practice.)

The Storage Orthicon Target

While the ordinary orthicon is operated so that the picture charge at any point is neutralized by a single beam scansion, it can be seen that the Teleran application requires a picture charge removal only after several hundred scans. This can be accomplished either by decreasing the beam current (see Figure 4) or increasing the target capacity. In actual practice both devices are used in the storage orthicon. Increasing the target capacity increases the storage time without sacrifice of signal-to-noise ratio. Merely decreasing the beam



a. High beam current. b. Low beam current.

Fig. 4—Reproduction of a rotating spot by the Storage Orthicon.

current increases the storage time at the expense of signal-to-noise ratio, the latter varying as the square root of the beam current.

The convenional orthicon has a mica target the capacity of which can only be increased to the point corresponding to the thinnest sheet into which mica can be split. In practice this usually turns out to be of the order of a half mil. In order to increase further the capacity it was necessary to make a target of dielectric material other than mica. Two approaches to this problem were tried. One was to make

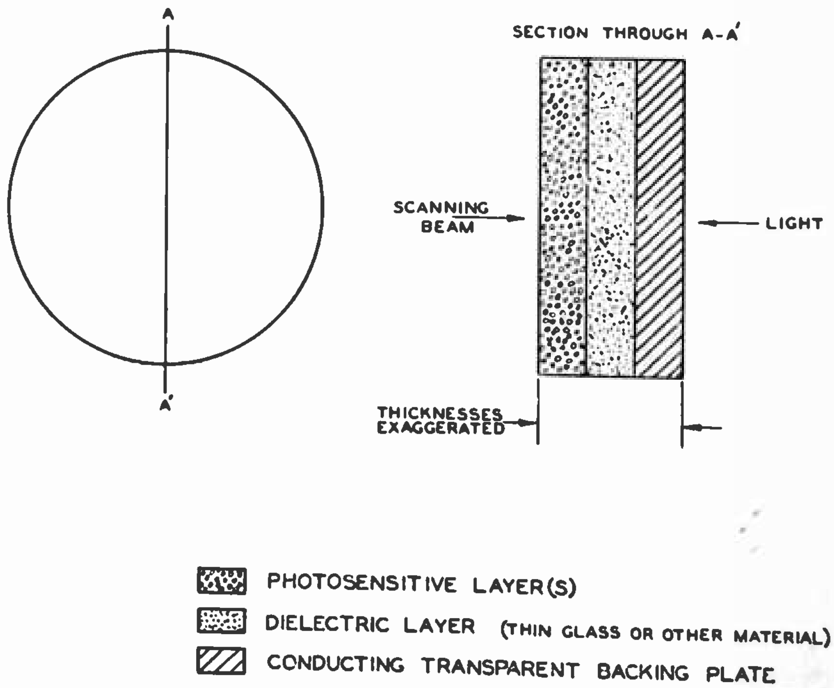


Fig. 5—Storage Orthicon target. (Section through functional layers)

a very thin glass target; the other was to form a thin dielectric layer on a transparent conducting surface.

The glass target is formed and mounted in a manner similar to that used in the production of thin glass targets for the image orthicon. However, unlike the image orthicon target, that of the storage orthicon must have a very high resistivity. A semi-transparent conducting backing plate is formed on the unscanned side of the glass surface, while one of the conventional photo-sensitive layers is formed on the scanned side after the tube has been sealed.

The storage time of the thin glass target is much longer than that of the normal orthicon by several orders of magnitude. If, however, for any reason a smaller diameter target or longer time lag than is produced by the glass type is needed in the future or for other applications, other material is available which has been found in actual operation to provide desirable and usable characteristics.

Figure 5 shows diagrammatically the functional layers of the storage orthicon target.

TARGET CHARGE-DISCHARGE CHARACTERISTICS

General Remarks

Normally, in pick-up tube operation, efforts are made to insure that picture signals are completely erased each scanning cycle if the defect of time lag (fuzzing of moving objects) is to be avoided. However, this defect for conventional television operation is a definite asset for reproduction of radar scope information, and has been exploited in a greatly exaggerated form in the storage orthicon.

The storage orthicon exhibits this time lag as a result of a very high target capacity in combination with a low velocity scanning beam. These are, of course, just the conditions giving rise to a long time constant for a condenser being discharged, and it is as a condenser, charged positively by the photoemission process and discharged by the beam current, that one considers an orthicon target.

Target Charge-Up

As the field in front of the target is normally made quite high, the photoemission from the target is saturated. Therefore steady light on the target causes a steady current to leave it. Thus the target potential builds up linearly with respect to the time t, rather than following the normal exponential condenser charging curve. The slope of this charging curve is determined by the capacity C of the target and by the photo-current leaving it. This photo-current is in turn proportional to the light intensity B on the target, and to the photosensitivity S

of the target. The relation between these factors can be expressed simply by

$$v = \frac{BAS}{930C} t + V_i \tag{1}$$

where v = potential after t seconds and V_i = initial potential, in volts

B = illumination in foot-candles, A = target area in square centimeters

S = sensitivity of target in microamperes/lumen

C = capacity of target in microfarads (associated with area A).

Target Discharge

The mechanism of target discharge by the scanning beam is somewhat more involved. There are two significant ranges in the discharge cycle, each of which has its own characteristics of discharge versus time. In the "normal range" all of the scanning beam I_B reaches the target each scan, thereby reducing the target potential by a fixed amount each scanning period. In the "intrinsic range" only part of the scanning beam can land each cycle, this part being a function of the target potential found by the beam where it is landing at any instant, and is therefore also a function of time. The first range ends and the second begins when the target potential drops to a value low enough that some of the electrons within the beam no longer have sufficient axial energy to reach the target. The first range is therefore one of essentially constant output current, while the second range is one of falling output.

The axial velocity distribution within the beam is determined by the range of thermal emission velocities from the cathode and by the different amounts of helical motion imparted to the beam electrons when passing through the electron lenses and magnetic deflection field within the tube. In the production of this helical motion a transverse velocity component is imparted to the electrons at the expense of their axial velocity, for their total velocities at any point are determined by the space potential at that point. Thus electrons which would normally have sufficient energy to reach the target in the absence of this helical motion, might, in its presence, be unable to do so. Normally a tube is operated so as to minimize the helical motion, thereby insuring perpendicular approach to the target. For this reason a velocity distribution arising only from the initial thermal velocities will be considered herein.

In this case, then, the transition between the normal and intrinsic ranges will take place when the target potential just equals cathode potential, account being taken of contact potential differences. At this potential an electron with zero axial emission velocity would just reach the target. Those emitted with more than zero velocity will reach the target with their emission velocity. Since the secondary emission ratio is less than unity at these low velocities, the beam will tend to charge the target negatively. In actual operation, and in the "dark", the beam will charge the target negatively until the beam current landing equals the current leaving the target.

While the secondary emission ratio is low at these low striking velocities, it is not zero. Consequently only part of the beam strik-

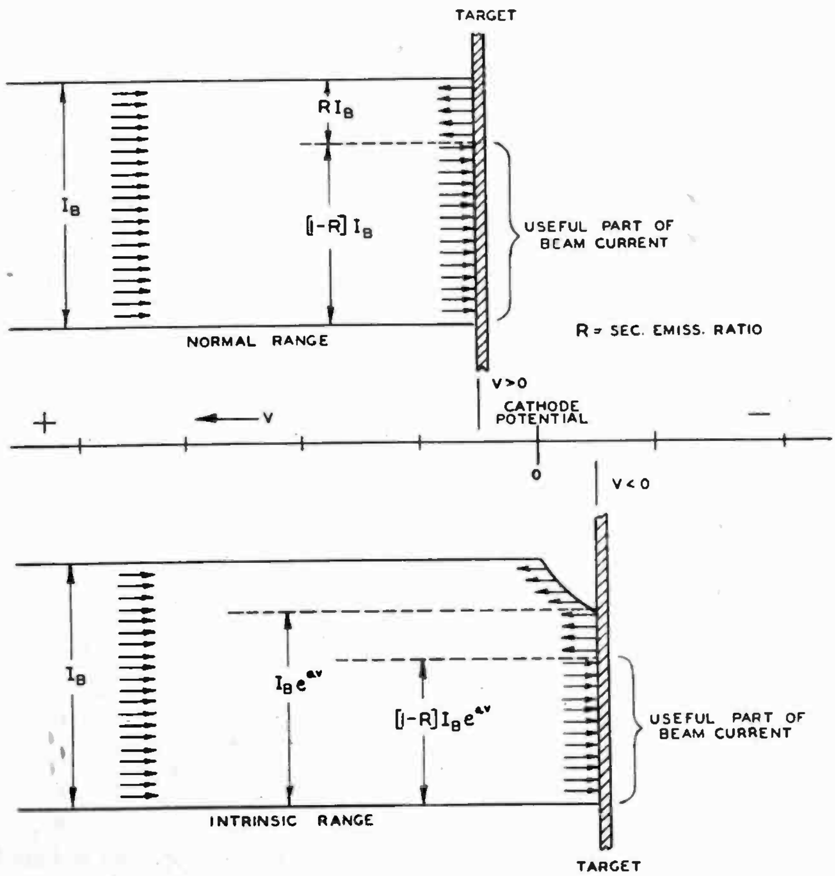


Fig. 6—Useful and non-useful parts of the beam current.

ing the target sticks. Like the fraction of the beam current that is energetically unable to land, this non-sticking current returns to the multiplier where it contributes to spurious signal and lowers the overall signal-to-noise ratio. Actually the secondary emission ratio R at the target is not strictly constant, and, in a rigorous derivation, its variation with potential would have to be accounted for. But the variation is small over the narrow voltage range of importance in the storage orthicon, and so will be treated here as a constant.

It has been seen that only a fraction of the full beam I_B leaving the gun is useful for discharging the target. In the normal range (target positive) all of I_B lands, but in the intrinsic range (target negative) only part of I_B has enough energy to land. It can be shown that the part of I_B that can reach the target when negative is given by $I_B e^{av}$ where $a = -\dfrac{e\,10^7}{kT}$ and $v =$ potential after t seconds. In either range only part of the beam that lands is absorbed, the ratio of absorbed to landing current being given by $(1 - R)$. Figures 6 and 7 summarize the information of this paragraph.

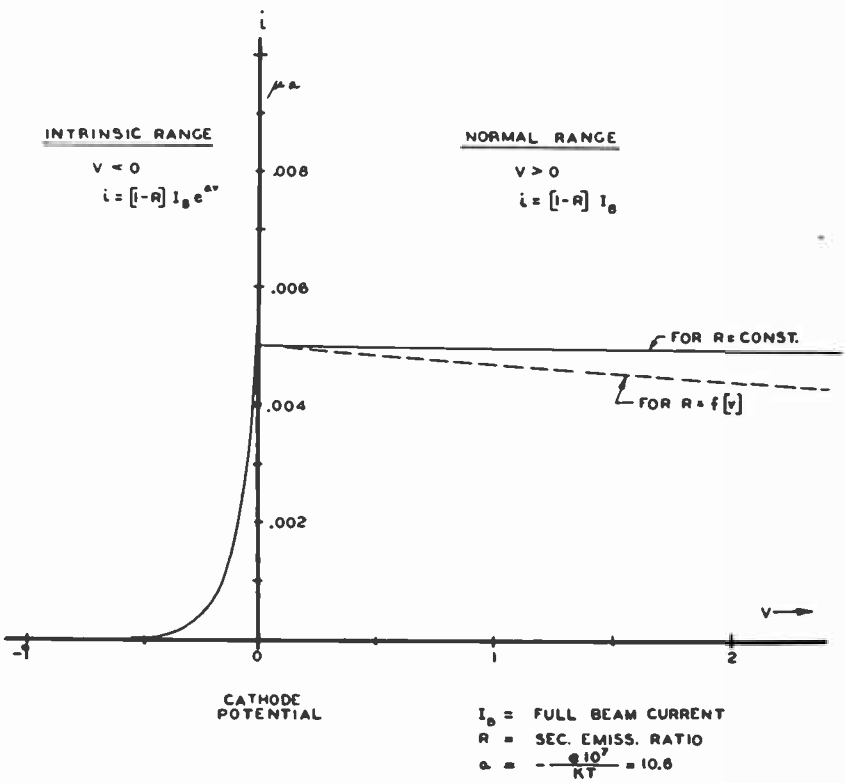


Fig. 7—Typical plot of the useful part of beam current as a function of the target potential.

After the storage orthicon target has been charged up by an exposure to light, the principal consideration is the rate at which the target discharges over periods of time very long compared to a scan time. Thus the fact that the potential of a picture element drops by a discrete amount each scan can be ignored and the potential can be treated as though it were dropping continuously. The entire target may be considered as a single condenser, initially charged up to some potential V_0 by uniform light over the whole target, and discharged by the beam acting through all of each frame time. It is legitimate thus to consider the whole target as a unit rather than to consider a small picture element alone, even though it may be the potential of the latter that is eventually wanted. This is true because the charge present on a given area and the length of time the beam is on that area are both proportional to that area. Thus the potential of a small element drops by the same amount each scan as would the potential of the entire target if the latter were charged to the same initial potential.

The discharge curves for the storage orthicon after the target is charged up by an exposure to light may now be considered. Leakage currents can be ignored here as the tube can retain a picture charge for hours after exposure to light and in the absence of the beam.

Normal Range ($v > 0$; Steady Output Current): Here the target, after being charged to some potential V_0 by a flash of light, is discharged *at a constant rate* by the useful part of the beam current,

$$i = (1 - R)\, I_B \tag{2}$$

The voltage v for any value of time t for which $v > 0$ is immediately given by:

$$v = V_0 - \frac{(1 - R)\, I_B}{C}\, t \tag{3}$$

where I_B is in microamps, V is in volts, t is in seconds

$C =$ total target capacity in microfarads.

Intrinsic Range ($v < 0$; Decreasing Output Current): Here the useful current i is given by $(1 - R)\, I_B e^{av}$, whence

$$\frac{dv}{dt} = -\frac{i}{C} = -\frac{(1 - R) I_B}{C}\, e^{av} \tag{4}$$

Solving this for v, and inserting the condition that $v = 0$ when $t = \dfrac{CV_0}{(1 - R)\, I_B}$ from (3) gives

$$V = -\frac{1}{a}\, Ln \left[1 - aV_0 + \frac{a\,(1 - R)\, I_B\, t}{C} \right] \tag{5}$$

and the useful current is

$$i = -C\frac{dv}{dt} = \frac{(1-R)\,I_B}{\left[1 - aV_0 + \dfrac{a\,(1-R)\,I_B\,t}{C}\right]} \tag{6}$$

where $a = -\dfrac{e\,10^7}{kT} = 10.6$, $\qquad e = -1.59 \times 10^{-19}$ coulombs

$k = 1.37 \times 10^{16}$ ergs/$k°$, $\qquad T = 1100°\ K$

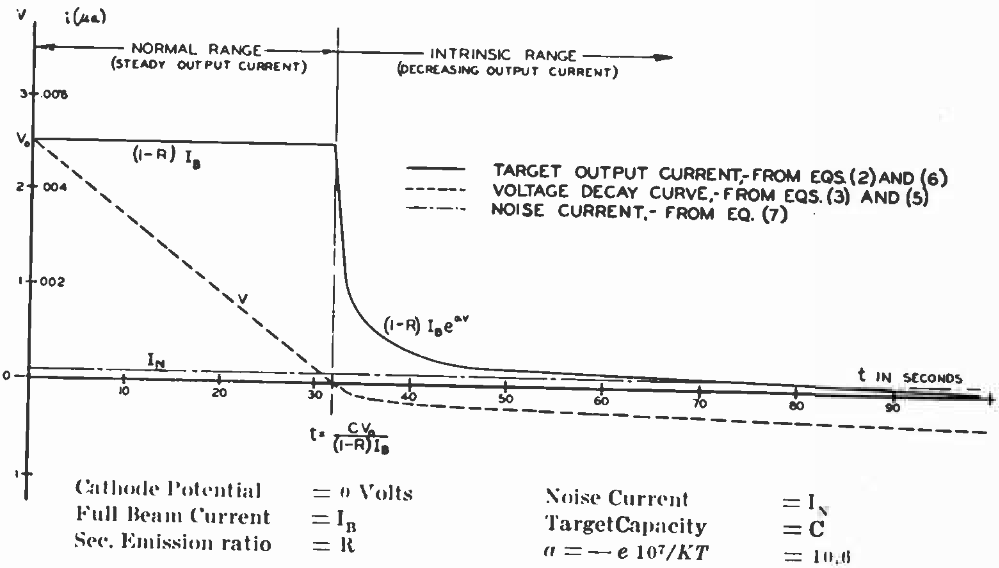


Fig. 8—Typical decay after single exposure to light.

A typical plot of the current and voltage as a function of time is shown in Figure 8. The useful current is proportional to the output current of the tube where the factor of proportionality is the multiplier gain. We see that the beam current tends to charge the target more and more negative as time progresses, but at an ever decreasing rate. The useful signal current eventually decreases to the point where it is masked by the noise current I_N associated with the full beam current I_B. The noise current is expressed by

$$I_N = (2e\,\Delta f)^{\frac{1}{2}}\,I_B^{\frac{1}{2}} = 1.26 \times 10^{-6}\,I_B^{\frac{1}{2}} \tag{7}$$

where $e = 1.59 \times 10^{-19}$ coulombs

$\Delta f =$ band pass, taken as 5 megacycles

The dot-dash curve of Figure 8 shows the noise current associated with the beam current used for plotting the two other curves of the figure.

Steady Light: If a steady light image is projected on the target, parts exposed to the image will come to an equilibrium potential such that the average current landing from the beam and sticking, just equals that leaving in the form of photoemission I_s. This equilibrium potential can be shown to be expressed by

$$V_e = -\frac{1}{a} Ln \frac{(1-R)\, I_B}{I_s} \tag{8}$$

Operation under steady light conditions takes place only at a voltage near zero (cathode) potential for I_s almost equal to $(1-R)I_B$, or at more and more negative voltages as I_s approaches zero. This is so, because anywhere in the "normal range" I_s must equal $(1-R)I_B$ if the voltage is not to change for a given value of I_s. This is not an equilibrium operating condition, however, for if the light intensity (and therefore I_s) increases somewhat, there is then not enough useful beam current $(1-R)I_B$ available to hold the target potential down, and the target "goes positive". If, on the other hand, the light is decreased slightly, the useful beam current is more than enough to neutralize the photoemission and charges the target to zero or below.

As opposed to this, the intrinsic range is a relatively stable operating range. If the light is increased so that the target would climb in potential because of the greater photoemission, more useful beam current $(1-R)\ I_B e^{av}$ becomes available and tends to hold the target down. Likewise a decrease in light is followed by a decrease in beam current landing, the two effects tending toward balance.

There can also be stable operation in the immediate vicinity of zero volts, where the photoemission swings the target just above zero volts before passage of the beam and the latter charges the target back down to slightly below zero volts. This resembles operation of the standard orthicon in normal television practice, where, however, the capacity of the target is so much lower, and the currents so much higher, that these voltage swings above and below the axis are much greater, and take place within a frame time.

Some Miscellaneous Characteristics

Half Tone Reproduction

In the normal range, by definition, all of the beam lands each scan irrespective of the actual potentials existing on the target. Half tone steps therefore would not be reproduced as long as the secondary emission ratio at the target were a constant over the small range in target potentials encountered. If the secondary emission ratio did vary significantly, half tone steps could be reproduced, but in reversed polarity. This is true, because the secondary emission ratio increases for higher target potentials (higher light) which means that more electrons (RI_B) would return from the lighter areas. On the other hand, in the intrinsic range, the current landing is a function of the target potential found by the beam. Here half tone steps can be discerned, and in the correct polarity.

Coplanar Grid Effects and Stray Light

If a bright pip corresponding to a plane is to be picked up successfully by the storage orthicon, it must be bright enough to charge up the target to the desired voltage in one sweep of the PPI beam. The momentary photoemission from such a pip is many times that of the useful storage tube beam current.

On the other hand the light intensity from a steady image, such as a map pattern projected on the storage tube, may be quite low, and must in fact be low enough that the photoemission from an area is equal or less than the average useful beam current to the area. This is to prevent the target from "going positive". Accordingly, when the map projection is suddenly turned on, it may take an appreciable time for the target to build up to operating potential. This is epecially true if it has previouly been discharged by the beam all the way down to its dark potential.

If, as is likely to be the case, the map information consists of narrow white lines, or small white dots on a dark background, the low potential of the large adjacent dark areas may cause enough of the potential barrier in front of the light areas that the beam cannot land on these areas (coplanar grid action) until the photoemission has charged them above their normal equilibrium potential. This, of course, greatly exaggerates the time necessary for build-up. The same grid action prevents small areas from charging to collector potential under normally excess lighting conditions. If such a transient effect temporarily prevents the beam from landing, the situation may be corrected by the use of a judicious amount of uniform stray light falling upon the target. This stray light falling on the previously very negative dark areas charges them up sufficiently so that their coplanar grid effect no longer is sufficient to prevent the beam from landing on the light areas. Although this stray light may only be needed long enough to put the tube back into operation, it may be helpful to use a permanent small part of it to prevent the high velocity electrons in the beam from charging the dark areas too negatively.

Photoemission Streaking

In the presence of a bright pip or strobe line on the PPI scope, very strong and objectionable horizontal lines or dashes were frequently found to be transmitted by the storage orthicon. These spurious signals were generated by the photoemission from the brightly illuminated parts of the target reaching and scanning the multiplier in the same manner that photoemission in an image dissector tube scans over its defining aperture. The corresponding aperture here is the persuader (G_3) aperture (see Figure 2).

The lengths of the streaks in the transmitted picture have been correlated with the time that the bright pip or strobe signals of the PPI are on. If one is dealing with only one or at most a few pip signals which are on but a few microseconds total time, this effect is usually negligible. On the other hand, if, as is usually the case, there are many pips visible each radar scan, or if the radar strobe is on some of the time, this streaking effect is likely to be objectionable. By reducing the size of the persuader aperture area by a factor of five in later model tubes this effect has been reduced in about the same ratio.

Signal to Noise Ratio Discrimination

It was found in operation that the storage orthicon could effectively reproduce PPI information even when the signal-to-noise ratio of the latter was quite low. In some tests made* to ascertain what minimum signal-to-noise ratio could be transmitted, it was found that ratios below about 1:1 could still be reproduced. Signals that could not be discerned visually from a noise background on a PPI screen, could be successfully picked up by the storage orthicon and then viewed on a Kinescope screen.

Resolution

Under carefully set-up laboratory conditions, resolutions greater than one thousand television lines have been observed.

Conclusions

The operating characteristics of the storage orthicon tube have been described with particular reference to its application to Teleran.

* These tests were carried out by G. L. Fernsler and H. Kihn of RCA Laboratories Division.

These characteristics include optical insertion of the signal, storage and continuous reproduction of the signal for tens of seconds (see Figure 9) and discrimination against noise equal to or better than that of the eye in viewing PPI presentations directly. Other applications of the tube, for example, for the purpose of obtaining bright radar pictures, large screen presentation and remote transmission are clearly suggested.

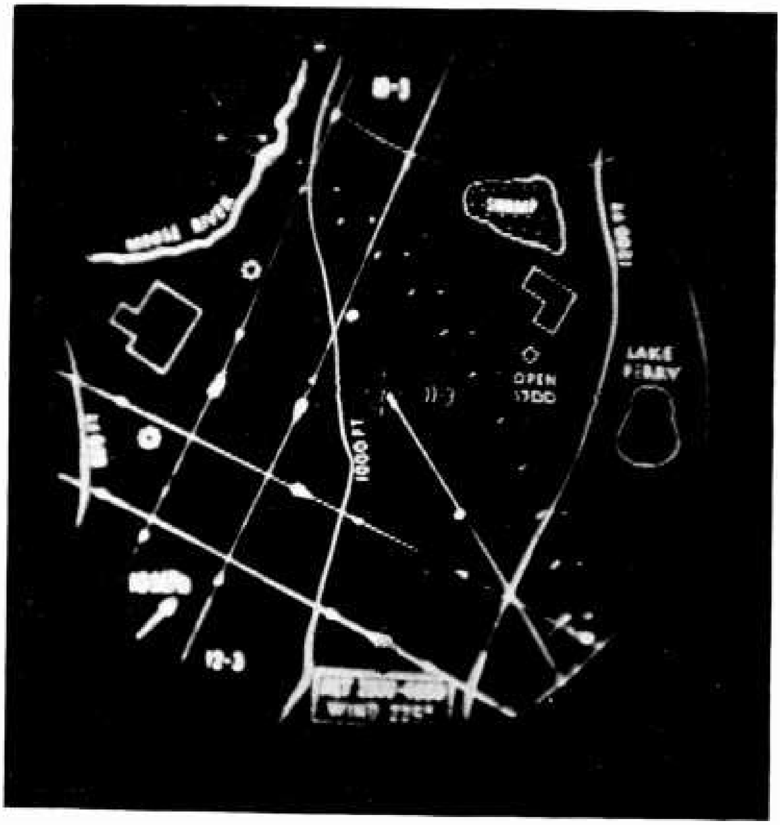


Fig. 9—PPI information and optically superimposed aerial map simultaneously picked-up by a Storage Crthicon.

ACKNOWLEDGMENTS

The writer is especially indebted to A. Rose and H. B. Law for their valuable contributions to the development of this tube. Some of the results reported here are based on tests carried out in cooperation with G. L. Fernsler and H. Kihn. The writer also wishes to express his appreciation for the support and encouragement of V. K. Zworykin, under whose direction the work was carried out.

INTERLOCKED SCANNING FOR NETWORK TELEVISION*

By

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Summary—The benefits of operating the scanning systems of two or more independent television broadcasting plants in locked coincidence are discussed. The problem of producing locked coincidence is explored, and methods of achieving the desired result are indicated. Some of the possible benefits of using stable (high inertia) frequency sources for scanning systems are noted.

BENEFITS OF INTERLOCKED SCANNING

THE desirability of maintaining continuity of synchronizing information on the television broadcast signal was recognized by the designers of early television plants, and the plants were so arranged that no interruption of the synchronizing signal occurred while switching between cameras or between local studios. However, no means for maintaining this continuity of synchronizing signal has evolved as yet for operational use when switching between local studios and remote pickup points or network programs where independent synchronizing signal generators are involved. To render less objectionable the attendant loss of synchronizing signal at receivers when such switches are made, it is usual practice to go to a dark screen before the switch is made and to fade up from a dark screen after the switch is completed. This allows time for synchronizing circuits in the receivers to lock in on the synchronizing signal transmitted after the switch before a picture reappears on the receiver, and hence the resultant disturbance of the image due to the momentary loss of the synchronizing signal is less noticeable. As an additional aid in bridging this gap in synchronizing signal continuity, it is good practice to have the 60-cycle components of the two synchronizing signals involved phased for approximate coincidence.

If complete time coincidence between the two signals (local and remote) could be maintained, it would be possible to preserve continuity of the transmitted synchronizing signal and thereby eliminate the necessity of going to a dark screen during switching, or, of checking and maintaining the vertical phasing of the two signals. From

* Decimal Classification: R583.13.

the standpoint of smooth program presentation, this is of importance because it is frequently necessary to switch between the local and remote pickups several times in the course of one program, particularly when presenting commercial announcements. In addition, if the local synchronizing signal is transmitted continuously on either the local or remote picture signal (this requires processing the remote signal to remove its synchronizing signal and adding the local synchronizing signal) the receivers are fed a relatively more noise-free synchronizing signal. Also made available between the local and remote signals are lap dissolves, super-impositions and all other processes normally available between local cameras or studios. As network television broadcasting grows, the foregoing aids to smooth presentation of programs assume increasing importance. Therefore, an analysis of what is involved in producing locked coincidence between a local and a remote signal is timely.

Requirements for Producing Interlocked Scanning

Complete coincidence is required between the local and remote signal to achieve the above benefits, and coincidence must therefore be on a line, field, and frame basis. This means that even fields must coincide with even fields, etc., and in the final analysis that each line of the 525 lines in a frame of one signal must coincide with its counterpart in the other signal. This complete matching is required before the local synchronizing signal can be used on the incoming signal and comply with FCC standards of transmission. If even and odd fields are matched and line blanking in the two signals coincide, the local synchronizing signal would align with the incoming signal as shown in Figure 1, where one line frequency synchronizing signal pulse (local) rests on the vertical blanking (remote), or one equalizing (2 times line frequency) pulse (local) rests on the last horizontal blanking (remote), or is lost in the video of the last line of alternate fields. When the required complete coincidence is obtained, the lock applied to maintain the coincidence must be quite rigid. Any hunting permitted by the lock would: first, render impossible the use of local synchronizing signal on the remote picture signal and hence, continuity of synchronizing signal transmission; and second, render ineffective the use of superimpositions and lap dissolves. The only gain in using a loosely-locked coincidence between the local and the remote signal lies in the fact that the vertical components of the two signals will remain approximately in phase and therefore the tendency for vertical scanning at receivers to lose synchronization momentarily following a switch will

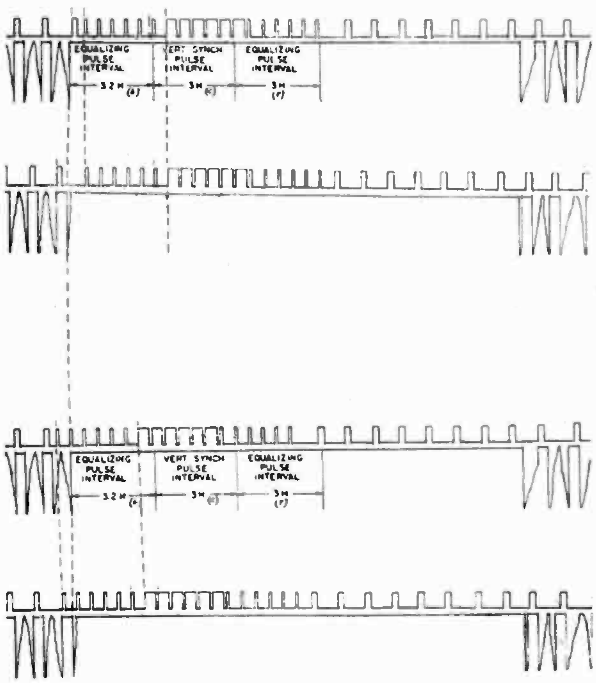

Fig. 1—Local synchronizing signal superimposed on a remote video signal showing the two possible conditions, both mis-matches, which can obtain when even vs. odd fields are locked in coincidence.

be reduced. Figure 2 serves to illustrate the coincidence required to achieve the desired results.

Methods of Achieving Interlocked Scanning

The block diagram of Figure 3 is an arrangement which has been used in "On the air" demonstrations of the operation of two independent television plants under conditions of "locked coincidence." While admittedly an experimental arrangement, it served to confirm the

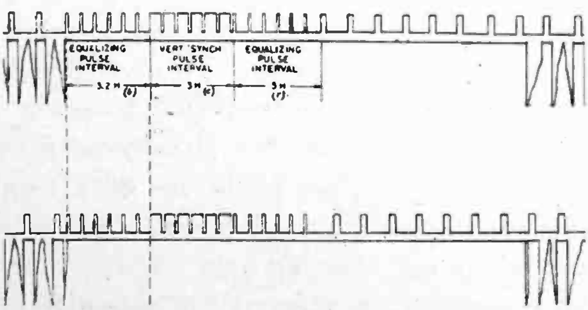

Fig. 2—Local synchronizing signal superimposed on a remote video signal showing the desired result which obtains when even vs. even fields are locked in coincidence.

possibility of realizing the benefits mentioned heretofore, and also to demonstrate effectively the program possibilities during the hours when election returns were coming in for a fall election, 1941. The regularly scheduled program on that occasion was wrestling from a sports arena via the Telemobile Unit. In the Radio City studios a camera was focused on a black-board upon which election returns were recorded. By using only the top of the black-board and by keeping the top of the picture from the Telemobile Unit relatively clear of action (normally the case) a superimposition of the election results upon the incoming sports picture provided the latest election returns without interrupting the sports event.

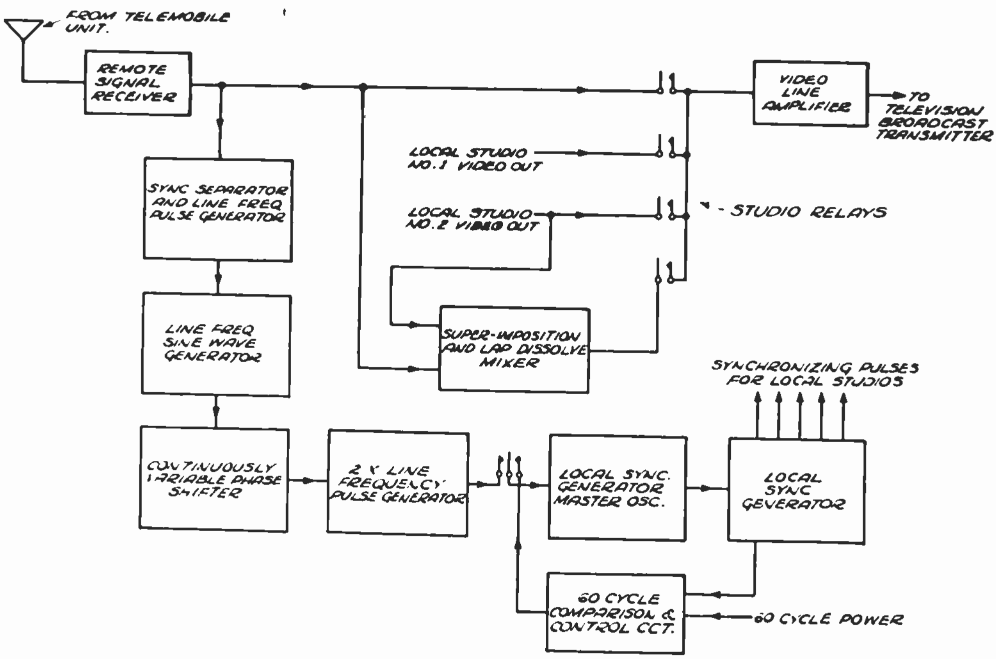

Fig. 3—Block diagram of equipment used to lock the local synchronizing signal generator with the incoming remote video signal.

Figure 4 is a schematic of the block diagram of Figure 3. The synchronizing signal separator and line frequency pulse generator are omitted, the former consisting merely of a conventional synchronizing signal separator, which drives a line frequency blocking oscillator. The line frequency sine wave generator has already been described.[1] The continuously-variable phase shifter made use of a rotating magnetic field and a pickup coil whose physical position could be advanced or retarded without limit in the rotating field. At the time of the demonstration the most convenient equipment for accomplishing this result

[1] R. A. Monfort and F. J. Somers, "Measurement of the Slope and Duration of Television Synchronizing Impulses," *RCA REVIEW*, Vol. VI, No. 3, pp. 370-389, January, 1942.

was a small Selsyn unit. The circuit for deriving the three-phase excitation for the Selsyn from the single phase output of the sine wave generator is shown in Figure 4. The resistance-capacitance components on the first grid of two of the legs are chosen to get the desired shift as indicated by the vectors. Following the continuously variable phase shifter the sine wave is processed to provide the locking information to the master oscillator of the local synchronizing generator. As stated before, the lock must be quite rigid before the desired benefits can be realized. In the demonstration referred to previously, the local synchronizing generator was a standard commercial unit.[2] The master

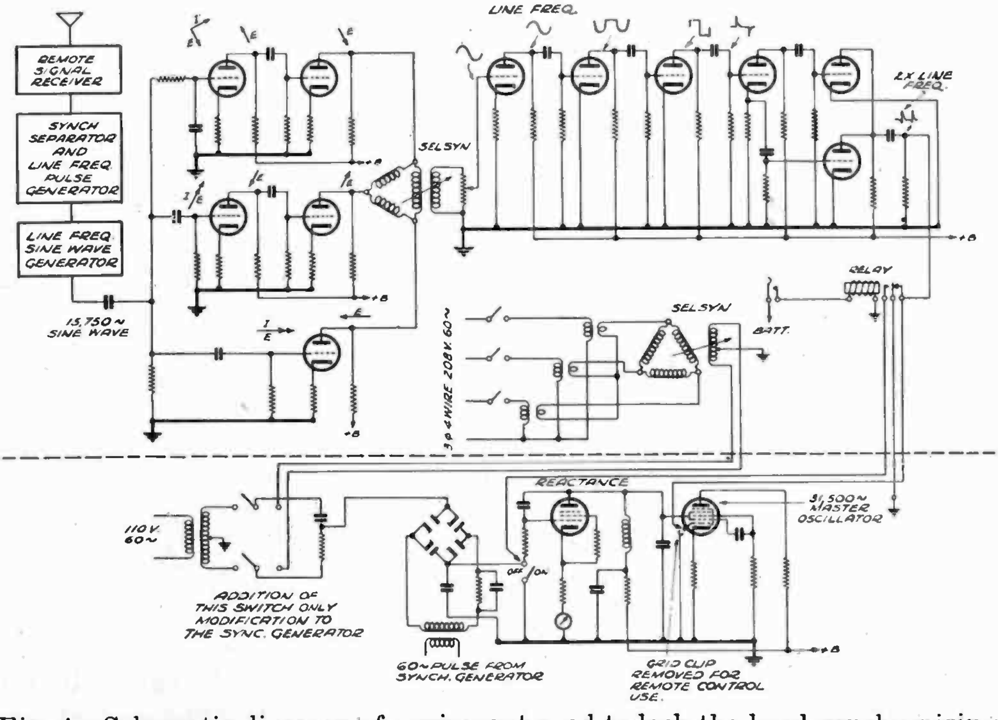

Fig. 4—Schematic diagram of equipment used to lock the local synchronizing signal generator against the incoming remote video signal.

oscillator in this unit is of the negative transconductance type and operates at twice line frequency. By trial and observation it was determined that control or lock of the desired degree of rigidity could be obtained by injecting a pulse of twice line frequency into the first grid of this master oscillator. Therefore, for the demonstration the processing required was the conversion of the line frequency sine wave output of the continuously variable phase shifter into twice line frequency pulses. This was accomplished by the symmetrical clipping of the sine wave to produce a symmetrical square wave which was in turn differentiated. A push-pull input into grid current biased grids

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[2] A. V. Bedford and J. P. Smith, "A Precision Television Synchronizing Signal Generator", *RCA REVIEW*, Vol. V, No. 1, pp. 51-68, July, 1940.

of two tubes, the plates of which were in parallel, provided the desired result. An alternative and equally effective method would be the use of an unfiltered output from a full wave rectifier driven by the continuously-variable phase shifter.

The method of coupling this double frequency pulse into the master oscillator and of transferring the control of the local master oscillator from the 60-cycle power frequency to the incoming video signal is also indicated in Figure 4. No major modification of the local synchronizing generator was required, three clip leads from a relay clipped on at the proper points and the lifting of the grounded grid cap to the first grid of the master oscillator sufficing. However, it will be noted that the 60-cycle power into the comparison circuit which normally controls the master oscillator of the local generator was modified by the insertion of a continuously variable phase shifter. The same style Selsyn was used for both the 60-cycle and 15750-cycle phase shifters. The reason for this modification to the local generator is apparent when the task of securing coincidence between the two signals by the use of the line-frequency phase shifter alone is considered. A maximum of approximately 262½ revolutions of the line frequency phase shifter may be required if the control should be transferred to the remote signal when coincidence between an even field of one signal existed with an odd field of the other signal. It is much faster to use the 60-cycle phase shifter for rough setting of coincidence, transfer control, and finish the exact alignment of coincidence between the two signals by means of the line-frequency phase shifter.

For checking coincidence, the pulse cross unit, which was described in the paper on the sine wave generator,[1] provides an effective indicator. The local and incoming signals are mixed and applied to the pulse cross monitor and the phase shifters (60 and 15750 cycles) are adjusted to secure coincidence of the two signals on the pulse cross. Figure 5 shows blanking from one signal generator and synchronizing signal from another. There is lack of coincidence in terms of both line and field. Figure 6 shows coincidence for line but not for field. Figure 7 shows odd vs. even field coincidence. Note that one line frequency synchronizing signal pulse is sitting on the field frequency blanking. Figure 8 illustrates the same condition except the line frequency phase shifter has been rotated one revolution from the condition of Figure 7. Here an equalizing pulse is resting on the last line frequency blanking preceding field blanking. Figure 9 shows even field vs. even field coincidence.

Figure 9 is the same as Figures 7 and 8 except one signal is shifted through 262½ lines. Any hunting between signals is readily discerned

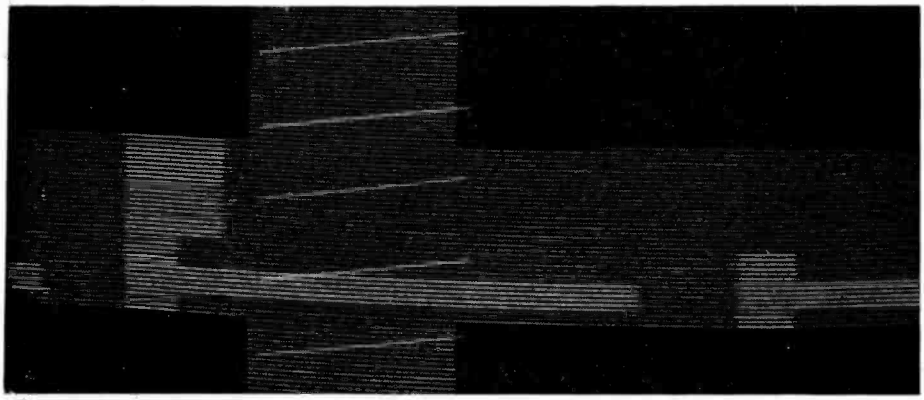

Fig. 5—Pulse cross pattern photograph showing complete lack of coincidence.

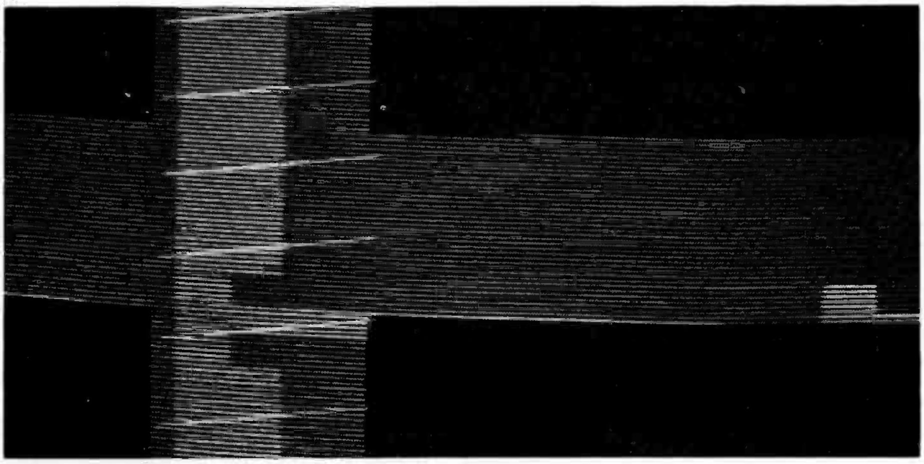

Fig. 6—Pulse cross pattern photograph showing coincidence at line frequency but not at field frequency.

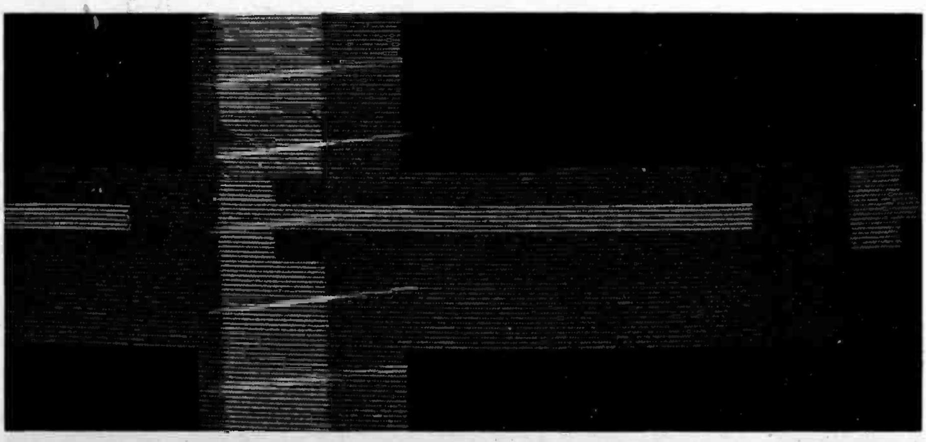

Fig. 7—Pulse cross pattern photograph showing odd vs. even field matching (the last line frequency synchronizing signal pulse is superimposed on the first line of the field frequency blanking.)

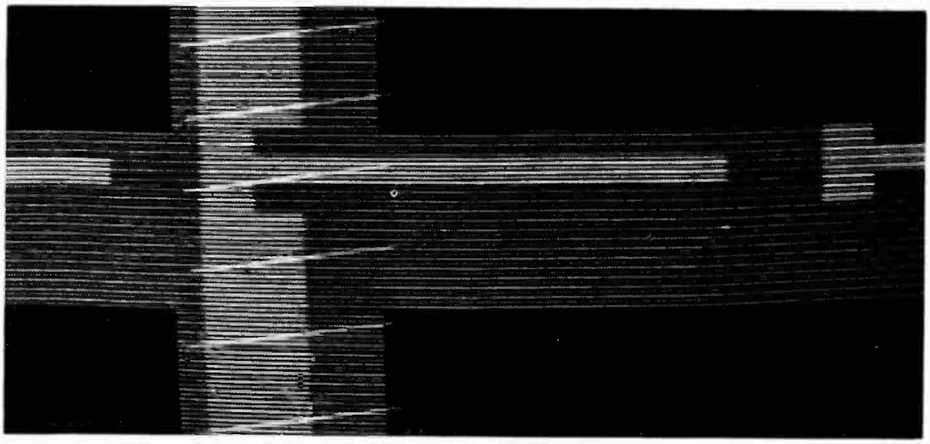

Fig. 8—Pulse cross pattern photograph showing odd vs. even field matching (the first equalizing pulse is superimposed on the last line frequency blanking pulse.)

on the pulse cross. Variations in the width of the "front porch," or delay of line-frequency synchronizing signal with respect to line-frequency blanking, are easily noted as shown in Figure 10.

The method and equipment outlined function satisfactorily and do not require excessive time in aligning the two signals for coincidence provided the continuity and stability of the received signal are good — i.e., provided one alignment will suffice for the transmission. For regular operational use, a means of quickly reverting to independent systems would be mandatory to cover the contingency of momentary interruption in the incoming signal. A means of automatically establishing coincidence and lock between the two signals as well as reverting to independent operation would of course be a highly desirable feature.

The results obtainable from interlocked scanning will be enhanced by the use of synchronizing generators that are not locked to local

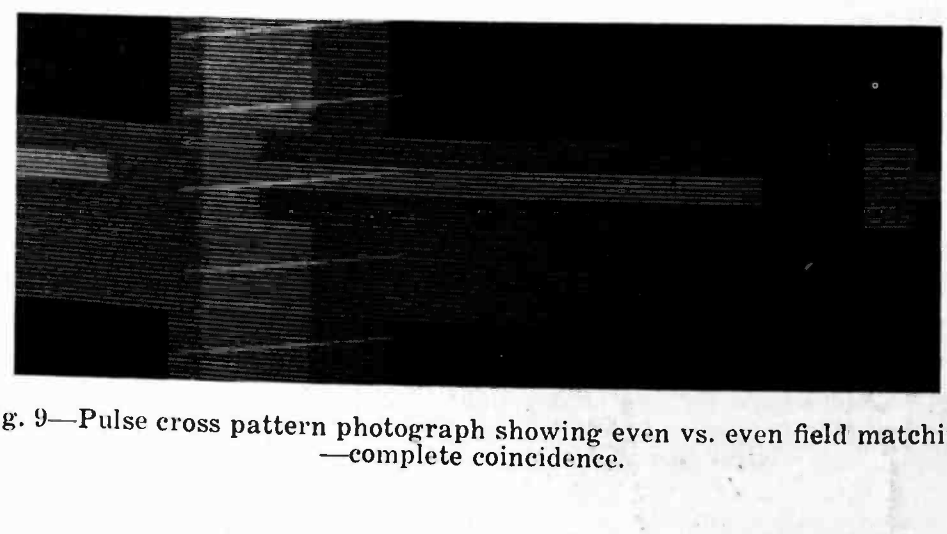

Fig. 9—Pulse cross pattern photograph showing even vs. even field matching —complete coincidence.

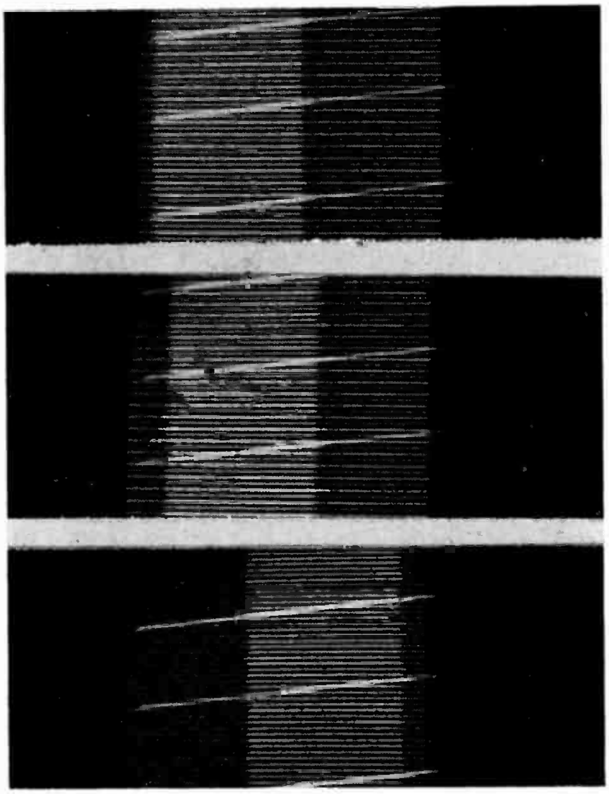

Fig. 10—Portions of pulse cross patterns showing variations in "front porch" or in the delay of line-frequency synchronizing signal with respect to the line-frequency blanking signal.

60-cycle power supplies, but which are controlled by high-inertia frequency sources. The most obvious benefit from the use of high-inertia systems lies in the fact that a momentary loss of the incoming signal would not ordinarily produce the same discontinuity of control of interlocked scanning that is inevitable with the system described. In fact, it appears possible to provide high-inertia controls for individual synchronizing generators of such excellence that appreciable discontinuity in incoming signal can be tolerated before, from an operational point of view, it would be necessary to sever the interlocked scanning tie-in.

A disadvantage in using the high-inertia frequency control system for synchronizing generator control lies in the fact that projector motors in film studios could no longer be synchronously driven from the local 60-cycle power.

Conclusions

The general direction of work toward one solution to the problem of interlocked scanning systems has been indicated. The work done has served more to show the nature and magnitude of the problems involved than to provide a complete answer. The use of the incoming signal for control is indicated by the economics of the problem.

PERFORMANCE CHARACTERISTICS OF LONG-PERSISTENCE CATHODE-RAY TUBE SCREENS; THEIR MEASUREMENT AND CONTROL*†

BY

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Summary—Cathrode-ray tube indicators employing long-persistence screens have been developed for use in Radar, Teleran, and other systems where field-repetition rate is in the order of seconds. Tube performance in such applications is dependent upon the phosphorescent intensity of the screen following a given excitation. This parameter in turn is dependent upon the fluorescent intensity, rate of build-up and rate of decay of the phosphor components. In this paper, a laboratory system for pulse-exciting long-persistence cascade-type cathode-ray tube screens and quantitatively evaluating these charactertistics in finished tubes is described. Correlation of these values with field performance is illustrated.

A range of efficiencies of screen components was measured under steady-state conditions in the conventional manner and compared with their performance in cathode-ray tube screens under pulsed excitation. It is shown that the efficiency of the blue zinc-sulphide phosphor layer is a good indication of its performance but that the efficiency of the yellow zinc-cadmium-sulphide phosphor layer (phosphorescent) phosphor bears no consistent relation to ultimate screen performance. A method for pulse-exciting the phosphor with blue light which gives values correlating with screen performance is described.

The thickness of each phosphor layer and the exhaust-bake temperature employed in tube processing have been found to affect screen characteristics appreciably. Average curves are presented to show the effect of each variable.

INTRODUCTION

LONG-PERSISTENCE screens were developed for and used extensively in radar indicators during World War II. This paper, therefore, necessarily refers chiefly to radar requirements although, basically, it is of interest in current cathode-ray tube applications such as oscillography and in the projected development of peacetime marine radar and Teleran.

It is now well known that the means for converting the information carried by the radar pulse to an interpretable medium has been supplied

* Decimal classification: R138.313 × R200.

† This paper is based in whole or in part on work done for the Office of Scientific Research and Development under contract OEMsr-441 with Radio Corporation of America.

by the cathode-ray tube. When the beginning of an oscilloscope sweep is synchronized with the release of energy from the transmitter, a "pip" can be made to appear on the sweep after a time interval that is directly proportional to target distance. The scan just described is the type "A" scan and is all that is required for accurate ranging. However, modifications were introduced in the form of type "B", PPI, and Eagle scans designed to extend the usefulness of radar. In contrast to the multi-kilocycle repetition rate of the type "A" scan, the recurrent scan periods associated with these other displays varied from 0.2 seconds to as much as 30 seconds. Conventional P1 and P4 screens in popular use for oscillography and television are highly efficient but have short persistence, decaying to less than one percent of their peak brightness in 0.06 second.[1] These qualities make them well-suited to the "A" type scan but virtually useless in exploiting the advantages of the later types. Therefore, a long-persistence phosphor or phosphor combination of relatively high efficiency was needed to track targets and furnish the map-like displays characteristic of the PPI and Eagle scans.

For high luminescent efficiency, the decay for the ideal phosphor would be similar to curve *A*, Figure 1, where high phosphorescence is retained for the complete radar scan period at the end of which time it effectively disappears. No luminescent material has been found with that characteristic or even with the concave downward characteristic of curve *B* which approaches it in the practical sense. H. W. Leverenz[2] has demonstrated that such decay is unattainable on the basis of the classical theory of luminescence. However, much effort had been concentrated on high-efficiency sulphides, primarily for television applications, during the period of 1935 to 1940.[3-11] Two of these sul-

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[1] R. T. Orth, P. A. Richards and L. B. Headrick, "Development of Cathode-ray Tubes for Oscillographic Purposes", *Proc. I.R.E.*, Vol. 23, No. 11, November, 1935.

[2] H. W. Leverenz, "Cathodoluminescence as Applied to Television", *RCA REVIEW*, Vol. V, No. 2, pp. 131-175, October, 1940.

[3] T. B. Perkins, and H. W. Kaufman, "Luminescent Materials for Cathode-ray Tubes", *Proc. I.R.E.*, Vol. 23, No. 11, November, 1935.

[4] Frederick Seitz, "Interpretation of the Properties of ZnS Phosphors", *Jour. Chem. Phys.*, Vol. 6, No. 8, August, 1938.

[5] S. T. Martin and L. B. Headrick, "Light Output and Secondary Emission Characteristics of Luminescent Materials", *Jour. Appl. Phys.*, Vol. 10, No. 2, February, 1939.

[6] W. DeGroot, "Luminescence Decay and Related Phenomena", *Physica*, Vol. 6, No. 3, February, 1939.

[7] R. B. Nelson, R. P. Johnson and W. B. Nottingham, "Luminescence During Intermittent Electron Bombardment", *Jour. of Appl. Phys.*, Vol. 10, pp. 335-342, May, 1939.

(References continued on next page)

phides were subsequently combined in an ingenious manner to form the P7 and P14 type cathode-ray tube screens that performed creditably over the relatively wide range demanded by radar. A typical decay curve for these screens during the first scan period is shown as curve C in Figure 1. During the second scan period, the phosphorescent level rises as shown by the curve C' and continues to rise each subsequent scan period until an equilibrium condition is reached.

THE P7 SCREEN

The P7 screen is a cascade or two-layer screen and is different in this respect from screens generally employed in cathode-ray tubes.

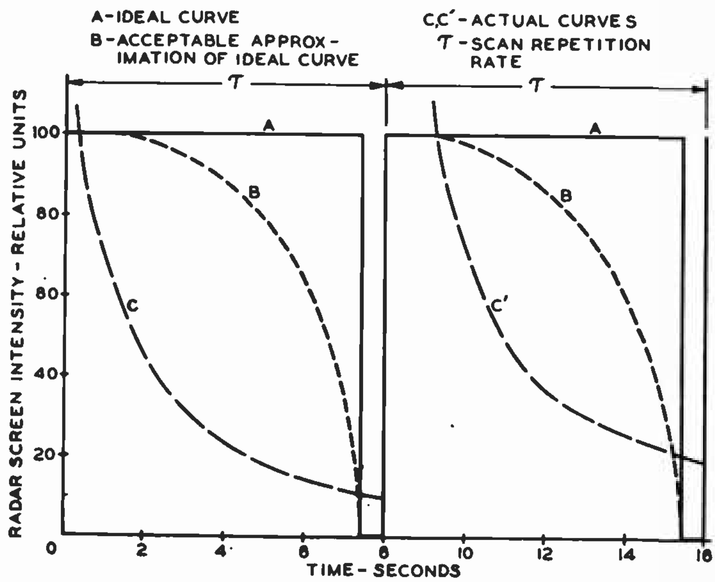

Fig. 1—Decay curves for radar indicator screens.

It consists of a layer of yellow-phosphorescing zinc cadmium sulphide (hex.—9ZnS.CdS:Cu(0.007)) *applied directly to the bulb face* and a layer of blue-fluorescing zinc sulphide (hex.—ZnS:Ag(0.015)) *on top of the yellow layer*. When the phosphor layers are applied properly, the separation of the layers, necessary for good screen performance, can be achieved. The P14 is fabricated in a similar manner but employs

[8] W. B. Nottingham, "Electrical and Luminescent Properties of Phosphors Under Electron Bombardment", *Jour. Appl. Phys.*, Vol. 10, pp. 78-83, 1939.

[9] R. P. Johnson, "Luminescence of Sulphide and Silicate Phosphors", *Jour. Opt. Soc. Amer.*, Vol. 29, pp. 387-391, 1939.

[10] W. DeGroot, "Miscellaneous Observations on Rise and Decay of Luminescence of Phosphors", *Physica*, Vol. 7, pp. 432-446, May, 1940.

[11] H. W. Leverenz, "Optimum Efficiency Conditions for White Luminescent Screens in Kinescopes", *Jour. Opt. Soc. Amer.*, Vol. 30, No. 7, pp. 309-315, July, 1940.

a zinc cadmium sulphide whose spectral emission characteristic is shifted toward the red. Use of this phosphor produces an accelerated decay during the first 0.1 second after excitation. Beyond one second, decay proceeds at approximately the P7 rate but at somewhat lower intensity so that it appears to have shorter persistence.

In operation, the layer of zinc sulphide is excited to blue fluorescence by electron bombardment. During this period, the yellow layer absorbs and stores much of the blue-light energy generated. When bombardment ceases, the fluorescence subsides immediately while the yellow layer releases its energy as an intense phosphorescent glow. Thus, the image traced by the electron beam is retained for some time after the signal has ceased.

If the two layers become mixed, the layer of zinc cadmium sulphide will be subjected to direct electron bombardment and attendant heating of the phosphor. Under this mode of excitation, decay is greatly accelerated and the absolute level of persistent light at any given time after excitation will be greatly reduced. In effect, the whole purpose of the cascade principle, permitting deeper penetration of the lattice structure by ultra-violet energy as contrasted to the low penetrating power of electron excitation, is defeated.

It is to be noted that the storage property is cumulative. Therefore, if the screen possesses some energy at the time of excitation, the newly acquired energy will ordinarily add to that already possessed to result in a higher phosphorescent level at any time after excitation than would prevail if the phosphor were excited from zero energy level. A notable exception would occur when the excitation is attended by excessive heating of the phosphor sufficient to accelerate its decay and result in decreased phosphorescent intensity at some given time after excitation. In normal operation, successive pulses of moderate strength and in fairly rapid succession, such as are encountered in radio detection and ranging, raise the signal brilliance to a level much higher than a single pulse and afford a distinct advantage over any single-layer screen of comparable efficiency not having a build-up characteristic.

From the foregoing discussion the following are attributes for good screen performance:

(1) the absolute level of the persistent light at a given time after a single pulse excitation should be as high as possible;

(2) the screen should show a good cumulative increase in light level on repeated excitation; and

(3) the conversion efficiency of the screen, i.e., the ratio of inten-

sities of the phosphorescent light to the fluorescent light producing it, should be as high as possible.

It is also necessary that the phosphorescence does not persist long at a level sufficiently high to interfere with information presented in subsequent sweeps. However, since carry-over has not been a problem in the course of this work, it is not included as one of the characteristics to be evaluated.

METHOD FOR OBJECTIVE TESTS ON P7 SCREENS

To evaluate these attributes in any particular P7 screen, it would basically be necessary to pulse a finite area of the screen with an electron energy density comparable to that encountered in the field.

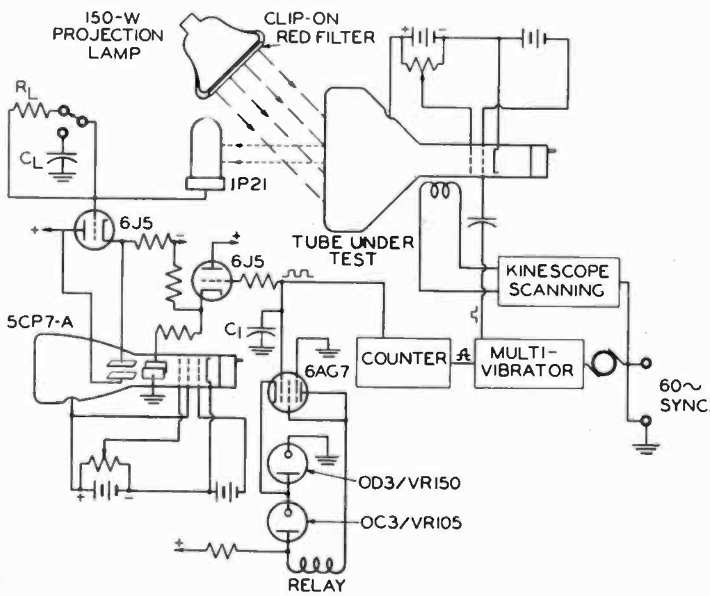

Fig. 2—Schematic diagram of P7 screen test equipment.

This pulse should then be repeated in this case at the radar scan interval and continued at this rate until either the target progresses to a new location or the screen reaches a condition of equilibrium. Regarding the latter instance, a point will be reached on pulse operation where the energy input to the screen per pulse will just restore that lost by decay between pulses. With this broad principle in mind, the following test procedure was developed and adopted for rating P7 and P14 screens.

Test Procedure and Equipment

By means of equipment illustrated in Figure 2, quantitative measurement of P7 and P14 characteristics is made with voltages on the cathode-ray-tube elements equal to those employed in field operation. The tube to be rated is set up under steady-state conditions and a 60-microampere beam is deflected into a 7×7 centimeter raster on the P7 screen. Interlaced scanning having a frequency of 13,230 cycles per second horizontally and 60 cycles per second vertically is employed. The tube is then biased off, usually by -25 volts, and the screen reduced to approximately zero-energy level. This reduction is accomplished by irradiating the screen with red and near-infrared light energy which is supplied from a 150-watt projection lamp operating through an Electrolyte footlight red filter. A series of square-topped pulses of 1/60-second duration are then generated by the multivibrator and impressed on the grid of the cathode-ray tube. The pulse amplitude is equal to the bias selected above for the tube but of opposite polarity so that with each pulse the tube is fired to the 60-microampere beam level established under steady-state conditions. The multivibrator is normally biased off or blocked between pulses and is triggered by the 60-cycle synchronizing pulses used to lock in the scanned raster on the cathode-ray tube. The sync pulses are admitted to the multivibrator circuit only once each second through rotating contacts on a synchronous motor. In this manner, driving pulses of 1/60-second duration are impressed on the cathode-ray tube once each second and exactly 1/2 of the interlaced raster is scanned on the screen once each second at the predetermined beam intensity. The result is a periodic pulse excitation of the P7 screen. The relation of this pulse excitation to the intensity and duration of the "average" radar signal is quite definite and will be discussed later.

Between pulses, for 59/60 of each second, the phosphor will decay in a manner illustrated by curve C of Figure 1. A type 1P21 multiplier phototube, positioned as shown in Figure 2, responds continuously to the light output from the P7 screen. The signal is impressed on the deflection electrodes of a type 5CP7-A oscillograph tube through a resistive load so that vertical deflection of the spot is proportional to screen brightness. The multivibrator pulses operating through the counter circuit provides the horizontal sweep for the oscilloscope. Each pulse is amplified and impressed successively on a capacitor in such a manner that (1) each increment of charge displaces the spot on the oscillograph tube a desired amount horizontally and (2) the rate of accumulating charge can be adjusted to operate a relay when the desired number has been delivered.

The pattern traced by the oscilloscope while a screen is being rated will resemble the solid lines in Figure 3. As the first pulse is delivered, the screen is excited to high fluorescence. Since the oscilloscope is adjusted to give a readable displacement under the relatively weak phosphorescent light, the spot of the oscillograph tube is deflected off scale at this time. As the phosphor decays, however, the spot will return to the face of the oscillograph tube and drop slowly as decay proceeds. This is represented by the first vertical line in Figure 3. Thus, it is seen that the height of the spot above the sweep line is proportional at any time to the brightness of the screen under test.

When the second pulse is applied to the tube under test, the spct will again be deflected off the face of the oscillograph tube and the

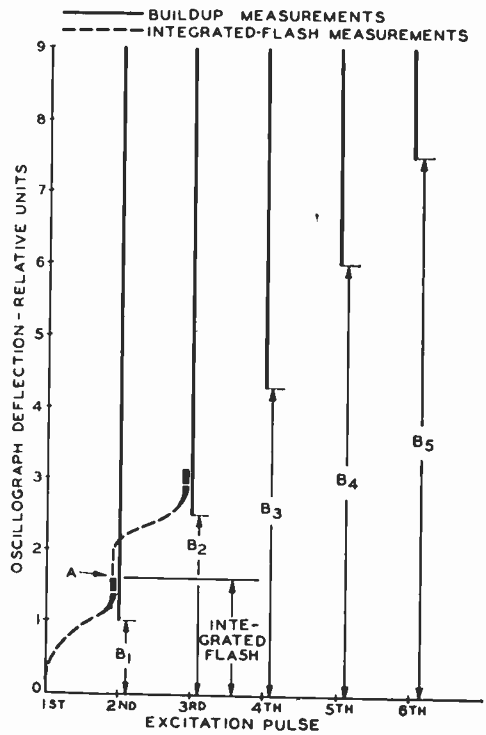

Fig. 3—Pattern traced on the oscilloscope screen for measurements of build-up and integrated flash.

cycle repeats as additional pulses are delivered. The lowest point that the spot reaches just before a subsequent pulse drives it off scale, is a measure of the intensity of the persistent light of the screen at this time and is identified as B_1 in Figure 3.

Because the standard pulse is repeated at one-second intervals, a series of brightness measurements such as B_2, B_3, B_4, etc. are automatically recorded as noted in Figure 3. A quantitative indication of the cumulative increase in phosphorescence is then given by the numerical ratios B_5/B_1, B_{10}/B_1, etc. Under the mode of excitation selected, the screens approach a state of dynamic equilibrium soon after the fifteenth pulse when the energy imparted per pulse is just sufficient to restore that lost by decay between pulses. The ratio of B_{10}/B_1 has also been found to be proportional to B_5/B_1, so that it became customary to rate tubes on the basis of the lesser number of pulses. Therefore, the following parameters have been set up as standard in rating long-persistence P7 screens:

B_1—the absolute level of persistent light intensity 59/60 of a second after excitation from near zero-energy level by a single, standard pulse;

B_5—the absolute level of persistent light intensity 59/60 of a second after excitation by the fifth pulse in a series of standard pulses.

The brightness is measured in milli-foot-lamberts. The cumulative increase in the level of the persistent light is denoted by the buildup ratio (*BR*), calculated directly from B_5/B_1.

When the resistive load is replaced with a capacitive load and the procedure repeated, an electrical charge proportional to the total light emitted, both fluorescent and phosphorescent, will be stored in the capacitor. Deflection of the oscillograph tube is now proportional to the total light output integrated over one second and the "flash" measurement is obtained from the oscilloscope as indicated by the broken line of Figure 3. In this case, however, the gain is adjusted so that deflection is on the face of the oscillograph tube at all times since it is desired to read the maximum deflection values. The point "A" at which the measurement is taken is readily noted in practice by the change in speed of the spot as the second pulse comes in from the multivibrator.

Because point "A" represents an integration (photo-coulombs), and a value assigned to B_5 represents a rate (photo-coulombs per unit time), the ratio of "integrated flash" to B_5 is not a pure numeric as is the ratio B_5/B_1. One can, however, estimate the averaged "rate" during the flash by assuming that most of the charge had accumulated well within the 1/10-second period (approximately) over which the human eye tends to average. If V_c is the voltage across the input capacitance C, the charge accumulated is

$$Q = V_c C$$

If, however, a voltage V_R is applied across an input resistance R, the charge passed during 1/10-second is

$$Q = 0.1 \frac{V_R}{R}$$

If the Q's are set equal, the current in the second instance is equal to the average current flowing to the capacitance during the 1/10-second observation interval. It is to be noted however that for equal signals or equal visible screen stimuli, when, for example, $R = 2$ megohms and $C = 2$ microfarads, then $V_R = 40\ V_C$. In other words, to derive a numeric roughly indicative of the subjective comparison by the human eye of the "flash" to the phosphorescence, the input voltage represented by point A must be multiplied by 40 for comparison with the input voltage represented by B_5. Similar calculations can be made for any particular combination of calibrating resistance and capacitive load.

Relation of Test Conditions to Field Operation

The pulse duration and repetition rate for the standard test procedure were selected to approximate the "average" conditions encountered in field operation. The necessary equipment was designed to use those facilities one might normally expect to find in a cathode-ray tube manufacturing plant. Beyond this, test conditions have been set to judge the more critical instance or the limit of acceptability of the P7 screen. Calculations have been prepared and tabulated to demonstrate the relation existing between test and field conditions.

The pulse time per screen element (t) is given for any cathode-ray tube screen by the expression

$$t = \frac{d\,t'}{l} \text{ microseconds}$$

where $d =$ diameter of the cathode-ray beam in millimeters;

$l =$ length of trace on cathode-ray screen in millimeters and the equivalent of maximum range;

$t' =$ time in microseconds required to scan the trace (not including retrace) and is identical with the transit time of the pulse at maximum range.

These parameters in turn are dictated by the resolution capabilities of the tube, the screen diameter, and the range of the radar, respectively. Under the test conditions these same parameters are determined by the type of tube, raster width, and the scanning frequency. Table I is a tabulation of pulse time per picture element for the more common sizes of "scopes" employed and for a number of practical radar ranges.

Table I—Calculations of pulse time per picture element under field and test conditions.

Field Conditions

Useful Diameter of Radar 'Scope (inches)	Spot Size (millimeters)	Radar Range (miles)	Approximate Pulse Time per Element (microseconds)
10	1.00	100	4.0
		50	2.0
		10	0.5
7	0.75	100	5.0
		50	2.5
		10	0.5
5	0.50	100	5.0
		50	2.5
		10	0.5

Test Conditions

Pattern Width (millimeters)	Spot Size (millimeters)	Number of Lines	Approximate Pulse Time per Element (microseconds)
70	0.50	220	0.5
	1.00	220	1.0
35	0.50	220	1.0
	1.00	220	2.0
20	0.50	220	2.0
	1.00	220	4.0

Similar calculations are presented for the test conditions. In this instance, however, it has been found possible to retain a constant scan frequency (RMA standard television frequency which is generally readily obtainable in most cathode-ray tube plants) and achieve a range of pulse excitations by varying the length of the trace. It is to be concluded from these data that, at the frequency selected, raster sizes may be chosen which can be accommodated by most 'scopes and which will duplicate most of the pulse rates encountered in field operation.

However, energy imparted to the screen depends on the intensity of the excitation as well as on the duration. It has been stated from observations made in the field that radar signals range from a weakest usable signal of 10^{-11} coulombs/square millimeter to 2×10^{-9} coulombs/square millimeter for the best signals.[12] Assuming an average

[12] See following page.

pulse duration of one microsecond (justified from Table I), the instantaneous peak current to the screen would vary from 2 to 400 microamperes. From additional field observations the "average" signal can be expected to lie in the range of 60 to 100 microamperes. This range can easily be attained in the conventional radar 'scopes and the anode voltage which determines the final velocity of the beam can obviously also be duplicated. In view of this situation, the only decision required was to determine the conditions to be selected for the standard test. Since it would actually be impossible to predict the conditions at which a particular tube would be operated, not the least of the reasons being that it would probably be operated under all of

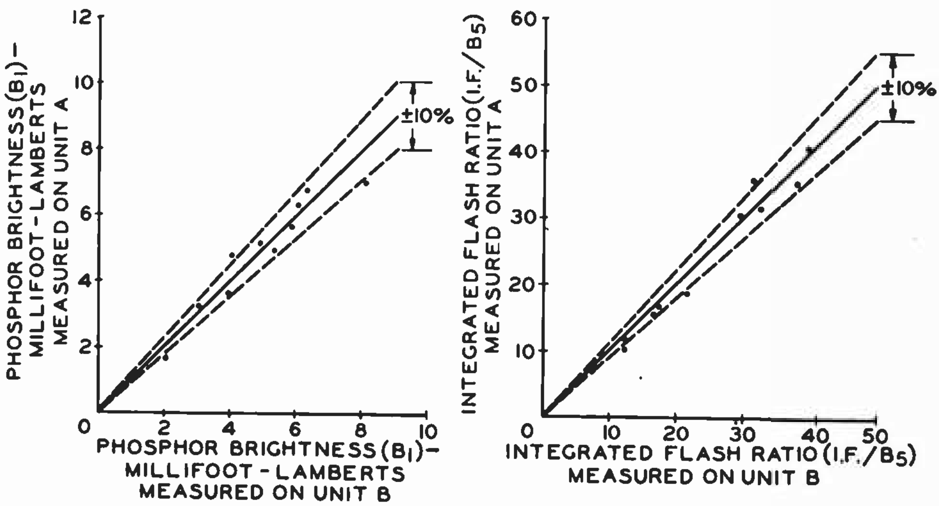

Fig. 4—Relative accuracy obtained with different persistence-measuring units.

them, the obvious solution was to choose those conditions which duplicated the more critical range of 10 to 30 miles (where tracking and ranging are most closely followed) and the weaker end of the still very usable signal. Hence, the choice was made of a 70-millimeter raster and a 60-microampere beam in the test procedure.

Accuracy of Measurement

The curves of Figure 4 indicate the accuracy to be attained with the persistence-measuring unit. The data presented are from different units and the measurements are made by different operators in order to take into account all possible sources of error. Identical measurements on each unit would cause all points to fall on the 45-degree line

[12] H. W. Leverenz, "Final Report on Research and Development Leading to New and Improved Radar Indicators", PB 25481 and PB 32546, Office of the Publication Board, Department of Commerce, Washington 25, D. C. See also, H. W. Leverenz, "Luminescence and Tenebrescence as Applied in Radar", *RCA REVIEW*, Vol. VII, No. 2, pp. 199-239, June, 1946.

and the two lines on either side represent deviations of 10 per cent between measurements. Extensive measurements were made on many tubes in the laboratory in the course of screen development and the per cent of error encountered compares favorably with the values shown here. In general, the great majority of readings can be repeated with an accuracy of ±10 per cent but ±15 per cent has been assumed to include the extreme cases.

Correlation of Tube Rating With Tube Performance

For the purpose of correlating tube ratings with performance in the field, a group of tubes representing a range of characteristics were tested in an SCR-520-B radar set. The following general observations were very well established in the course of this investigation.

(1) The B_1 reading in this application gives little indication of tube performance and is of importance only as it is related to B_5.

(2) Tube performance varies directly with the B_5 rating;

(3) Build-up ratio is not of first order importance beyond a certain minimum. When screen layers become mixed, however, the build-up ratio drops rapidly and tube performance is greatly impaired.

(4) The flash ratio is not directly reflected in tube performance. However, high flash ratio is esthetically undesirable and indicates low conversion efficiency. This information is of importance in developing better phosphors. Abnormally high flash ratio can also be used to indicate poor screen fabrication since it will be often found to vary inversely with build-up ratio.

FACTORS AFFECTING P7-SCREEN CHARACTERISTICS

After an accurate laboratory method for rating screens that would reliably predict performance in the field was established, variables in the manufacture of cathode-ray tubes were investigated to discover their effect on P7-screen characteristics. Tests were selected which would include variables in phosphor manufacture, screen application, and tube processing and which would relate the variables to the final screen characteristics.

The Efficiency of Screen Components

The efficiency of the zinc-sulphide phosphor and zinc-cadmium-sulphide phosphor have ordinarily been expressed in the following terms:

Visual Efficiency—the fluorescent light output of the phosphor under

Table II—ZnS phosphor efficiency tests.

Test No.	Visual Efficiency %	Peak Efficiency %	Average Persistence Characteristics B5	BR
6	140	116	31.4	7.5
7	98	123	26.9	8.0
8	107	116	24.0	8.0
9	96	95	19.5	5.8
10	92	69	18.0	5.0

a standard electron bombardment measured with a photocell or phototube corrected to the eye sensitivity curve;

Peak Efficiency—the fluorescent energy output of the phosphor under a standard electron bombardment measured at the wavelength of maximum emission. For acceptance, the peak must fall within a specified wavelength range for each phosphor.

For the zinc-cadmium-sulphide phosphor a phosphorescence value was also recorded. It is a measure of the intensity of the persistent light of the phosphor one second after the steady-state election bombardment employed for the efficiency measurement has been stopped. All values are recorded as a percentage of a standard phosphor sample.

In the course of manufacture a relatively wide variety of efficiencies will normally be encountered. It was the purpose of this investigation to show the relation of phosphor efficiency to final screen performance. Phosphors representing a range of efficienices from the lowest to the highest available were selected and about fifteen tubes were made from each phosphor sample. The data presented in Table II and Table III are averages of each tube group. Variables other than phosphor efficiency were held to a minimum. In particular, screen settling was very closely supervised to assure uniform screen weight, and tubes were run consecutively through sealing and exhaust so that production

Table III—ZnCdS phosphor efficiency tests.

Test No.	Visual Efficiency %	Peak Efficiency %	Phosphorescence %	Average Persistence Characteristics B5	BR
1	83	75	110	17	7.5
2	95	97	92	20	5.4
3	92	95	79	15	5.8
4	102	106	67	16	6.5
5	106	118	67	18	6.3

variables would be minimized for each group in the test. The zinc-sulphide test samples were used with a standard zinc cadmium sulphide and vice versa. The relation of the efficiency of the blue zinc-sulphide (fluorescent) layer to screen characteristics and hence, final tube performance, is shown in Table II.

Approximate correlation between the B_5 level and the peak efficiency of the blue material is immediately evident. On going from 69 per cent efficiency to 123 per cent, the average B_5 was raised from 18 to 27 milli-footlamberts, an approximate 80 per cent increase in phosphor efficiency resulting in a 50 per cent increase in B_5. It is notable that this increase was achieved while maintaining the B_1 level thus resulting in a higher build-up ratio as well. Operational tests, it will be remembered, have shown this to be the most desirable type of gain. High visual efficiency made conspicuous contributions to the screen characteristics by maintaining B_5 in test #10 despite a very sharp drop in peak efficiency (from test #8) and greatly increasing B_5 in test #6 with peak efficiency unchanged. In each instance, however, the build-up ratio dropped. It was felt that a good correlation had been established between the efficiency of the zinc-sulphide phosphor and characteristics of the finished tube. The best material would have a high peak fluorescence with the ratio of visual to peak efficiency held to less than 1.5 if it were desired to preserve build-up ratio. Subsequently, use of higher-efficiency zinc sulphide in screen fabrication was accompanied by marked improvement in tube performance.

On the other hand, it may be seen from Table III that the efficiency measured for the P7 yellow material showed no consistent relation to ultimate screen characteristics. Average persistence characteristics showed some slight variation which might be attributable to error introduced by other variables that could not be controlled exactly. However, it could also be due to differences in the phosphor which were not being revealed by efficiency measurements under steady-state conditions. To pursue the latter assumption, it was necessary to investigate efficiency under 3650-Å radiation, the photoconductivity effect occasioned by ultraviolet radiation,[13] and the efficiency, build-up, and phosphorescence under blue-light excitation. Measurement of the magnitude of the build-up phenomenon produced by pulsed blue light (a procedure similar to the pulsed test procedure developed for the finished tube) proved to be the most reliable and sensitive method and gave results which correlated well with finished tube performance.

As has been previously explained, the yellow layer of the cascade

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[13] A. E. Hardy, "Photoconductivity of Zinc Cadmium Sulphide as Measured with the Cathoderay Oscillograph", *Trans. Electrochem. Soc.*, 87, 1945.

screen of a cathode-ray tube with a P7 screen is excited by the blue radiation from the zinc-sulphide phosphor. Therefore, in order to evaluate the yellow phosphor before using it in a tube, an excitation source was set up which had a spectral energy distribution essentially the same as that of the blue phosphor. This source was a Wratten #47 filter over a frosted 60-watt incandescent lamp operated at 115 volts a-c (Figure 5).

In order to avoid turning the lamp off and on for irradiating the phosphor, the lamp was placed in a box equipped with a camera shutter.

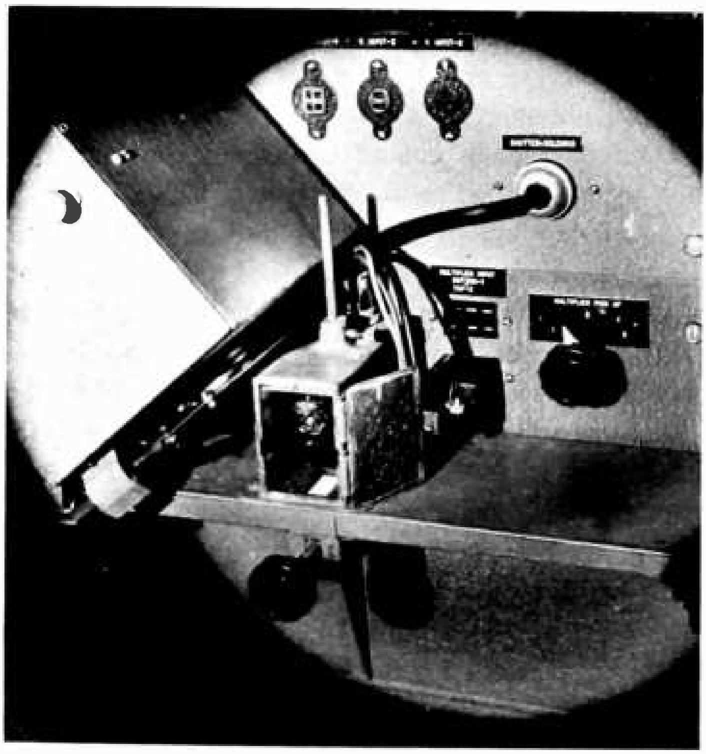

Fig. 5—Blue-light pulser unit showing relative position of light source, sample and multiplier phototube.

In addition, a solenoid was attached for opening and closing the shutter and for firing the shutter when it was set for instantaneous exposure (1/10, 1/25, 1/50 of a second).

The lamp radiation is directed through confining slits to the phosphor sample at an angle of about 40 degrees. Mounted directly over the sample (about 5 centimeters away) is a 1P22 multiplier phototube which measures the light emitted by the phosphors. Although the angle of incidence was chosen to minimize direct reflection of blue radiation, considerable scattering occurs at the surface of the sample which is absorbed by a Wratten #16 filter placed directly in front of the multiplier phototube. The signal developed by the multiplier phototube is amplified by a direct-coupled amplifier[14] and applied to the vertical deflection coils of a developmental 12-inch cathode-ray tube similar to the 12AP4 but with a P7 screen. However, a 12DP7 tube might also have been used.

A horizontal sweep which is variable from one second to one minute is provided by a motor-driven potentiometer (General Radio type 371-A). The Bakelite stops at either end of the resistance-wire winding were filed off and the gap between the ends of the winding filled in with a short piece of Bakelite. This alteration was necessary to permit the potentiometer arm to rotate continuously. While the arm is on the insulated section between the ends of the winding, no voltage is applied to the horizontal amplifier and the spot on the cathode-ray tube starts to drift back slowly. The use of a 0.1-microfarad capacitor between the arm and the starting end of the potentiometer minimized the back drift. In operation, the spot moves at a predetermined rate from left to right, remains at the extreme right while the arm is on the insulated sector, and then snaps back to the extreme left as the cycle begins again.

Coupled to the potentiometer is a microswitch arrangement which supplies a pulse at the beginning of each cycle. This pulse may be used for actuating the shutter on the source of the blue light. With the shutter set for time exposure, the shutter opens at the beginning of the cycle and the spot on the cathode-ray tube starts to move horizontally. As the luminescence grows, the spot is deflected in an upward course and the resultant graph is a plot of fluorescence build-up as a function of time as produced by the blue light. If desired, the shutter can be closed on the next cycle and a graph traced of the phosphorescence decay (Figure 6).

In addition to the initial pulse, a toothed wheel synchronized with the potentiometer produces pulses at one-second intervals. This arrangement allows the phosphor sample to be pulsed at one-second intervals with pulses of 1/10, 1/25, or 1/50 of a second duration. The use of a relatively low electric amplification in the phototube circuit makes it possible to observe the top of the pulses on the cathode-ray tube. By increasing the electric amplification, the pulse peaks may be sent off scale and the bottoms of the pulses may be observed (Figure 7). The increasing vertical displacement of the bottom of the pulses represents the phosphorescence build-up. It will be seen from the curves that each phenomenon represents a build-up to equilibrium. Furthermore, the absolute equilibrium level is a function not only of the phosphor efficiency but also of the intensity and duration of excita-

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[14] A. E. Hardy, "A Combination Phosphorometer and Spectroradiometer for Luminescent Materials", *Trans. Electrochem. Soc.*, 91, 1947.

tion. In addition it is dependent on the extent of the previous excitation and the elapsed time after excitation. Before making any quantitative measurements, samples were quenched by a deep red light (as in rating finished screens) to bring them to a so-called standard state.

A pulse width of 1/25 of a second was chosen for the standard test for two reasons. First, at 1/25 of a second the diaphragm needed to be only about half open to produce the same numerical build-up as would be recorded by the pulsed excitation test devised for the finished tube. This permitted wide adjustment on either side without altering the voltage on the lamp. Second, the shutter tripped more reliably at 1/25 of a second than it did at 1/50.

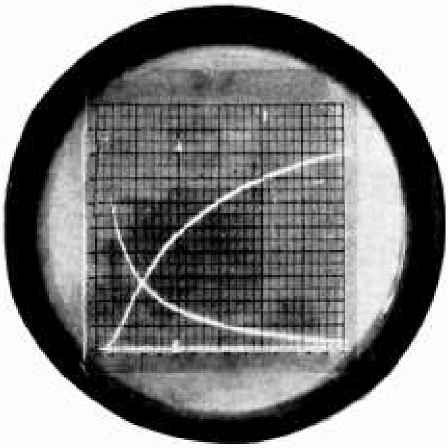

Fig. 6—Fluorescence and phosphorescence of P7 yellow, using 10-inch sweep and continuous excitation (blue light)

Fig. 7—Phosphorescence build-up of P7 yellow, using B_1 to B_7 pulsed excitation (blue light).

Because the phosphorescent brightness was measured on the finished tube at the end of the fifth pulse, the material specifications were set on the same basis. However, information gathered showed that the phosphorescent level after any pulse from 3 to 10 was just as significant in predicting tube characteristics. This relationship had been previously observed in the development of a suitable test for the finished tube.

Upon being tested with this new technique, the original yellow phosphor (as used for the efficiency test) had a B_5 of only 67 per cent of a reference sample. After considerable developmental work, the B_5 level was raised to an average of 150 per cent with one or two samples having gone as high as 200 per cent. It is significant that under direct cathode-ray bombardment the 67 per cent sample and the 150 per cent sample had the same efficiency. It is only in the finished tube or under pulsing with blue light that a difference is revealed.

Variables in Screen Fabrication

The field of screen application provided a most lucrative choice of variables for investigation, but the yield of factors to improve P7 screen characteristics over those currently obtainable was proportionately small. For instance, prolonged exposure of wet sulphides to ultraviolet radiation and the excessive grinding of phosphors have been known to reduce screen efficiency considerably. In the practical case, P7 screens were seldom allowed to settle for more than a few hours under the relatively weak (compared to sunlight) ultraviolet radiations of fluorescent lamps. Furthermore, with the relatively larger phosphor particle sizes used in P7 screens, periods of grinding were never very long. Therefore, tests of these variables through the ordinary range found in production resulted in changes of persistence characteristics of little more significance than the normal percentage of error encountered in the unit used for measuring persistence.

In the general discussion of the P7 screen, the importance of obtaining a good separation of layers was stressed and later the very poor performance of mixed screens was demonstrated in field tests. Many modifications of screen-settling techniques were investigated to improve this layer separation. In addition the layers were applied in separate operations as individual screens and various artificial layers transparent to light but opaque to electrons were attempted. Although research methods have demonstrated the effectiveness of excellent layer separation, no consistent significant improvement of P7 screens above the proved average obtainable with properly executed standard procedure resulted from these efforts.

The most important variable of this nature was associated with the physical disturbance of the layers as they were settling. This may seem elementary, in a sense, but under the wartime urgency for increased production, it came to be the single item most likely to get out of control. Emphasis on a suitable period for the screen to settle following application of each layer resulted, in one month, in the improvement per tube type tabulated in Table IV.

"Allowing them to settle" had to be further translated to mean:

(1) no physical contact with settling tables during the entire period of settling;

(2) no inspection of the primary layer for blemishes during the settling process because it involved lifting the bulbs off the table.

The shape of the bulb was an important factor in this respect and the larger diameter, flat-face tubes that were more stable consistently

Table IV—Improvement in build-up characteristic obtained by more care in screen application.

Tube Type	% Improvement in one month
3FP7	27
3HP7	27
5CP7	63
5FP7	43
7BP7	15
12DP7	43

exhibited better screen characteristics. Tubes such as the 5CP7, on the other hand, with curved faces and high ratios of length to diameter were most critical.

Phosphor Weight Per Layer

Another important variable in screen application and one over which positive control can be exercised is the thickness of the screen components. To investigate this parameter, a series of tests in finished tubes were made and their characteristics later evaluated by the persistence-measuring equipment. In each instance, a given weight was chosen for the first layer and the second layer was varied over the practical range. In this manner, since the selection of weights for the first layer covered the practical range, all possible combinations were assembled. The thinnest screens possible are limited by graininess due to insufficient coverage of either the glass or the yellow layer while the limit in the other direction is set by screen adherence. Persistence characteristics for the weights commonly employed in each layer as the other layer component is varied over the permissible range are shown by the curves in Figures 8 and 9. General conclusions to be drawn from the curves are:

(1) For any given thickness of the yellow layer, increasing the blue layer beyond the minimum thickness required for coverage will

(a) reduce B_1;

(b) reduce B_5;

(c) increase build-up ratio;

(d) reduce flash ratio.

The rate of change and total change, however, is also a function of the primary-layer thickness as is shown by the second group of curves.

(2) When the thickness of the blue layer is on the order of 20 milligrams/square centimeter, varying the thickness of the yellow layer will show an optimum point for all characteristics. When the blue layer is made thinner, i.e., in the ordinary range as in Figure 9, increasing the thickness of the primary layer will

(a) reduce B_1;
(b) show an optimum for B_5;
(c) increase build-up ratio;
(d) reduce flash ratio.

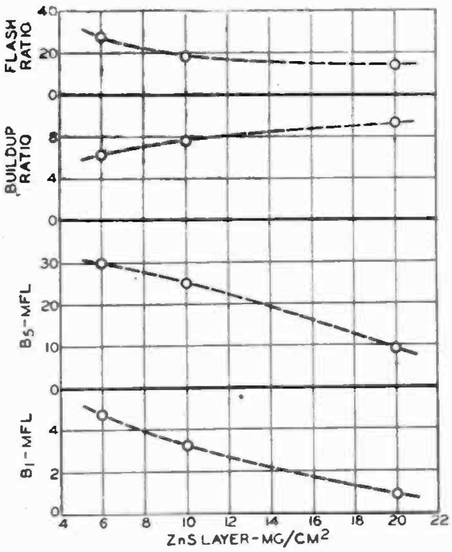

Fig. 8—P7 screen-thickness tests with ZnCdS layer of 12 milligrams/square centimeter.

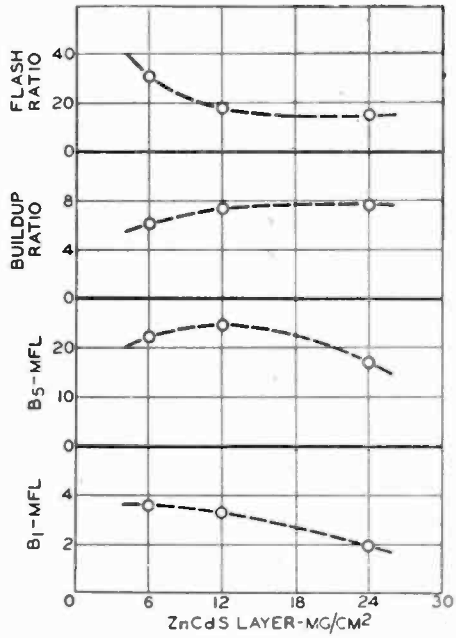

Fig. 9—P7 screen-thickness tests with ZnS layer of 10 milligrams/square centimeter.

Quantitative variations may be taken directly from each curve. The curves have proved very useful in predicting correct screen weights for experimental types where specific characteristics (high B_1, low flash, etc.) were desired.

Processing Variables

As might be expected, there are few variables in the normal process of exhaust that will affect screen characteristics. Contamination of screen by water vapor, metallic evaporation, etc., do occur but are to be eliminated rather than evaluated. The baking temperature on exhaust for outgassing bulbs, however, can be measured and controlled. The permissible range of temperature is generally quite limited between adequate degassing and collapse of the bulb due to glass softening. Nevertheless, this variable was investigated in the permissible range with the results shown in Table V. Data presented are again the averages of a number of tubes containing screens of the same composition made from the same lots of phosphors and run through exhaust consecutively. The data indicate that excessive temperatures are to be avoided during exhaust. Frederick Seitz[4] advances a theory to explain the phenomena which proposes that evaporation of sulphur takes place at this temperature in vacuum. When this happens the delicate balance of elements naturally acquired in the synthesis of zinc sulphide is disturbed and leaves free zinc atoms in the lattice structure of the phosphor crystal. The luminescent efficiency of the phosphor would, therefore, be more or less impaired depending on the time and temperature of heating.

Conclusion

The pulse method devised for determining persistence characteristics served a very useful purpose in the development of P7 screens. Ratings on screens established in the laboratory were used to give reliable predictions of performance in the field. Subsequently, it was also used to great advantage in investigating manufacturing variables for their effect on persistence characteristics. Some of the variables have been found to have considerable effect on the characteristics while others were not very critical. Both types of information proved to be of value in the evolution of better long-persistence screens.

During the course of the investigation, the B_5 characteristic and the buildup characteristic have been established as the dominant factors affecting screen performance. Other characteristics were important only insofar as they affected these characteristics and in general are controlled by them. It was proposed, therefore, to simplify the persistence-measuring procedure in the following manner. Excite the screen with a fixed number of standard pulses and set a minimum

Table V—Effect of exhaust-bake temperature on persistence characteristics.

Baking Temperature °C	Average Persistence Characteristics B5	BR	FR
425	25.8	7.9	163
400	29.0	7.7	128
375	31.6	7.1	108

value for the attained brightness of the screen at this point. The rating would then be a complex function of B_5 and build-up which would accurately predict tube performance in a single figure of merit. Although some encouraging data were collected on this simplified method, insufficient information was available at the time P7 screen development on radar indicators became inactive to correlate the new method with performance in the field. Both the old and new methods have subsequently been used to good advantage in arriving at the best combination P7 screen for oscilloscope use and for early development work on screens for Teleran. The results indicate the broad, general usefulness of the method in fields other than that for which they were specifically developed.

ACKNOWLEDGMENTS

In the course of development, the efforts of many persons contributed directly to the success of the project. Dr. R. F. Bacher (later —Dr. L. J. Haworth) and his staff at the Massachusetts Institute of Technology made notable contributions to the work. In particular, of this group, Dr. T. Soller and Mr. A. Y. Bentley gave much assistance in field requirements and tests while Dr. W. B. Nottingham contributed a large share of the original methods incorporated in the test procedure. Mr. B. Kazan and Mr. M. E. Crost of the Camp Evans Signal Laboratory were very cooperative in supplying test results from the Belmar equipment to promote correlation of final tube rating with measurements made on the service unit.

H. W. Leverenz, R. E. Shrader and E. J. Wood (deceased) of the RCA Laboratories Division were responsible for the phosphor research that preceded manufacture of the tubes. W. H. Painter and M. Sadowsky contributed the water-settling technique for screen application. Much of the final design of the radar 'scope was based on the development work of L. E. Swedlund and J. T. Wilson, in the Tube Department. Revisions in the phosphor manufacturing process to improve the B_5 screen characteristics reflect the work of J. H. Markowski and B. E. Artau. Construction of the test gear was done by S. T. Hill (deceased).

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RADIOPHOTO STANDARDS*

By

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Summary—Radiophoto, an outgrowth of Alexander Bain's facsimile development in 1842, has progressed through the years in a rather uncoordinated manner. An attempt to correlate and standardize radiophoto equipment and practices at the 1938 International Conference in Stockholm was postponed by war conditions in Europe. Recently, the Radio Technical Planning Board has submitted some very encouraging proposals on standards. The two International Coordinating Committees on Radio and Telegraph have questions on their new agenda which can lead to one standard applicable to wire line and radio services and enable the linking of the wire and radio picture networks of the world. A brief history of radiophoto development introduces the paper which proceeds with a discussion of the various types of equipment in current use. Objectives in further development are listed and suggested answers to the questions referred to above are given.

Introduction

THE average person not active in the communication, press, or publication fields generally considers, when he sees a Radiophoto in a newspaper or magazine, that the art of picture transmission has suddenly been born. Actually, the transmission of pictures electrically had its inception about the same time as straight telegraphy, when in 1842 Alexander Bain, an English physicist, first proposed a device to send pictures from one place to another by electric wires. His plan was so basically sound that the systems used today follow his original scheme, at least, in principle. In his apparatus there were two pendulums (Figure 1) arranged electrically in such a manner that they would swing back and forth at a constant rate, much as the pendulum on a clock. The transmitting pendulum carried a flexible wire that would scan or scrape over the top of a block of type faces making electrical contact with the type as it passed over. The resulting electric pulses were transmitted by wire to the receiving pendulum. The receiving pendulum swung back and forth across a piece of chemically treated paper and discolored the paper at each point where the electric current passed between the receiving pendulum contact wire and the paper, thereby recording what was scanned by the transmitting pendulum. A ratchet mechanism was so adjusted that each time the pen-

* Decimal Classification: R581 × R020.

dulums swung across the type face and the paper, both the type face and the paper would advance one notch, so that on the next stroke, the pendulums would scan and record the next line.

The systems in use today are considerably more practical and incorporate many new technical achievements. Material for scanning is not limited to a set of type, but, instead any picture can be wrapped around a revolving cylinder and scanned line by line with a small light spot and photo cell. At the receiving end, various methods of recording are employed which not only use the discolored paper scheme but also use photographic, electrochemical, electrolytic, electrothermal, electromechanical and carbon recording methods.

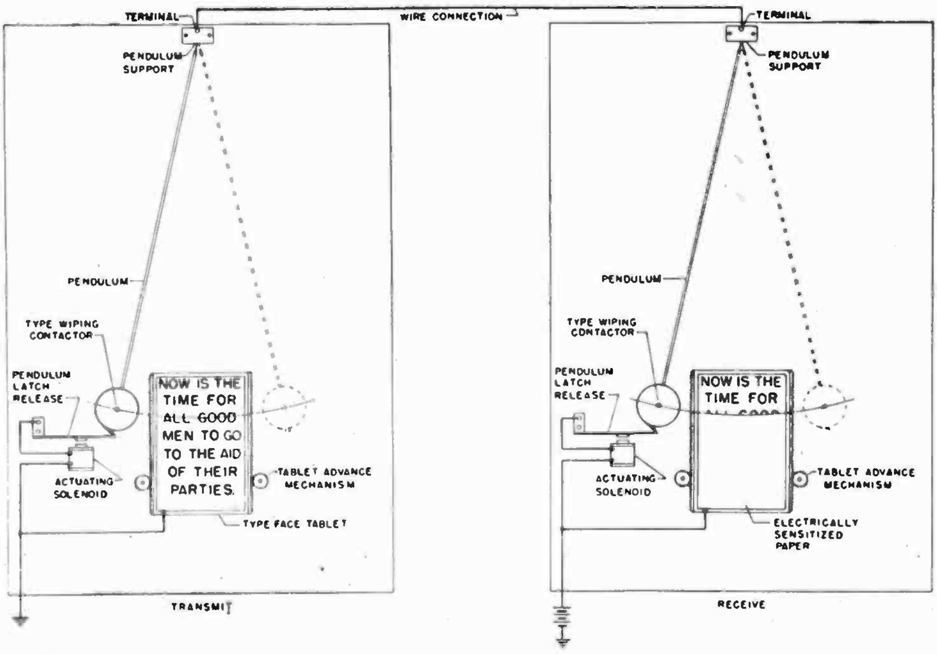

Fig. 1—Author's conception of Alexander Bain's first Facsimile System in 1842.

In the early part of this century, the development of phototelegraphic equipment in the United States and in Europe was concentrated on the transmission of pictures over wire line circuits. However, in 1924, RCA transmitted the first radiophoto, a picture of Charles Evans Hughes, from New York to London, where it was automatically retransmitted back to New York and recorded (Figure 2).[1]

From that time until 1938, practically all discussions of phototelegraphy requirements, as far as radio circuits were concerned, were relative to time modulation methods, i.e., a system wherein a picture was transmitted and received in a series of telegraphic dots that varied

[1] R. H. Ranger, "Transmission and Reception of Photoradiograms", *Proc. I.R.E.*, Vol. 14, No. 4, pp. 161-180, April, 1926.

in length depending upon the density of the picture. The Hughes picture (Figure 2) is composed of random dots, whereas in the later development of the constant frequency variable dot system[2], pictures were composed of uniformly spaced dots (Figure 3) that were equivalent to the screen produced in the average newspaper photograph.

At the time, this was a great accomplishment, but the system left much to be desired. Pictures could not always be printed in the exact screen or scale of the original transmission. Also, fading could eliminate a few dots or static might add a few additional dots in a strategic portion of a picture.

During the recent war, frequency modulation methods[3] of transmitting pictures on a radio circuit completely supplanted time modula-

Fig. 2—The first radiophoto transmitted from New York to London, where it was automatically retransmitted back to New York and recorded (July 6, 1924).

tion and apparatus was brought into use which operated with characteristics different from those recommended by the two International Coordinating Bodies, the CCIT* and CCIR.**

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[2] J. L. Callahan, J. N. Whitaker, and H. Shore, "Photoradio Apparatus and Operating Technique Improvements", *Proc. I.R.E.*, Vol. 23, No. 12, pp. 1441-1482, December, 1935.

[3] R. E. Mathis, and J. N. Whitaker, "Radio Facsimile by Sub-Carrier Frequency Modulation", *RCA REVIEW*, Vol. IV, No. 2, pp. 131-153, October, 1939.

* CCIT—Consultative Committee on International Telegraph, the coordinate body for standards, operating practices, tariffs and international arrangements for land line and cable systems.

** CCIR—Consultative Committee on International Radio, the coordinate body for standards, operating practices, tariffs and international arrangements for radio communication systems.

Communications has brought the various parts of the world much closer together and, at present, there is a serious need for standard methods which will permit faster and more economical transmission of pictures over wire and over radio circuits, or a combination of the two, so that the immense wire networks of Europe and the United States can be tied together on a worldwide basis.

For clarification in this article, the term "radiophoto" is used to denote transmissions by radio and "telephoto" is used to denote transmissions over wire lines. The word "facsimile", although a general term used to cover the entire field, has recently been more closely

Fig. 3—Constant Frequency Variable Dot (CFVD) from London to New York.

related to the transmission over wide-band channels at higher speeds using apparatus for continuous recording.

INDEX OF COOPERATION

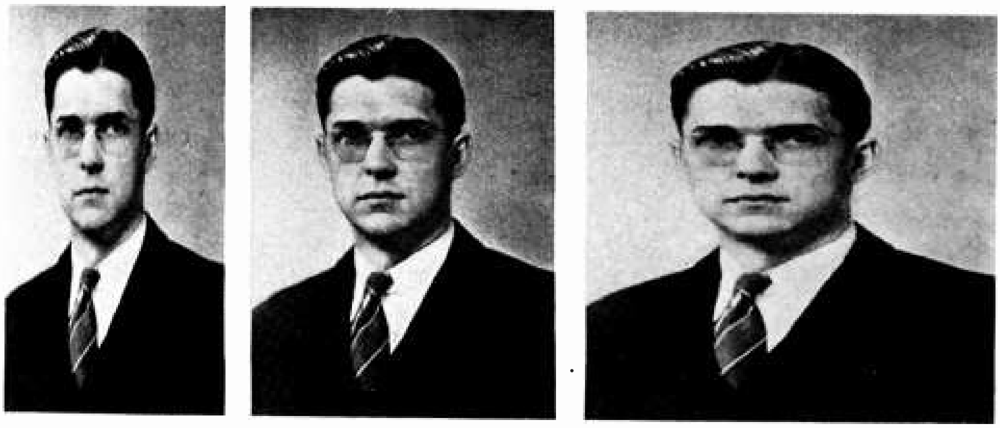
Before discussing standardization problems, it is necessary to have a clear understanding of what is meant by the *index of cooperation.* This is a factor—literally an index of rectangular proportions—that can be calculated for any machine—regardless of its cylinder size—

which indicates whether or not one machine can work or cooperate with another machine having different specifications.

The international index of cooperation, as defined by the CCIR, is the product of the line advance (in lines per millimeter) and the cylinder diameter (in millimeters). It should not be confused with the Basic Index which is the product of the line advance and the length of scanning line. This is applicable but is more generally used in continuous recording facsimile systems.

If the index of cooperation of two machines is the same (within 5 per cent) pictures can be exchanged between them, as far as proportion is concerned, but there may be a reduction or enlargement in actual size.

Figures 4(a) and 4(c) are examples of transmission between two machines with indices of cooperation of 350 and 264. In Figure 4(b)



(a) (b) (c)

Fig. 4—Photographs transmitted and received on cylinders of 70-millimeter diameter but different Indices of Cooperation due to different line advances. Figure 4-b is a normal transmission at 3.77 lines/millimeter; Figure 4-a was sent at 3.77 lines/millimeter and recorded at 5 lines/millimeter; Figure 4-c was sent at 5 lines/millimeter and recorded at 3.77 lines/millimeter.

both indices are 264 and the received picture is rectangularly proportional.

Types of Equipment

At present there are some fourteen or more types of radiophoto and telephoto equipment in general use throughout the world. Fortunately, only four types predominate and need be considered as far as international radiophoto communication is concerned. The four important types of equipment considered here are:

1. CCIT STANDARD

Cylinder diameter	66 millimeters	2.598 inches
Cylinder length	132 millimeters	5.196 inches
Line advance	4 and 5.33 lines/millimeter	101.6 and 135.4 lines/inch
Maximum picture size	13 x 18 centimeters	5.118 x 7.086 inches
Cylinder rotation	60 revolutions per minute	
Index of cooperation	264 and 352	

This machine was designed for amplitude modulation over wire lines and has been used on the European wire line systems for a number of years. There are probably 100 to 150 machines of this type still in existence on the continent of Europe; however, this figure is approximate since no accurate check has been possible since the war. The number of machines still in active use must be considered in any standardization problems because it is impossible to change types and methods of operation overnight. The size of copy on this machine is limited to 5 x 7 inches, which is fairly convenient for many press photographs, but unsatisfactory for numerous commercial applications. Fortunately, frequency modulators and demodulators have been produced in recent years which adapt this equipment for use on radio circuits.

2. CCIR STANDARD

Cylinder diameter	88 millimeters	3.464 inches
Cylinder length	300 millimeters	11.81 inches
Line advance	4 and 5 lines/millimeter	101.6 and 127 lines/inch
Maximum picture size	26 x 29 centimeters	10.236 x 11.417 inches
Cylinder rotation	60 revolutions per minute	
Index of cooperation	352 and 440	
Frequency shift limits	1600 cycles per second (black), 2000 cycles per second (white)	

This large diameter machine was suggested by several of the international radio communications organizations for adoption at the 1938 International Conference which was postponed due to war conditions in Europe. A number of these machines were built and forty or fifty are still in active operation by radio carriers throughout the world. The machine was designed when time modulation methods were the best-known means of transmission on radio circuits and the large cylinder diameter permitted the transmission of a large photograph which could be reduced at the receiving end and thereby eliminate, to a certain extent, the dot pattern described previously.

This machine, at an index of cooperation of 352, will cooperate with the 66 millimeter machine, as far as rectangular proportions are concerned, but will result in a ¼ reduction, or a ⅓ enlargement, depending upon the direction of transmission. In the direction of enlargement, there will also be an apparent loss of quality due to the change in line advance in transmitting from 5.33 to 4 lines per millimeter. A further problem with this large diameter machine is that its keying speed is comparatively high for the same speed of transmission.

The frequency shift range used on the CCIR equipment is 1600 cycles per second for black, 2000 cycles per second for white. This offers such a narrow frequency difference between black and white that setup from an original picture must be done very carefully and makes the possibility of automatic transmission questionable. Any error in setup, or an equipment drift of 40 cycles, gives a 10 per cent change in picture density. Since noise increases proportionally with deviation from the carrier, if the black were represented by the higher frequency, spurious noise would be less apparent in the final recording.

3. TIMES FACSIMILE CORPORATION

Cylinder diameter	70 millimeters	2.756 inches
Cylinder length	186 millimeters	7.32 inches
Line advance	3.77 lines/millimeter	96 lines/inch
Maximum picture size	17.8 x 20.3 centimeters	7 x 8 inches
Cylinder rotation	90 revolutions per minute	
Index of cooperation	264	
Frequency shift limits, old	1800 cycles per second (white), 3000 cycles per second (black)	
new	1500 cycles per second (white), 2300 cycles per second (black)	
Fork Frequency	1800 cycles per second	

4. ACME NEWSPICTURES, INC.

Cylinder diameter	73.78 millimeters	2.904 inches
Cylinder length	188 millimeters	7.401 inches
Line advance	3.93 lines/millimeter	100 lines/inch
Maximum picture size	17.8 x 21.5 centimeters	7 x 8.46 inches
Cylinder rotation	100 revolutions per minute	
Index of cooperation	290	
Frequency shift limits	1000 cycles per second (white), 2500 cycles per second (black)	
	1500 cycles per second (white), 2500 cycles per second (black)	
Fork Frequency	1920 cycles per second	

Interchangeable Light Tight film cartridge for daylight recording.

Comments from Operating Organizations

Much correspondence has been exchanged with the foreign operators concerning experience in the radiophoto field. This correspondence has produced the following comments on problems and difficulties with present equipment:

1. One of the principal limitations which must be considered in

discussing combined radiophoto-telephoto standards is that of the low keying speeds permissible on radio circuits. Keying speed is a function of cylinder diameter, revolutions per minute, and line advance. In order to obtain greater detail through finer line advance, and increased speed of transmission through higher cylinder revolutions per minute, it is necessary to keep cylinder diameter reduced to the smallest practicable size.

2. The CCIT 66-millimeter machine, although well established, is believed to be too small for general commercial use. Doubling the cylinder length would greatly increase its usefulness without increasing keying speed.
3. The CCIR 88-millimeter machine is very useful for large copy, but the high keying speed, due to the large diameter cylinder, is a disadvantage and generally will not permit an increase in speed above 60 revolutions per minute. A 20 per cent smaller cylinder would accommodate the majority of the subjects filed for transmission and permit lower keying speeds. The frequency shift limits (1600 cycles, black—2000 cycles, white) are too narrow and require very accurate settings.
4. The Times cylinder diameter is practical, but cylinder length should be increased.
5. The Acme cylinder diameter is practical but its length should be increased to 30 centimeters.
6. Separate drives for the cylinder and leadscrew would be advantageous.
7. A light-tight cartridge loading system for daylight recording would be highly desirable.
8. A practical size of cylinder would be one which will accommodate the two standard sizes of correspondence sheets: European, 21 x 29.9 centimeters (8.25 x 11.7 inches), American, 21.5 x 28 centimeters (8.5 x 11 inches).
9. A single set of frequency shift limits, for use with all types of equipment on combined wire line and radio circuits, should be provided. This will simplify central office and circuit operation, and facilitate interconnection of radiophoto and telephoto services.
10. Skew and synchronizing problems will be practically eliminated by providing standard frequencies of greater stability and accuracy than those now specified by the CCIR and CCIT or by providing an automatic synchronizing system.
11. A storage system is required at terminal points (like tape storage for telegraph) to facilitate economical switching be-

tween various trunk circuits. Such a system will permit the storage of relatively high-speed telephoto signals for transmission on radiophoto circuits at slower speeds.

OBJECTIVES IN FURTHER RADIOPHOTO DEVELOPMENT

In order to make phototelegraphic transmission practical, economical, and available to the public in general, there must be international and universal standards in equipment covering both radio and wire line use, and an agreement on standards and methods of operation. The ideal situation, of course, is one universal machine for radiophoto, telephoto (and, if possible, facsimile) use on all types of communication circuits. The electrical properties of such an equipment should cover all the requirements of the international coordinating bodies—the CCIT, CCIR and CCIF.*

The design of, and regulations for, any new equipment should be farseeing and flexible in order to allow for or permit future expansion or improvement.

Effort must be directed toward a system which will make possible the automatic and semi-automatic relay of phototelegraphic material by radio and wire line links in the same way that telephone and telegraph messages are now handled.

Such an ideal cannot be accomplished immediately, but an agreement by all concerned, particularly the International Committees, will permit continual development and construction toward this goal.

A major portion of the phototelegraphic apparatus now in use will have to be replaced in the next several years, and it is proposed that the International Committees guide this new construction toward a universal standard for worldwide use and development.

COORDINATION WITH THE CCIT

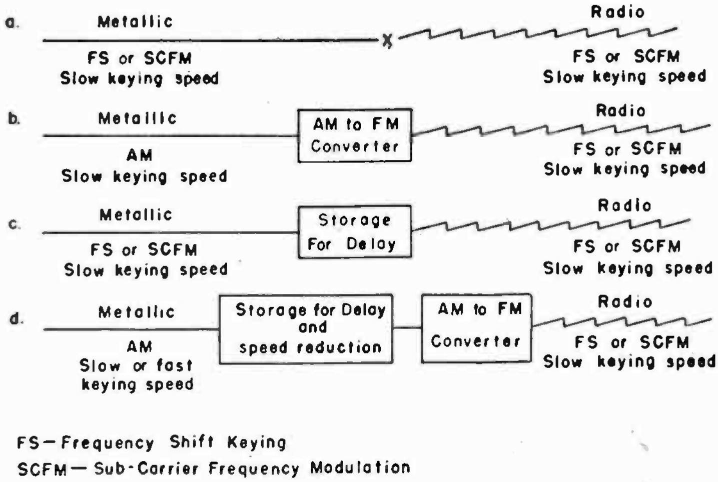
Although the coordination and eventual standardization of radiophoto will be handled by the CCIR, activities must also be coordinated with the CCIT which is concerned with standardization of all wire line activities and equipment. The CCIT has on its agenda for the 1948 meeting two questions regarding phototelegraphic standards. These questions are listed below followed by suggested answers to the problems.

Question V, 8, a: "Which are the most recommendable measures for the transmission of halftones over cable circuits,

* CCIF—Consultative Committee on International Telephone, the coordinating body for standards, operating practices, tariffs and international arrangements for telephone systems.

and over mixed metallic and radio circuits, with the view of obtaining the best results?

1. In considering operation over mixed metallic and radio circuits the following combinations are possible:



2. One of the problems of primary importance is the agreement upon both machine and electronic equipment standards which will permit the automatic or semiautomatic relay of phototelegraphic material between radio circuits and metallic circuits without recording and rescanning a picture at terminal points. In the present state of the art, the radio channel is a limiting factor in band width and keying speed.

3. At present, amplitude modulation is predominantly used for the transmission of telephoto over wire lines, cables, and carrier channels because phase delay or distortion is less troublesome with this type of modulation. Most of the wire line and cable channels now in use are not sufficiently corrected for uniform phase delay at all frequencies in the transmission bandwidth to permit good frequency modulated picture transmission.

4. Frequency modulation of an audio carrier (SCFM) is predominantly used for radiophoto transmissions. Some carrier shift, or frequency modulation of the radio-frequency carrier, is used for the transmission of pictures via radio. However, there are very few transmitters and receivers whose stability is good enough to permit picture transmission by this means. This same problem is apparent on carrier circuits.

5. Fortunately, the practical carrier frequencies, bandwidths and keying speeds for radio circuits fall almost exactly within the telephone limitations, specified by the CCIF, of 300 to 3400 cycles. Although there still exist a large number of circuits whose bandwidth

is limited to about 2600 cycles, satisfactory halftones can be transmitted over these circuits. Since it is the general intention to install all new telephone transmission systems with higher cut-off frequencies of 3400 cycles, it would be a step backwards to limit picture transmission quality and speed because of the old standard.

6. The fact that many of the existing wire and cable circuits are not capable of satisfactorily handling frequency modulation at the present time, should not affect the general standardization of frequency modulation parameters because modern modulators and demodulators for converting AM to FM and FM to AM are both simple and relatively

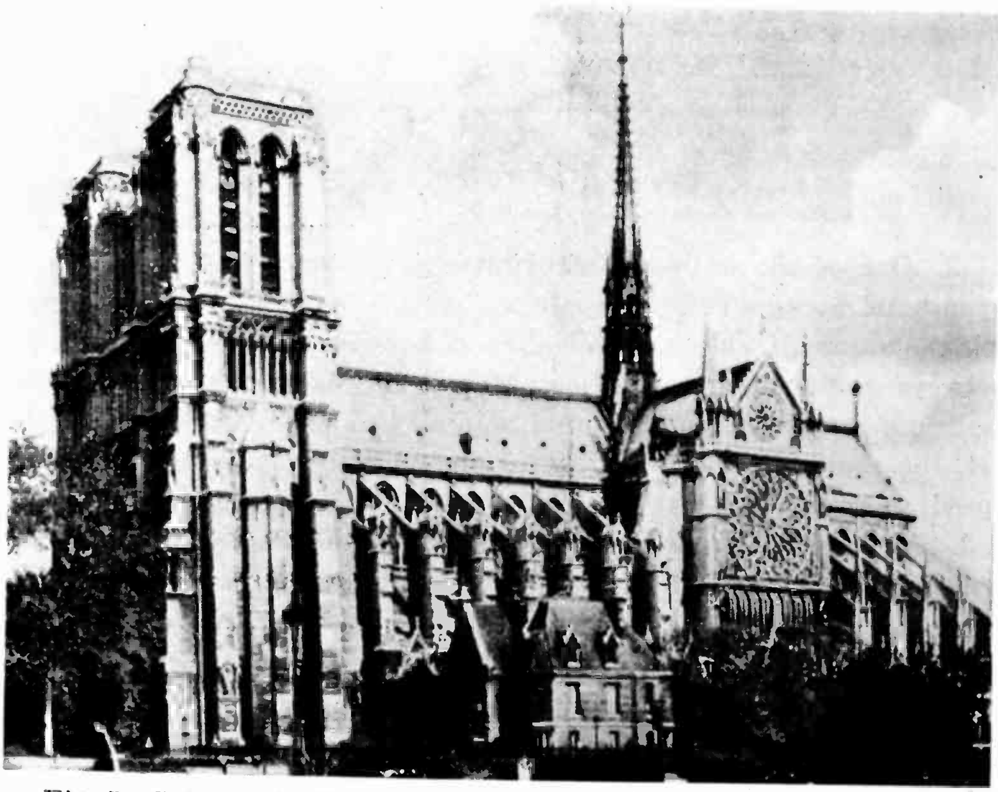


Fig. 5—Sub-Carrier Frequency Modulation from Paris to New York.

inexpensive. It would be helpful if all these converters could be of a linear type so that a straight line conversion characteristic is apparent throughout a series of systems connected together.

7. Sub-carrier frequency modulation, and eventually radio frequency carrier shift (RFCS), will be the most satisfactory means for transmitting radiophoto material, since selective fading, signal-to-noise ratio, harmonics and general amplitude variation are either less detrimental or easier to control.

8. Recent theoretical studies and multipath measurements indicate that except in extremely rare instances, phototelegraphic material,

because of the relatively high keying speeds associated with its transmission, is subject to a certain amount of distortion in the radio circuit. This distortion varies from a minimum, when the use of optimum frequencies and antennas with well proportioned and directed vertical beams reduces the delay to that inherent in single ray propagation, up to a maximum, when operating conditions dictate the use of other than optimum frequencies and antennas with wide vertical patterns which admit propagation of a multiplicity of rays. For example, on a radio circuit, equivalent to the New York-London, or the New York-Tangier, using average frequencies and ordinary antennas, the average maximum keying speeds for certain percentages of distortion are as follows:

	10 Per Cent Distortion	20 Per Cent Distortion	40 Per Cent Distortion
Summer......	62.5 cycles	125 cycles	250 cycles
Winter.......	185 cycles	370 cycles	740 cycles

With optimum frequencies and antennas capable of rejecting all but single ray propagation, keying speeds increase to the following maximums:

	10 Per Cent Distortion	20 Per Cent Distortion	40 Per Cent Distortion
Summer........	240 cycles	480 cycles	960 cycles
Winter.........	340 cycles	680 cycles	1360 cycles

9. The use of higher operating frequencies and antennas with better pattern control permitting sharper vertical angle selectivity will alleviate this situation to a great extent but at this time, and probably for some time to come, it is necessary to make the best use of available equipment. An automatic relay station at the mid-point of the above circuits would reduce, by almost a half, the multipath distortion and thereby almost double the allowable keying speeds.

10. By reference to the Appendix, which shows keying speeds for various machines in use, it can be seen that some distortion must be accepted with even the smallest diameter cylinder unless cylinder diameters and rotational speeds are reduced to an uneconomical and impractical value.

11. These limitations are, in reality, not as drastic as they appear. The keying speeds shown in the Appendix are calculated on the basis of equal horizontal and vertical definition with the horizontal definition taken as the width of the scanning line. This fineness of detail is not encountered in all pictures transmitted and when it is, it may cover such a small percentage of the subject that the picture as a whole is

commercially acceptable. The limitation does indicate, however, that in order to obtain the maximum detail at economic operating speeds, the smallest diameter cylinder practicable for commercial operation must be adopted.

12. No practical method has been devised for measuring the effects of distortion on halftones, but it is known from experience that subjects of large detail and good contrast (such as a portrait) can stand some 30 per cent more multipath distortion than a subject of fine detail and low contrast (such as a picture of building wreckage) and still be of commercial quality.

13. Since there is presently no practical means of continuously varying transmission speeds over radio circuits to conform to circuit conditions and type of copy, a compromise must be accepted on the best practical and average size, speed, and quality of equipment, improving each as technical progress permits.

> *Question V, 8, b:* "Which are the circuits that may be used for high speed phototelegraphic transmission, both from a technical as well as operational viewpoint?"

1. For higher transmission speeds than permitted by the present radiophoto-telephoto standards, it is suggested that those circuits proposed by the CCIF for the international transmission of music, be adopted.

2. For the interconnection of slow radiophoto circuits and fast telephoto circuits, the methods shown in item I under Question V, 8, a can be utilized.

Recommendations

During World War II the Radio Technical Planning Board (RTPB) was organized to assist the armed forces and production interests in standardizing equipment and component parts within the electronics industry. Recently Section 4 of Panel 7 of the RTPB reviewed the situation in the phototelegraphic field and made the following recommendations:

A practical radiophoto-telephoto machine would have the following specifications:

Cylinder diameter	70.00 millimeters	2.75 inches
Cylinder circumference	219.91 millimeters	8.66 inches
Gripping loss	13.00 millimeters	0.51 inch
Maximum skew	3.71 millimeters	0.15 inch
Minimum useful circumference	203.2 millimeters	8.00 inches

Useful cylinder length	300.00 millimeters	11.81 inches
Maximum size picture	205 x 296 millimeters	8.07 x 11.65 inches
Line advance	3.77 and 5 lines/millimeter	96 and 127 lines/inch
Index of cooperation	264 and 350	
Cylinder speed	*90 and 60 revolutions per minute	
Audio frequency shift limits	White 1500, black 2300 (cycles)	

Standard Frequency: A multiple or submultiple of 300 cycles
Accuracy of calibration one part per million

Stability:	Short time variation (30 minutes)	± 1 per 200,000
	Long time drift (6 months)	± 1 per 200,000
	Adjustable	± 1 per 50,000

Conclusions

It is believed that the above RTPB recommended standards are the best compromise possible under present conditions. The 70-millimeter diameter cylinder with 3.77 lines per millimeter line advance will work with Times equipment at 90 revolutions per minute and, with 5 lines per millimeter, it will cooperate with the CCIT and CCIR standards at 60 revolutions per minute (with an area change in transmission). It will not cooperate with the Acme machine, as it is, but the Acme machine is so designed mechanically that it can be easily made to cooperate merely by changing the cylinder diameter and/or changing the speed of the lead screw.

The standard frequency, a function of both 50 and 60 cycles, received thought beyond the radiophoto field. The ideal, of course, would be a primary standard, correct to one part in 20,000,000, in every communication center for use, not only on radiophoto, but also for multiplex control, circuit channeling, time standards, television use and laboratory measurements. Unfortunately, the present economic and equipment situations throughout the communications world will not permit such a radical change at this time.

The fact that the short time and long time variations are both plus and minus one part in 200,000 may be disconcerting. In compromising on available and desirable stability it was finally decided that the speed of a machine could be allowed to vary within that amount in 30 minutes, however, over a long period the error could not become accumulative. It must either reverse drift or be corrected.

Little has been said about the new frequency shift limits (1500 cycles-white, 2300 cycles-black) since so many international carriers have adopted or are adopting them that the problem may be considered to have been solved. Today, London, Paris, Rome, Cairo, Bombay and

* Double or triple these speeds for telephoto.

RCA Communications, Inc., New York are using the new limits. Rome, Bern and Lisbon are installing new linear modulators and demodulators on the new limits. This 800-cycle shift fits in with frequency shift practice in telegraph service looking forward to the time when receivers and transmitters will permit radiophotos to be intermingled with telegraph traffic.

Because of the recent war these recommendations represent, as far as is known, the first practical step toward a world phototelegraphic standard since 1938. Several undesirable factors still remain, but, if no limiting factors are injected, an economical worldwide picture handling system is in prospect. Technical advances in the last few years point the way to answering practically every problem.

APPENDIX

Maximum Keying Speeds
(Square dot assumed)

Cylinder diameter 66 millimeters, Circumference 207.345 millimeters (8.16 inches)

	3.77 lines/mm.	4 lines/mm.	5 lines/mm.	6 lines/mm.
40 RPM		276 cycles	345 cycles	415 cycles
60 RPM		415 cycles	518 cycles	622 cycles
90 RPM		622 cycles	777 cycles	933 cycles
100 RPM		691 cycles	864 cycles	1036 cycles
120 RPM		830 cycles	1036 cycles	1244 cycles

Cylinder diameter 70 millimeters, Circumference 219.912 millimeters (8.65 inches)

	3.77 lines/mm.	4 lines/mm.	5 lines/mm.	6 lines/mm.
40 RPM	276 cycles	293 cycles	366 cycles	440 cycles
60 RPM	414 cycles	440 cycles	550 cycles	660 cycles
90 RPM	621 cycles	660 cycles	825 cycles	989 cycles
100 RPM	690 cycles	733 cycles	916 cycles	1100 cycles
120 RPM	828 cycles	880 cycles	1100 cycles	1320 cycles

Cylinder diameter 74 millimeters, Circumference 232.47 millimeters (9.15 inches)

	3.77 lines/mm.	4 lines/mm.	5 lines/mm.	6 lines/mm.
40 RPM		309 cycles	387 cycles	464 cycles
60 RPM		465 cycles	581 cycles	697 cycles
90 RPM		697 cycles	871 cycles	1046 cycles
100 RPM		774 cycles	968 cycles	1162 cycles
120 RPM		930 cycles	1162 cycles	1394 cycles

Cylinder diameter 88 millimeters, Circumference 276.46 millimeters (10.88 inches)

	3.77 lines/mm.	4 lines/mm.	5 lines/mm.	6 lines/mm.
40 RPM		369 cycles	460 cycles	552 cycles
60 RPM		553 cycles	691 cycles	829 cycles
90 RPM		829 cycles	1036 cycles	1244 cycles
100 RPM		921 cycles	1151 cycles	1382 cycles
120 RPM		1106 cycles	1382 cycles	1658 cycles

Formula: $$\frac{\text{Circumference in millimeter} \times \text{Lines/mm.} \times \text{RPM}}{60 \times 2}$$

= Maximum fundamental keying speed in cycles per second

3.77 Lines/mm. = 96 lines per inch
4 Lines/mm. = 101.6 lines per inch
5 Lines/mm. = 127 lines per inch
$5\frac{1}{3}$ Lines/mm. = 135.4 lines per inch
6 Lines/mm. = 152.4 lines per inch

UNDERWATER SOUND TRANSDUCERS*†

By

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Summary—Subaqueous signalling by means of sound waves is far superior to optical, magnetic or electrical means because water is a good medium for the transmission of sound. Therefore, save for certain specific applications, the acoustical method is almost universally used for the transmission of intelligence in water. The transmission of intelligence in water requires the use of underwater loud speakers for converting electrical energy into acoustical energy and underwater microphones for converting acoustical energy into electrical energy. Sound waves have been produced, transmitted and detected by means of these electroacoustic transducers over the frequency range from 20 cycles to 20 megacycles. Dynamic, magnetic and crystal subaqueous loud speakers, and dynamic, condenser, velocity and crystal subaqueous microphones have been developed covering this frequency band.

Introduction

There are four general methods for the transmission of signals underwater, namely: optical, magnetic, electrical and acoustical. Water is very opaque to infrared and ultraviolet light and is not particularly transparent for visible light. Magnetic transmission and detection may be used over relatively short distances. Electromagnetic or radio waves are rapidly attenuated in passing through sea water because it is a good conductor of electricity. Subaqueous signalling by means of sound waves is far superior to the other methods mentioned above because water is a good medium for the transmission of sound waves. Therefore, save for certain specific applications the acoustical method is almost universally used for the transmission of intelligence in water.

The transmission of intelligence in water requires the use of underwater loud speakers for converting electrical energy into acoustical energy and underwater microphones for converting acoustical energy into electrical energy. Sound waves have been produced, transmitted and detected by means of these electroacoustic transducers over the frequency range from 20 cycles to 20 megacycles. It is the

* Decimal Classification: R800 (534).

† The work described in this paper was done in part for the Navy Department under Contract NOrd-504 with Radio Corporation of America.

purpose of this paper to describe underwater loud speakers[1] and microphones covering this frequency range. Subaqueous sound transducers are used in underwater signalling, ranging, sounding and numerous other applications. A subsequent paper will describe some of the applications of underwater transducers.

DYNAMIC SUBAQUEOUS LOUD SPEAKER

The direct radiator dynamic subaqueous loud speaker is a loud speaker designed to operate under water, in which a diaphragm is driven by a voice coil located in a magnetic field. (Figure 1.)

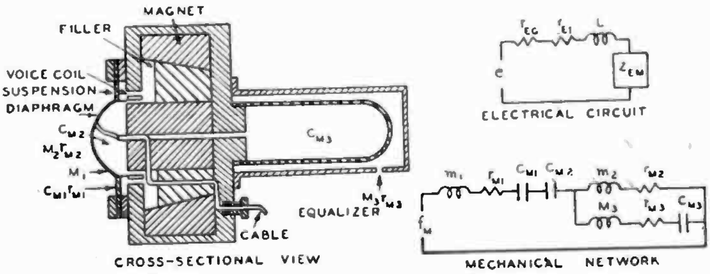

Fig. 1—Cross-sectional view, electrical circuit and mechanical network of a direct radiator dynamic subaqueous loud speaker. In the electrical circuit: z_{EM} = the motional electrical impedance. L and r_{E1} = the inductance and electrical resistance of the voice coil. r_{EG} = the electrical resistance of the electrical generator. In the mechanical network: m_1 = mass of the diaphragm and voice coil. r_{M1} and C_{M1} = the mechanical resistance and compliance of the suspension system. C_{M2} = the compliance of the air chamber behind the diaphragm. m_2 and r_{M2} = the mass and mechanical resistance of the water load. m_3 and r_{M3} = the mass and mechanical resistance of the aperture in the pressure equalizer. C_{M3} = the compliance of the pressure equalizer.

Except for the high impedance of the medium, the theory of the direct radiator dynamic subaqueous loud speaker is the same as that of the air direct radiator dynamic loud speaker. The higher acoustical impedance of the medium makes it expedient to incorporate some con-

[1] In this paper the term subaqueous loud speaker will be used to designate a system for converting electrical variations into the corresponding sound vibrations in water and the term subaqueous microphone will be used to designate a system for converting sound vibrations in the water into the corresponding electrical variations. In underwater sound the terms oscillator, transmitter and projector have also been used to designate a subaqueous loud speaker. The terms receiver and hydrophone have also been used to designate a subaqueous microphone. There is a possibility of ambiguity in the use of these terms. For example, when the term projector alone is used it usually means a still or motion picture projecting system. The same confusion exists in the terms receiver, oscillator and transmitter. Therefore, it is necessary to add the adjective subaqueous in the use of the terms to avoid confusion. Under these conditions, it seems logical to use the terms subaqueous microphone and subaqueous loud speaker. These terms have been used extensively during the past few years.

structional features which differ from the corresponding air type loud speaker. Since the acoustical impedance of water is about 3400 times that of air, the diaphragm of the subaqueous direct radiator loud speaker is relatively small as compared to that of an air direct radiator loud speaker employing a comparable driving system. In the case of the subaqueous loud speaker, it is not necessary to use a large baffle or cabinet. The back of the diaphragm can be terminated in a relatively small volume of air because of the relatively low acoustical impedance of air. As in the air direct radiator loud speaker, the response is independent of the frequency in the frequency region where the acoustical radiation resistance is proportional to the square of the frequency and when the system is mass controlled. In order to obtain mass control down to a relatively low frequency, a compliant or limp suspension system must be employed. A limp suspension will not support any appreciable differential pressure between the two sides of the diaphragm. Since the pressure in water increases about .44 pound per square inch per foot of depth, some means must be provided to maintain uniform pressure on the two sides of the diaphragm if operation at any appreciable depth is desired. In the one form of subaqueous dynamic loud speaker, the equalizing means consists of a limp rubber bag connected to the air space behind the diaphragm. It will be seen that this system automatically provides equal pressure on the two sides of the diaphragm.

An air bubble in the water in close proximity to the diaphragm provides a shunt series resonant acoustical circuit. The result is that very little energy can be radiated at the resonant frequency of the bubble. The presence of a bubble produces a serious dip in the response-frequency characteristic at the resonant frequency of the bubble. Since the compensating chamber is an air bubble, means must be provided to prevent the deleterious effects of the bubble in the response range. This is accomplished by enclosing the limp compensating chamber within a rigid case. A small aperture provides communication between the inside and outside of the case. Referring to the mechanical circuit, it will be seen that if the inertance of this aperture is made sufficiently large, the resonant frequency of m_3 and C_{M3} will occur below the desired response frequency range.

The sound power output, in watts, of the loud speaker is given by

$$P = r_{M2} \dot{x}_2^2 \, 10^{-7} \text{ watts} \qquad (1)$$

where r_{M2} = mechanical resistance due to radiation in mechanical ohms,

$\dot{x}_2$ = velocity in the mechanical radiation resistance in centimeters per second.

The velocity $\dot{x}_2$ is given by

$$\dot{x}_2 = \frac{f_M z_{M3}}{z_{M1}z_{M2} + z_{M1}z_{M3} + z_{M2}z_{M3}} \tag{2}$$

where $f_M = B\,l\,i$ = driving force in dynes,

B = flux density in the air gap in gausses,

l = length of the voice coil conductor in centimeters

i = current in the voice coil in abamperes,

$$z_{M1} = r_{M1} + j\omega m_1 + \frac{1}{j\omega C_{M1}} + \frac{1}{j\omega C_{M2}},$$

$$z_{M2} = r_{M2} + j\omega m_2,$$

$$z_{M3} = r_{M3} + j\omega m_3 + \frac{1}{j\omega C_{M3}},$$

m_1 = mass of the diaphragm and coil in grams,

r_{M1} = mechanical resistance of the suspension system in mechanical ohms,

C_{M1} = compliance of the suspension system in centimeters per dyne,

r_{M2} = mechanical radiation resistance in mechanical ohms,

m_2 = mass of the water load in grams,

C_{M2} = compliance of the air chamber behind the diaphragm in centimeters per dyne,

r_{M3} = mechanical resistance of the hole in the equalizing chamber in mechanical ohms,

m_3 = mass of the water in the hole in the equalizing chamber in grams,

C_{M3} = compliance of the equalizing chamber in centimeters per dyne.

The current, in abamperes, in the voice coil is given by

$$i = \frac{e}{r_{EG} + r_{M1} + j\omega L + z_{EM}} \tag{3}$$

e = voltage of the electrical generator in abvolts,

r_{EG} = damped electrical resistance of the generator in abohms,

L = damped inductance of the voice coil in abhenries,

$$z_E = \frac{(Bl)^2}{z_{MT}}$$

$$z_{MT} = \frac{z_{M1}z_{M2} + z_{M1}z_{M3} + z_{M2}z_{M3}}{z_{M2} + z_{M3}}.$$

The power output or response can be computed from Equations (1), (2) and (3).

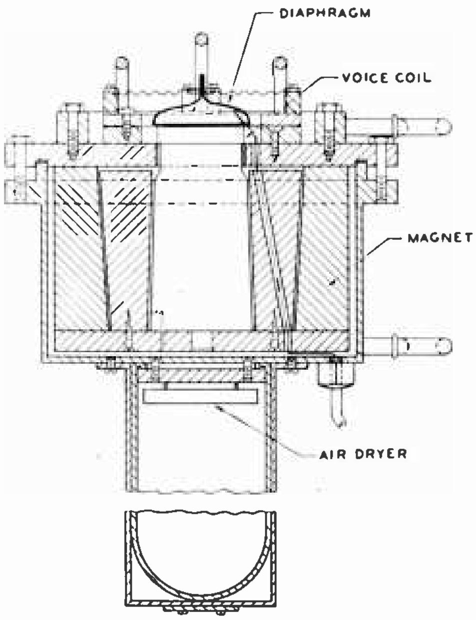

Fig. 2—Sectional view of the direct radiator loud speaker.

The direct radiator dynamic subaqueous loud speaker shown in Figure 2 is designed to operate with maximum efficiency in the audio frequency range. A photograph of the loud speaker is shown in Figure 3. The diaphragm is about two inches in diameter. The air gap flux is supplied by a permanent magnet. The flux density in the air gap is 15,000 gausses. The performance of the system may be deduced from the mechanical circuit and the constants of the system. The

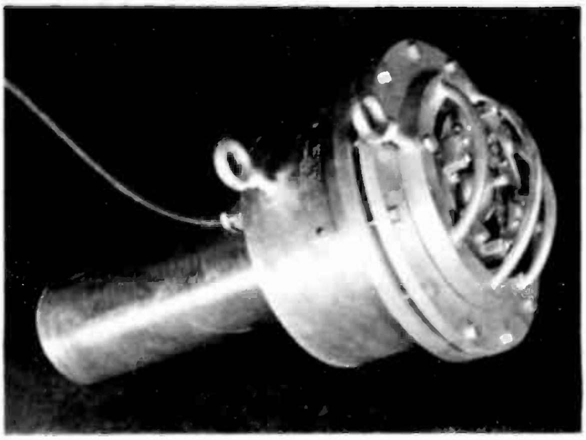

Fig. 3—Direct radiator dynamic subaqueous loud speaker.

measured response-frequency characteristic of the subaqueous loud speaker is shown in Figure 4. The variations in the response in the high frequency range are due to the lack of rigidity of the suspension

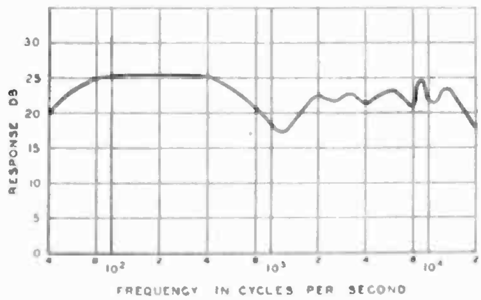

Fig. 4—Response-frequency characteristic of the direct radiator loud speaker.

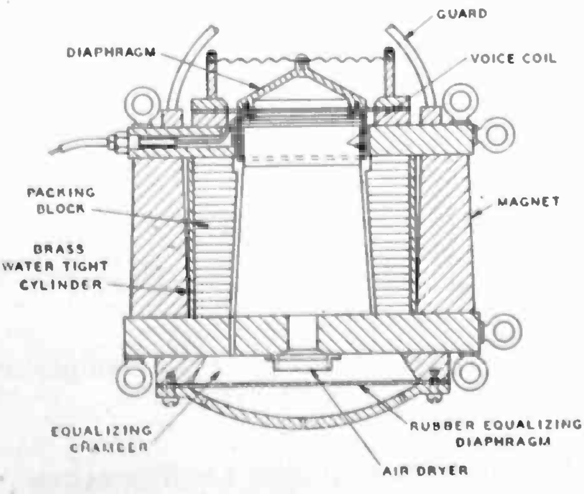

Fig. 5—Cross-sectional view of the high power direct radiator subaqueous loud speaker.

Fig. 6—High power direct radiator subaqueous loud speaker.

system. It may be mentioned in passing, that the response-frequency range is considerably greater than that of an air loud speaker with a comparable driving system.

A larger direct radiator subaqueous loud speaker capable of handling two-hundred watts input is shown in a sectional view of Figure 5. A photograph of the unit is shown in Figure 6. The diaphragm is six inches in diameter. The mass of the voice coil is 190 grams. The flux density in the air gap is 15,000 gausses. The weight of the permanent magnetic material is 500 pounds.

The performance of the system may be predicted by the same theory as that outlined for the smaller direct radiator subaqueous loud speaker. The measured response-frequency characteristic is shown in Figure 7.

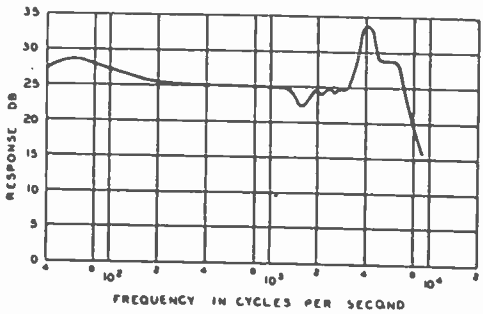

Fig. 7—Response-frequency characteristic of the high power direct radiator loud speaker.

Magnetic Subaqueous Loud Speaker

The magnetic subaqueous loud speaker is a loud speaker designed to operate under water consisting of a resonant diaphragm driven by forces resulting from magnetic reactions. The magnetic subaqueous loud speaker shown in Figure 8 is of the unpolarized armature type. The force on the armature, in dynes, is given by

$$f_M = \frac{C^2 i^2}{4\pi A a^2} \qquad (4)$$

where i = current in the coil in abamperes,

$C = 2\pi n A$,

A = area of the center pole in square centimeters,

n = number of turns in the coil,

a = spacing in centimeters.

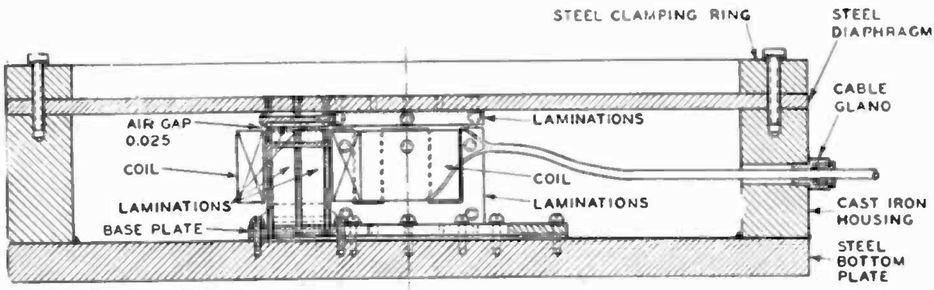

Fig. 8—Cross-sectional view of the magnetic subaqueous loud speaker.

If the current in the coil is sinusoidal, then the expression for the current can be written

$$i = i_{\max} \sin \omega t \qquad (5)$$

where $i_{\max}$ = amplitude of the current in abamperes,

$\omega = 2\pi f$,

f = frequency in cycles per second,

t = time in seconds.

Substituting Equation 5 for the current in Equation 4, the force on the armature is

$$f_M = \frac{C^2}{4\pi A a^2} i^2_{\max} \sin^2 \omega t$$

$$= \frac{C^2}{4\pi A a^2} i^2_{\max} \left(\frac{1}{2} - \frac{1}{2} \cos 2\omega t \right) \tag{6}$$

Equation 6 shows that there is a steady force and an alternating driving force of twice the frequency of the impressed current.

The performance of the system can be determined from a consideration of the mechanical and electrical circuits of Figure 9. The effective mass of the damped plate is one third of the total mass of the plate. The effective mechanical impedance load of the water upon the diaphragm may be obtained by assuming the effective area of the diaphragm is equal to a circular piston having one third the area of the diaphragm.

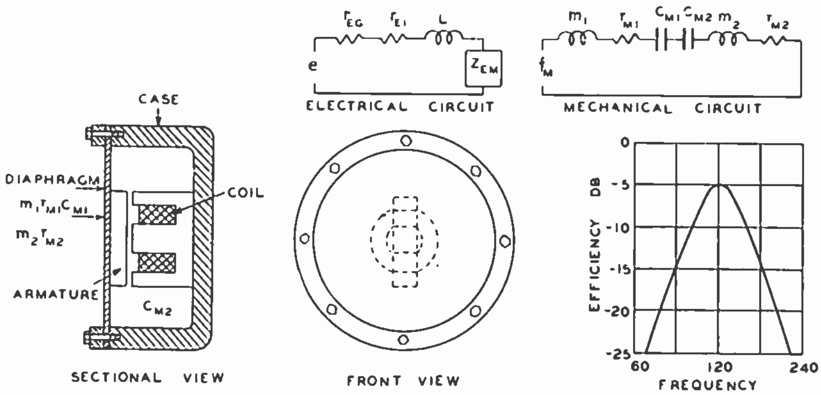

Fig. 9—Cross-sectional and front views, electrical circuit, mechanical circuit and efficiency-frequency characteristic of a magnetic subaqueous loud speaker. In the electrical circuit: z_{EM} = motional electrical impedance. r_{E1} and L = the electrical resistance and inductance of the coil. r_{EG} = the electrical resistance of the electrical generator. e = the voltage output of the electrical generator. In the mechanical circuit: m_1, r_{M1} and C_{M1} = the mass, mechanical resistance and compliance of the diaphragm. C_{M2} = the compliance of the air chamber behind the diaphragm. m_2 and r_{M2} = the mass and mechanical resistance of the water load.

The motional impedance of the system is

$$z_{EM} = \frac{2\pi^2 n^2 A^2 i^2}{a^2 z_M} \tag{7}$$

where z_{EM} = motional impedance in abohms,

n = number of turns,

A = area of the center pole in square centimeters,

i = current in abamperes,

a = spacing in centimeters,

z_M = total mechanical impedance in mechanical ohms.

At resonance z_{EM} becomes an electrical resistance.

The efficiency, in per cent, is

$$\mu = \frac{r_{EM}}{r_{EM} + r_{ED}} \times 100 \qquad (8)$$

where r_{ED} = damped electrical resistance of the coil in abohms,

r_{EM} = motional electrical resistance in abohms.

From Equations 7 and 8, it will be seen that the efficiency increases with the power input. This characteristic is typical of unpolarized driving systems.

A efficiency-frequency characteristic of the magnetic subaqueous loud speaker is shown in Figure 9. A photograph of the loud speaker is shown in Figure 10.

Fig. 10—Magnetic subaqueous loud speaker.

Subaqueous Velocity Microphone

A subaqueous velocity microphone is a microphone designed to operate underwater and in which the response corresponds to the particle velocity in the water. The subaqueous velocity microphone shown in Figure 11 consists of a corrugated phosphor bronze ribbon located in a magnetic field. Permanent magnets are used to provide the air gap magnetic flux. The low electrical impedance of the ribbon is stepped up to an impedance suitable for transmission over a line by means of a transformer. Since the conductivity of the ribbon is low compared to the shunt water path, it is not necessary to insulate the ribbon from the water. One end of the ribbon is connected to the frame of the microphone and the other end is insulated from the frame. Waterproof packing glands are used for the wires which enter the waterproof transformer case.

Fig. 11—Subaqueous velocity microphone with the protective screen removed.

The elements of the vibrating system and the acoustical circuit of the vibrating system are shown in Figure 12. The system is driven by the difference in pressure between the two sides of the ribbon.

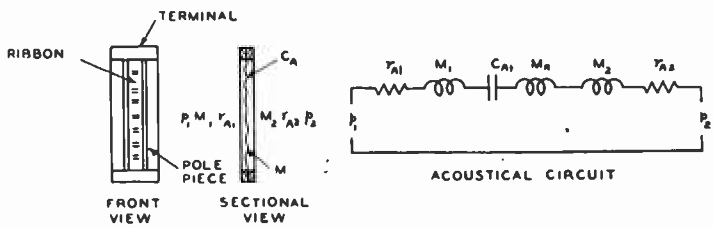

Fig. 12—Front and sectional views and acoustical circuit of the vibrating system of a subaqueous velocity microphone. In the acoustical circuit: p_1 and p_2 = pressures on the front and back of the ribbon. r_{A1} and M_1 and r_{A2} and M_2 = acoustical resistance and inertances on the front and back of the ribbon.

The velocity of the ribbon, in centimeters per second, is given by

$$\dot{x} = \frac{p_1 - p_2 \epsilon^{j\omega\theta}}{A\left[r_{A1} + r_{A2} + j\omega(M_R + M_1 + M_2) + \dfrac{1}{j\omega C_A} \right]} \tag{9}$$

where p_1 = sound pressure at the front of the ribbon,

p_2 = sound pressure at the back of the ribbon,

A = area of the ribbon,

r_{A1} and r_{A2} = acoustical resistances of the water load on the front and back of the ribbon,

M_1 and M_2 = inertance of the water load on the front and back of the ribbon,

M_R = inertance of the ribbon,

C_A = acoustical capacitance of the ribbon,

$$\theta = \frac{d}{\lambda} \cos \phi,$$

d = acoustical path from the front to the back of the ribbon,

λ = wavelength,

ϕ = angle between the normal to the ribbon and the direction of the incident sound.

The acoustical resistances r_{A1} and r_{A2} are small compared to the acoustical reactances of the vibrating system and may be neglected. The fundamental resonant frequency of the system is about 50 cycles. Therefore, the system is mass controlled in the range above the fundamental resonant frequency. If the distance from the front to the back of the ribbon is a small fraction of a wavelength, the pressure which actuates the ribbon will be proportional to the frequency. Under these conditions, the ratio of the velocity of the ribbon to the sound pressure in free space will be independent of the frequency.

The open circuit voltage, in abvolts, developed by the motion of the ribbon is

$$e = B\,l\,\dot{x} \tag{10}$$

where B = flux density in gausses,

l = length of the ribbon in centimeters,

$\dot{x}$ = velocity of the ribbon in centimeters per second.

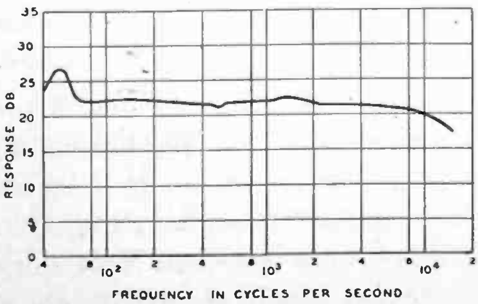

Fig. 13—Response-frequency characteristic of the subaqueous velocity microphone.

The measured response-frequency characteristic of the subaqueous velocity microphone is shown in Figure 13. It will be seen that the response is quite uniform with respect to frequency. The directional characteristic is of the cosine type as predicted by Equation 9.

Subaqueous Condenser Microphone

The subaqueous condenser microphone is a microphone designed to operate under water and which depends for its operation upon variations in electrical capacitance.

The subaqueous condenser microphone described in this paper was developed with the object of providing a microphone which could be used as an absolute standard. The specifications for this microphone included the following. The response was to be flat to within ± 2 decibels from 100 to 3000 cycles per second, and to within ± 6 decibels from 3000 to 8000 cycles per second. Its construction was to be such that the microphone could be submerged in water to a depth of 100 feet for a period of one year without servicing. During this time, the sensitivity was to remain constant, or means was to be provided for determining its sensitivity electrically at any time from a surface station. Furthermore, the unit was to be rugged enough to withstand the pressures due to moderate under water explosions without becoming inoperative. Taking these requirements into consideration, it appeared that a condenser microphone with an associated preamplifier would be the most suitable type to use. The diaphragm of a subaqueous condenser microphone consists of a relatively thick clamped plate. When submerged, it is stiff enough to withstand the static pressure of the water load with only a negligible amount of bending in so far as its operation is concerned. Furthermore, the clamped portion of the diaphragm and its fixed plate (the high potential electrode) can be rigidly fixed in relation to one another thus insuring a constant sensitivity for the microphone proper. The associated amplifier can be provided with means for an electrical calibration from a remote location and thus the sensitivity of the complete unit can be determined at any time.

In the design of a subaqueous condenser microphone, it is desirable to fix the resonant frequency of the diaphragm above the highest operating frequency in order to keep the response vs. frequency characteristic as flat as possible throughout the operating range. On the other hand, the diaphragm resonant frequency should not be placed too high or an excessive loss in sensitivity due to stiffness of the diaphragm will be incurred. Having chosen a diaphragm diameter to give a suitable directional pattern and adequate sensitivity, it then becomes necessary to calculate the diaphragm thickness such that the required resonant frequency will be obtained. The fundamental frequency of a clamped circular plate is given by

$$f = \frac{0.467\,t}{R^2}\sqrt{\frac{Q}{\rho(1-\sigma^2)}} \tag{11}$$

where t = thickness of the plate in centimeters,

R = radius of the plate up to the clamping boundary in centimeters,

ρ = density in grams per cubic centimeter,

σ = Poisson's ratio,

Q = Young's Modulus in dynes per square centimeter.

Choosing a resonant frequency in air somewhat higher than that desired in water, the thickness of the diaphragm can be calculated from Equation 11.

Taking the equivalent mass of the diaphragm as one-third the total unclamped mass, the compliance of the diaphragm can be calculated from the equation

$$C_M = \frac{1}{4\pi^2 f^2 m} \tag{12}$$

where C_M = compliance in centimeters per dyne,

f = resonant frequency in cycles per second,

m = equivalent mass in grams.

To the equivalent mass of the diaphragm must now be added the mass of the water load. This sum is the total mass of the vibrating system. The resonant frequency of the diaphragm in water can be calculated from Equation 12 using this total mass for m and the value of C_M previously obtained. If the resonant frequency in water thus obtained falls within the operating range of the microphone, a higher resonant frequency in air must be chosen and the calculations repeated.

A cross-sectional view of the subaqueous condenser microphone

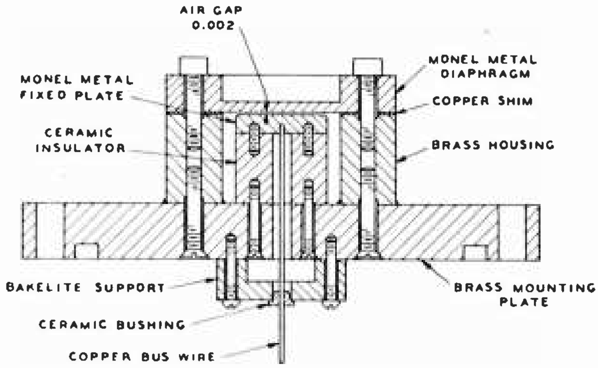

Fig. 14—Cross-sectional view of subaqueous condenser microphone head.

head is shown in Figure 14. The diaphragm and clamping ring are integrally turned from a single piece of material thus insuring perfect clamping and a minimum of fringing at the boundary of the movable part of the diaphragm. The fixed plate is insulated from the electrically grounded parts of the unit by four ceramic insulators. The diaphragm and clamping ring are bolted to a brass housing. The outer face of the housing and the fixed plate are ground so as to lie in the same plane. Accurate air gap spacing between the diaphragm and the fixed plate is obtained by means of a shim in the form of a flat circular ring between the clamped portion of the diaphragm and the surface of the housing. The housing and fixed plate assembly are rigidly fastened to a heavy brass mounting plate so that these parts are firmly fixed in place with respect to each other. The brass mounting plate also serves as the top of the amplifier case. An external view of the microphone is shown in Figure 15. A photograph of the microphone with the case removed to show the preamplifier is shown in Figure 16.

A schematic diagram of the preamplifier is shown in Figure 17. The preamplifier is two-stage resistance coupled. The polarizing voltage is applied to the fixed plate of the condenser microphone

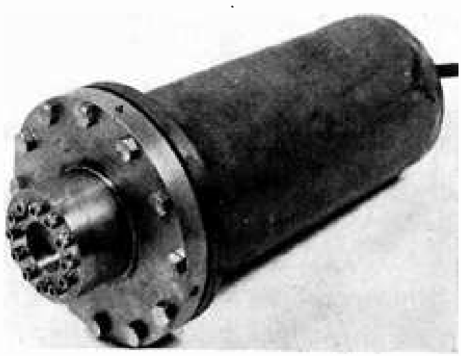

Fig. 15—Subaqueous condenser microphone.

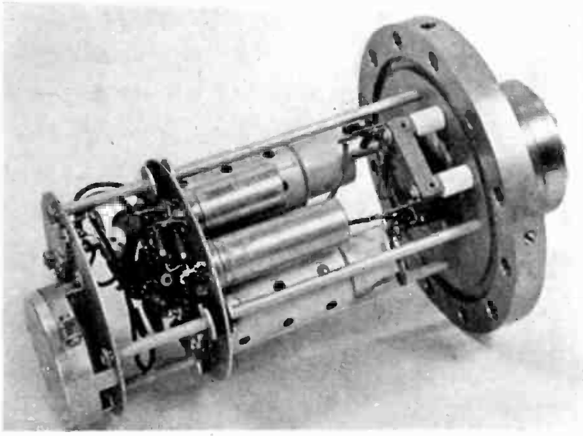

Fig. 16—Subaqueous condenser microphone with case removed to show preamplifier.

through the resistor R3 and the resistance-capacity filter R2, C1. Since the capacity of the microphone forms the internal impedance of the generator, the parallel impedance of R3, R4 and the input impedance of V1 must be large compared to the impedance of the condenser microphone in order to keep the response from falling off at low frequencies. The sensitivity can be measured at any desired frequency by passing a known current through the calibrating resistor R1. The voltage drop in R1 is thus applied to the grid of V1. By comparing this voltage with the voltage across the external load resistor, the sensitivity is obtained. All voltages to operate the preamplifier are supplied through a multi conductor cable from batteries located at the surface.

Response vs. frequency characteristics of the subaqueous condenser microphone head, the preamplifier, and the sum of the two are shown in Figure 18. The preamplifier characteristic is shown as the ratio of the output voltage developed across the external load resistor to the calibrating voltage developed across R1. The overall characteristic is an acoustic response curve obtained by placing the microphone in a

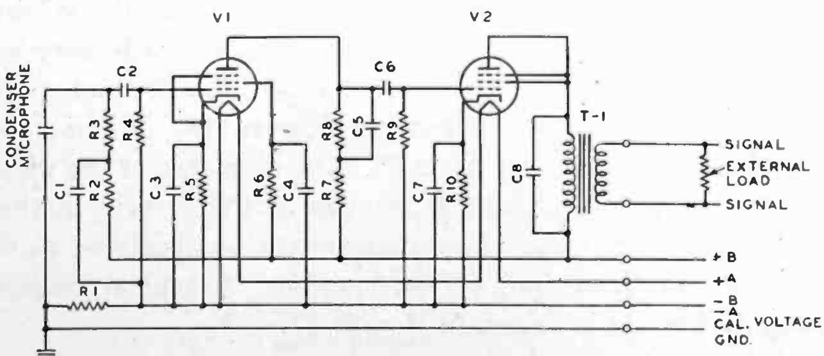

Fig. 17—Schematic circuit diagram of subaqueous condenser microphone preamplifier.

sound field of known intensity. The characteristic of the condenser microphone head is the difference between the other two characteristics. It will be noted that the useful overall response at the high frequency end of the range is extended by giving the preamplifier a drooping characteristic to compensate for the rising characteristic of the microphone head caused by the fundamental diaphragm resonance.

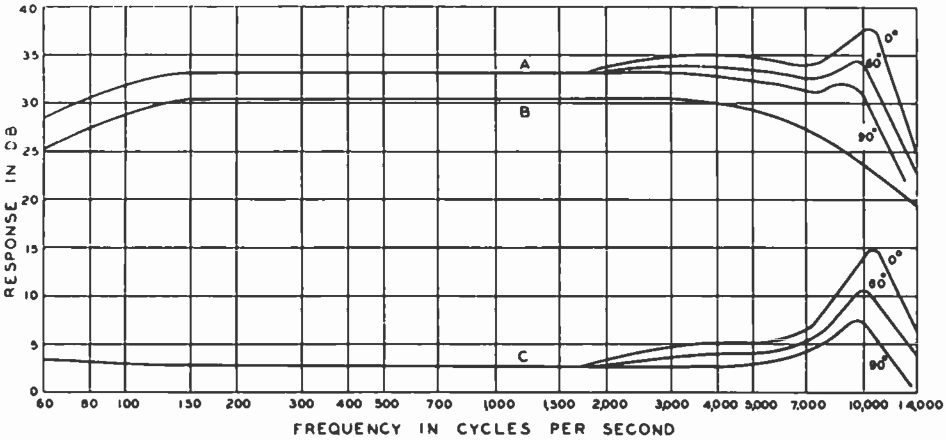

Fig. 18—Response-frequency characteristics of subaqueous condenser microphone.

A. Overall acoustic response for constant sound pressure on the axis of the microphone, 60° off the axis, and 90° off the axis.

B. Response of preamplifier.

C. Response of condenser microphone head for constant sound pressure on the axis of the microphone, 60° off the axis, and 90° of the axis.

High Frequency Dynamic Subaqueous Loud Speaker

The response of the dynamic loud speakers described in the preceding sections falls off above 15,000 cycles. In the frequency range above 10,000 cycles, it is possible to employ a smaller diaphragm because the amplitude for moderate power requirements is relatively small. Furthermore, the suspension system can be made very stiff and rigid and still retain mass control. A typical high frequency dynamic subaqueous loud speaker is shown in Figure 19. A cross-sectional view of the unit is shown in Figure 20. The diameter of the diaphragm is ¾ of an inch. The resonant frequency of the system in the water is about 12,000 cycles. The performance may be predicted as outlined in the section on the dynamic loud speaker. A typical response-frequency characteristic is shown in Figure 21.

The same unit may be used as a microphone over the same frequency range.

Fig. 19—High frequency dynamic subaqueous loud speaker.

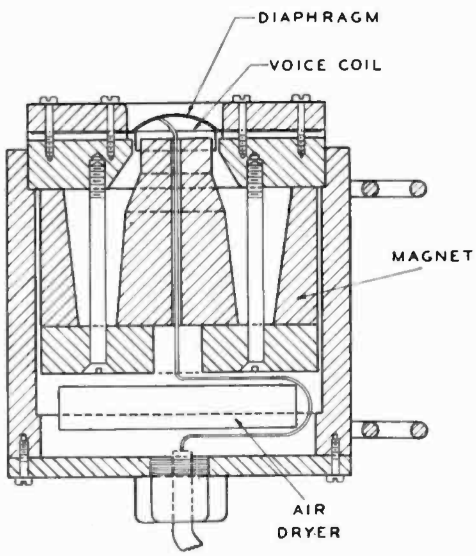

Fig. 20—Cross-sectional view of the high frequency dynamic subaqueous loud speaker.

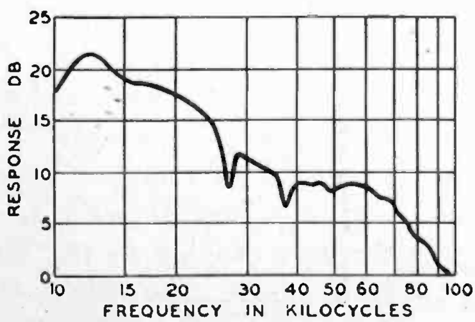

Fig. 21—Response-frequency characteristic of the high frequency dynamic loud speaker.

SUBAQUEOUS QUARTZ CRYSTAL MICROPHONE AND LOUD SPEAKER

A quartz crystal subaqueous transducer is a transducer designed to operate underwater which depends upon the converse piezoelectric properties that exist between mechanical and electrical forces in a quartz crystal.

In considering possible ways to make a transducer for the higher ultrasonic frequency range, the piezoelectric crystal seems to offer the best possibilities. Since stability under a host of conditions is one of the most desirable traits of a measuring transducer, quartz was selected as the best piezoelectric material. As is well known in the art of endeavor in the field of underwater sound, the simple systems offer: the best chances of achieving freedom from anomalous responses; reasonable ease in estimating response and sensitivity; and (perhaps most important of all) ease of construction. In this particular trans-

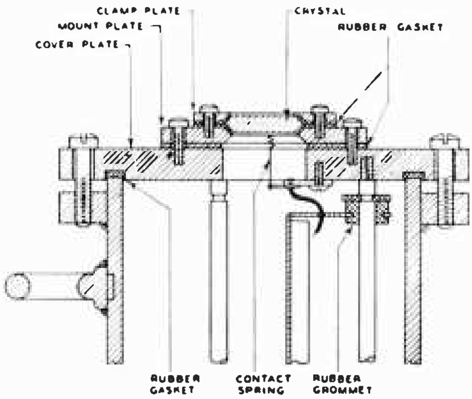

Fig. 22—Cross-sectional view of the quartz crystal subaqueous microphone.

ducer, the simplest construction conceivable is used. A circular x-cut quartz crystal is mounted in a watertight seal so that one face is exposed to the water which serves as the outer electrode. The opposite face of the crystal is coated with an evaporated aluminum electrode and is exposed to the air within the transducer case. The assembly is shown in Figure 22. A light phosphor-bronze spring serves to provide electrical contact to the aluminum coated crystal face.

For use of the transducer as a microphone, an amplifier and a detector were built and included in the same case with the transducer. A schematic diagram of the amplifier is shown in Figure 23. The amplifying stages are coupled with tuned circuits which have a Q such that the amplifier response brackets the transducer response. A diode detector is provided and coupled to the output terminals by means of a cathode follower stage. The output signal is, therefore, a direct current proportional to the sound pressure at the transducer outer face.

For use of this transducer as a loud speaker, the energizing signal was applied directly to the crystal by means of a concentric cable.

Matching microphone and loud speaker combinations were made for the following frequencies: 100, 200, 500, 1000, 2000, 5000, 10,000, 15,000 and 20,000 kilocycles. The crystal dimensions, for frequencies 100 to 10,000 kilocycles inclusive, were made inversely proportional to frequency so that the directivity would be the same for all models. All of these units operated on the fundamental thickness resonance of the crystal. The 15,000 kilocycles unit was operated on the second harmonic of the crystal thickness resonance. The 20,000 kilocycles unit was operated on the third harmonic of the crystal thickness resonance. For practical reasons, the lateral dimensions of the 15,000 and 20,000 kilocycles transducer crystals were made the same as the 10,000 kilocycles model.

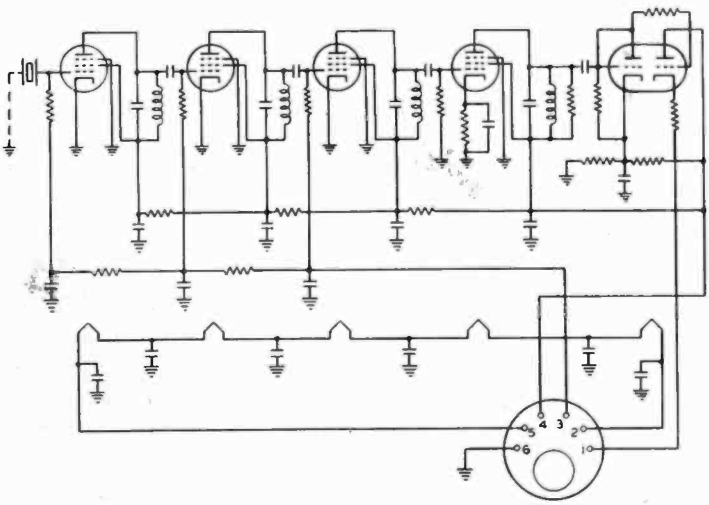

Fig. 23—Circuit diagram of the amplifier for the quartz crystal subaqueous microphone.

The response, sensitivity and directivity of this type of transducer were found to be in good agreement with theoretical calculations. The frequency and directivity characteristics of the 500 kilocycle microphone are shown in Figures 24 and 25. The sensitivity of the transducers as microphones (volts across the crystal electrodes at resonance) in microvolts per dyne per square centimeter ranged from 141 for the 100 kilocycle model through 45 for the 500-kilocycle model to 0.04 for the 20,000-kilocycle model. These sensitivities were determined by a reciprocity method.

Figure 26 shows the amplifier and transducer assembly of a 500-kilocycle model microphone. Figure 27 shows the 100-kilocycle microphone assembled in a watertight case.

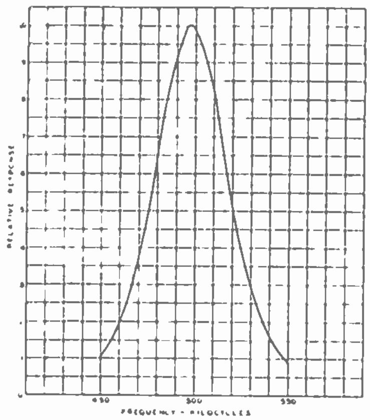

Fig. 24—Response-frequency characteristic of the quartz crystal microphone.

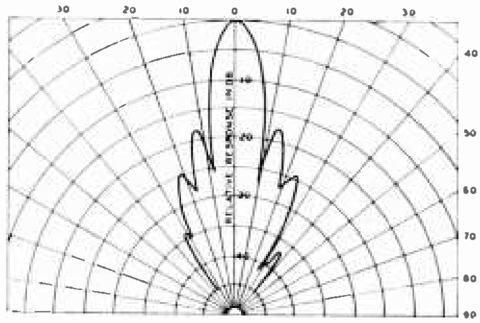

Fig. 25—Directional characteristic of the quartz crystal microphone.

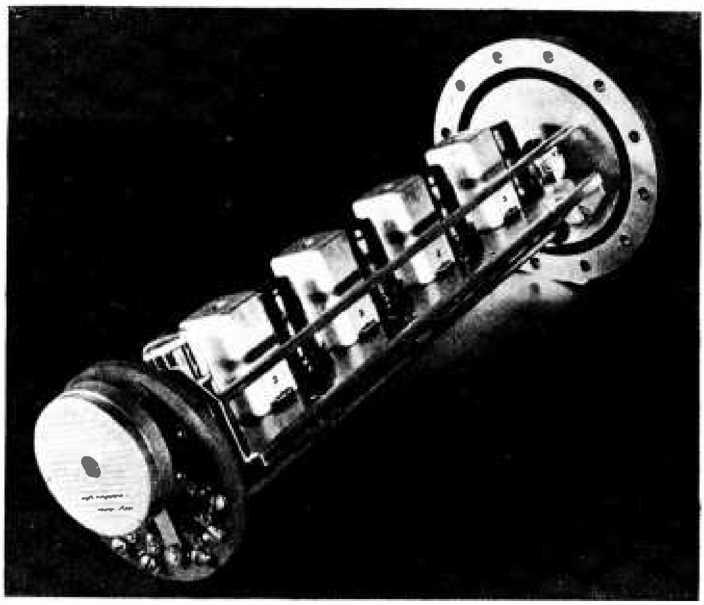

Fig. 26—Amplifier and quartz crystal subaqueous microphone assembly.

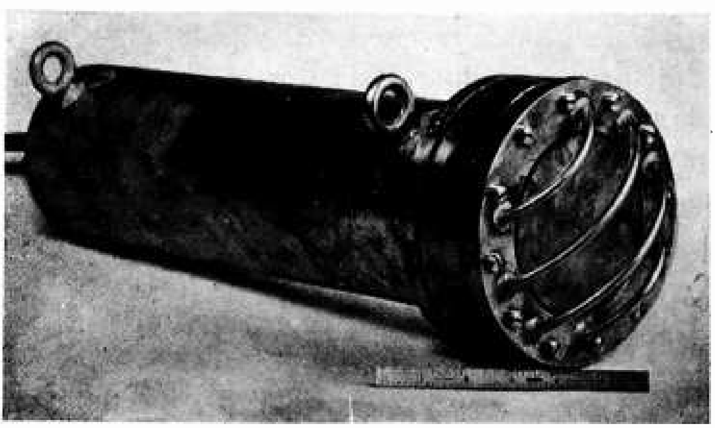

Fig. 27—Quartz crystal subaqueous microphone.

TECHNICAL ASPECTS OF TELEVISION STUDIO OPERATION*

By

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Summary—*This article describes the operating procedures employed in a television studio for the presentation of a live talent program. It is divided into two parts, one covering the operation of the technical equipment and the other describing the improvements that have been incorporated in the equipment as a result of operating experience. This latter section includes discussion of camera dollies, lighting, microphones, microphone boom, iconoscope cameras, dialogue equalization, audio perspective, transcription turntables, communication and cue systems, and the control room.*

Part I—Operation of Technical Equipment[1]

THE smooth operation of a television studio requires a high degree of cooperation and teamwork on the part of all personnel involved. This paper deals principally with technical phases of this problem. Proper technical operation should be completely unobstrusive, and generally speaking, should be of such a nature that it never diverts attention from the program itself.

The technical personnel in the live talent studio include the following:

(a) *In the control room:*

Technical Director
Video Control Engineer
Audio Control Engineer
Transcription Turntable Operator

(b) *In the studio:*

Camera Operators
Microphone Boom Operator
Light Direction Engineer
Light Operators (electricians)
Camera Dolly Operator

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* Decimal Classification: R583.2 × R583.3.

[1] A. W. Protzman, "Television Studio Technic," *RCA REVIEW*, Vol. IV, No. 4, pp. 399-413, April, 1940.

The titles are indicative of the duties performed by the individuals, but more detailed information on their duties and on the equipment operated by the personnel is included in this article.

From the inception of a program, the Technical Director works and confers with the Program Director, learning the sequence of scenes, desired camera shots, effects, etc. The Technical Director is available to give information and answers on technical problems. He is present on "dry rehearsals" (rehearsals without cameras) to determine the necessary movements of the cameras, microphone boom, and lighting equipment. In the studio the Technical Director is responsible for the technical quality of the program. He performs the camera switching, including dissolves, (as determined by the program director), previews monitor switching and issues directions to the engineering operating personnel.

The Video Control Engineer is responsible for all the controls for the cameras. He adjusts beam current, focus, deflection, centering, back light, shading, brightness and contrast and executes the desired video fade-ins and fade-outs. He communicates with the Light Direction Engineer on the lighting of the sets.

The Light Direction Engineer operates from the light bridge, an elevated platform in one corner of the studio. He adjusts the light units[2] that are mounted on the studio ceiling and has control switches for all light units. He directs the lighting personnel on the studio floor for proper placement of the floor units, and is also responsible for the placement and operation of special light units, such as spots. He takes precaution to see that all units are operating satisfactorily.

Each Camera Operator makes certain his assigned camera is in proper operating condition. He checks the iconoscope lens optical focus with the viewfinder lens focus and with the aid of the Video Control Engineer determines that his viewfinder frame corresponds to the picture obtained from his camera. In operation the Camera Operator positions his camera and maintains optical focus. Camera positions are determined by the Program Director in order that the scene and angle may be as he desires.

For sound pickup, a microphone suspended from the end of an arm, or boom, is employed. The boom is mounted on a movable support. The Microphone Boom Operator positions the boom and microphone to obtain the best sound. The microphone is raised or lowered according to the size of scene being televised, and moved in and out or back and forth to follow the actors. The floor position of the microphone

[2] W. C. Eddy, "Television Lighting," *RCA REVIEW*, Vol. IV, No. 4, pp. 414-424, April, 1940.

boom is determined by two conditions: the placement and movement of cameras and performers. It must be possible at all times to bring the microphone close enough to the performers to obtain a satisfactory pickup but the position of the boom should not interfere with the cameras.

The Audio Engineer maintains the proper sound level, as in standard broadcasting, and mixes the sound from all other sources such as transcriptions, additional microphones, film and outside pickups. He also controls the adjustments required for the dialogue equalizers and perspective filters. Further descriptions of these devices are given in Part II.

The Transcription Operator performs the functions of selecting and changing records, spotting the stylus at the desired point on the record, starting the transcription at the proper moment and turning the fader to the "ON" position. He must set the turntable speed to that for which the record is intended to be played. For special purposes the turntable can be operated at speeds somewhat different than standard. A description of this apparatus is given in a later section of this paper.

In some productions from the live talent studio, film inserts are provided to maintain continuity of the program or to provide a short time period for scene changes. To facilitate operations for these portions of the program, the video and sound controls are not switched to the film studio. By means of patch cords and additional relays the film studio is set up as an additional camera for the live talent studio. Sound from the film studio is fed to a fader on the audio control console. The video output of the film studio may be previewed in the live talent studio control room so that when the time comes to switch to the film insert, the Technical Director does the video switching and the Audio Engineer opens the proper fader to bring in the sound.

If the program includes a singer, there are two methods of providing accompanying music. If there is a record of the desired number made by the artist, this is used and the music fed into the studio loudspeaker so the movements of the artist will synchronize with the music and voice. The studio microphone is faded out so no feedback will occur. The second method is utilized when it is desired to mix the vocal portion (live) from the studio with the music accompaniment from a record. Audio output from the studio microphone is mixed with the turntable output to provide the broadcast sound. In addition, the turntable sound is fed to the studio loudspeaker so the artist can synchronize with the accompaniment. The level on the studio loudspeaker must be low to avoid feedback.

When special effects in the sound are desired, such as the representation of thoughts, recordings of the actors' voices are made prior to the show, employing filters necessary to produce the desired effect. These recordings are then played at the correct time in the program while the actor remains silent. In order for the cast to know when to resume speaking the actors may receive their cue from the Stage Manager or the recording may be fed into the studio loudspeaker. This latter method requires the Audio Engineer to fade out the studio microphone during the playing of the recording.

Musical portions are used for background accompaniment to provide desired moods. At times, the musical portions may be short, or only brief sections of some opus. The Transcription Operator needs a good sense of timing to be able to start the transcription on the proper cues.

Studio rehearsals enable the operating personnel to learn their required movements and proper sequence of scenes. From an operating standpoint it is essential that all personnel know the transitions from one scene or set to the next so that cameras, cables, and microphone boom will always be clear of each other.

The presentation of a studio dramatic show is now considered. The Program Department arranges for the production, assembles the staff and creates and manufactures the scenery. After the cast has learned their lines "dry rehearsals" are held. During these rehearsals the Program Director guides the actors in their movements to obtain the desired effect. The Technical Director attends these "dry rehearsals" and obtains from the Program Director the desired camera angles and shots. The Technical Director coordinates the necessary mechanical movements of camera and microphone boom so that there will be no confusion in changing from one scene to the following scene.

Usually on the day the program is to be presented, rehearsals are held in the television studio. The first rehearsal, called the walk-through, is usually done without the use of equipment. During this rehearsal the operating staff is assembled on the studio floor, watches the rehearsal and is given information as to special shots that are desired. This enables them to know what is desired during the entire program.

Following this walk-through, regular rehearsals with equipment begin. The Program Director in the control room talks with his Floor Manager who passes cues and instructions to the cast. The Technical Director gives the Camera Operators instructions and portions of the drama are rehearsed until it is satisfactory. The Video Control Engineer and the Lighting Engineer make certain the lighting is adequate and gives the effect desired by the Program Director. The

Microphone Boom Operator is responsible for keeping the microphone as close to the actors as is possible without having it show in the picture. It may be lowered when a close-up shot is being shown and must be elevated when a wide angle scene is shown. The Audio Control Engineer may cue the Boom Operator just prior to these camera switches in order to aid him in keeping the microphone out of the pictures. During rehearsal, the Boom Operator can learn when the microphone is to be raised and lowered and operate accordingly during the program.

Cues for music, titles, fades (both audio and video), lap dissolves, etc. all originate in the control room and are transmitted to the individuals who require this information.

The Video Control Engineer, Audio Engineer and Transcription Turntable Operator receive their cues directly since they are in the control room. The Program Director reminds the Technical Director of coming camera switches and he, in turn, cues the Camera Operators on coming shots. The Video Control Engineer informs the Lighting Engineer and the Audio Engineer cues the Microphone Boom Operator. The Stage Manager receives his cues by intra-studio radio directly from the Program Director. In this manner all operating personnel are constantly in touch and know what is occurring.

PART II—IMPROVEMENTS IN TECHNICAL EQUIPMENT

Since the original design of the Radio City television plant[3] twelve years ago, many changes and improvements have been made in order to keep abreast of the program requirements and take advantage of the experience gained. Comparison of Figure 1 (a photograph taken eleven years ago) with Figure 2 (taken recently) indicates some of the changes in and additions to the studio equipment. The detailed discussion that follows will show the needs and reasons for the changes.

Camera Dolly

The camera shown in the foreground of Figure 1 is mounted on a dolly. With this arrangement it was possible to make dolly shots (i.e. to move the camera smoothly and continuously toward or away from the scene while the camera was "on the air"). It became desirable, however, to have angle shots that could not be obtained with this unit. A small sized crane-type standard motion picture dolly was provided for one of the cameras which allowed an increase in the vertical motion of the camera. In addition the boom can be rotated horizontally

[3] R. E. Shelby and R. M. Morris, "Television Studio Design," *RCA REVIEW*, Vol. II, No. 1, pp. 14-29, July, 1937.

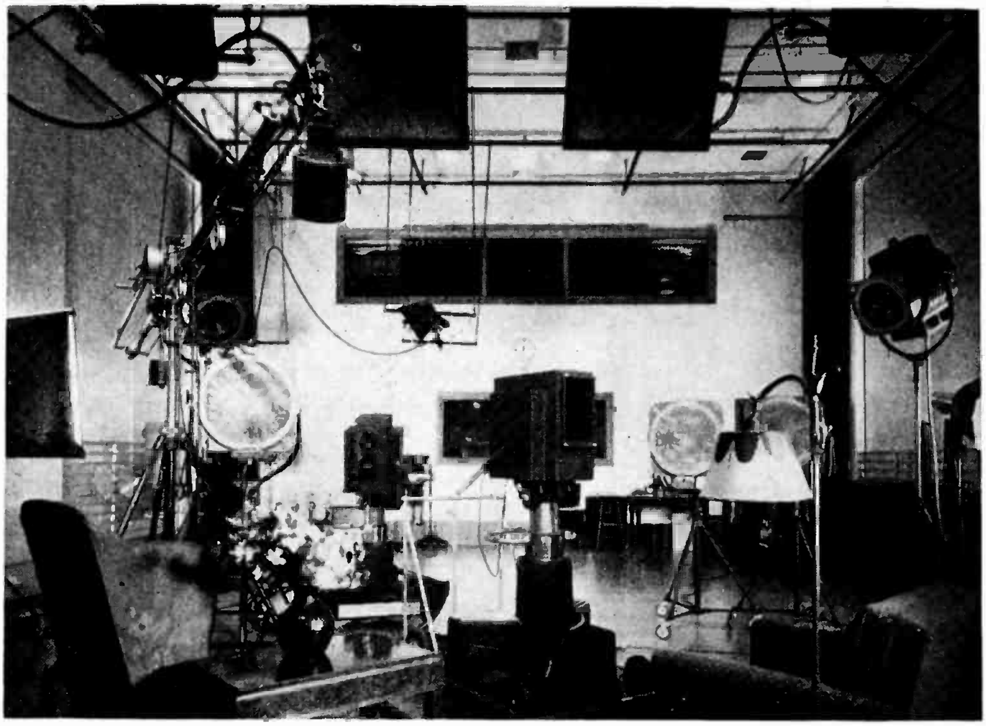

Fig. 1—Television studio—1936.

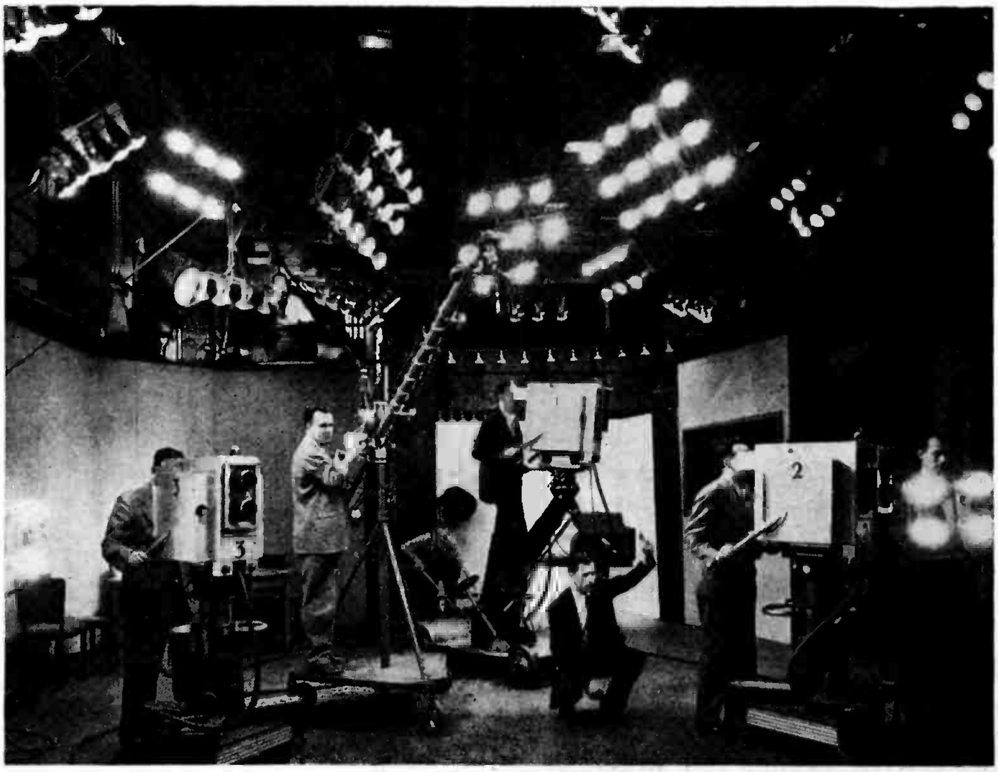

Figure 2—Television studio—1947.

360 degrees permitting shots alongside the dolly—scenes that could not have been obtained readily with the old unit.

Lighting

When the studio was first used, a large set in one end of the studio, and a small set could be satisfactorily illuminated. As the quantity and quality of programs increased the sets became larger and more numerous. At the present time it is not unusual to have all four walls lined with sets—in some places, two deep. Thus an evening's program schedule may require ten sets. This has had the net effect of reducing the limited floor space. To eliminate floor units as much as possible, lighting fixtures were installed in the ceiling.[2] The angle or direction of these lights may be changed rapidly by a Lighting Engineer on what is called a light bridge (upper center of Figure 2). From this position all ceiling lighting units can be turned "On" or "Off" and rotated or angled to any desired position. A catwalk from the lighting bridge around the studio sufficiently high to clear the scenery was installed later. This provided a mounting for additional light units such as spots or floods needed for special effects and a walk from the bridge for adjustment of these units.

Some floor units were retained for reducing shadows cast by the overhead lighting units on the eyes and under the chins of the actors. If the cameras are moved toward the actors, the floor units are moved in also.

The light bulbs now used are a combination of photofloods and photospots. These give the most satisfactory results from the standpoint of efficiency and life.

Microphones

Best sound quality is obtained by utilizing one microphone on a boom. During a program this microphone is positioned by an operator so it is above and a little in front of the actor who is talking. A sound broadcast studio microphone of the usual type is not wholly suitable for television usage because of its weight and size; also its mounting is such that rapid motion tends to introduce appreciable rumble into the sound channel.

The sound absorption frequency characteristics of a television studio differ from a sound broadcast studio in that the studio itself is designed to contribute substantially nothing to the sound quality as this will be determined largely by the characteristics of the set itself (and the microphone channel employed). The set, dependent upon whether it is, in effect, an enclosure as in the case of a "box" arrangement or three-sided arrangement, may reflect selectively appreciable high-frequency energy. The microphone is usually five to eight feet from the actors and located above their heads so that there is a tendency, because of this location, to discriminate against the higher audio frequencies. A small lightweight microphone has been developed by RCA Laboratories Division for experimental use in the television studio. The low frequency response of this microphone was reduced somewhat and with the use of improved magnetic circuits and close spacing of the ribbon it was possible to design a unit weighing less than eight ounces. The microphone is directional, reducing unwanted sound and noises from the rear of the microphone. This directional feature for reducing studio noise is important when it is remembered

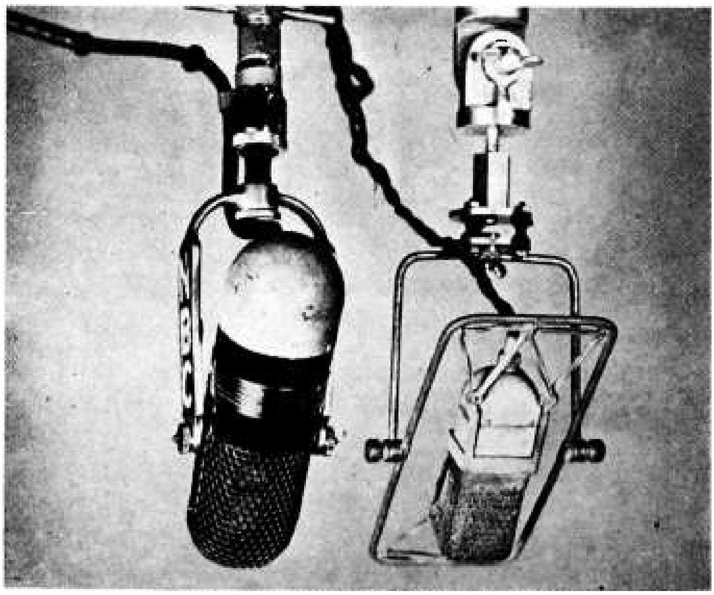

Figure 3—Standard and experimental type microphones.

that in television the microphone is outside the picture and consequently several feet away from the actor. From eight decibels to twelve decibels additional gain is required over that used in a sound broadcast studio. Figure 3 illustrates the size of a standard studio microphone in comparison to the special microphone developed for television.

Microphones in fixed positions are used for an announcer or commentator who is not being televised or whose position is fixed. Hanging microphones will be used when there is a change in pickups between sets and there is insufficient time to move the microphone boom into position for the second set. The actors starting the sequence on the second set must stay near the hanging microphone until the boom microphone can be positioned properly. It is also necessary that the characteristics of the hanging and boom microphones be almost identical so that in switching from one to the other there is no noticeable change in the quality of the sound.

Microphone Boom

Two minor but important changes have been made in the microphone boom. Originally the two connecting cables (microphone output and cue circuit to the operator) lay on the studio floor. This occasionally hampered the movement of the cameras and floor lighting fixtures and was generally in the way. This problem was solved by suspending both cables from the studio ceiling to the top of the pedestal of the microphone boom. The slack in the cables between the pedestal and the wall receptacles is taken up with no noise by a simple pulley and counterweight arrangement located on one of the studio side walls. The pulley and counterweight arrangement was chosen rather than a spring take-up device because of its simplicity and reliability.

To cover action taking place on the studio floor it became necessary to raise the boom or arm section of the microphone boom sufficiently to clear the top of the cameras and lighting fixtures. This increase in height necessitated either lowering the controls so they could be reached conveniently from the studio floor, or providing a platform for the operator on which to stand. The platform was decided upon as the simpler of the two and had the added advantage of increasing the vision of the operator above the cameras.

Iconoscope Cameras

The cases of the television cameras as originally built were painted a dull black. The cases absorbed heat from the television lights and during long periods of operation became hot enough to cause physical discomfort when touched by the Camera Engineers. Also, the internal temperature of the cameras rose to such a value as to affect the performance of the iconoscopes and some of the components in the amplifiers and associated equipment. Repainting the camera cases with a special highly reflective aluminum paint greatly reduced the heat absorption and decreased the internal temperature of the cameras to a suitable operating value.

Originally the camera cable entered the camera pedestal near the bottom of the pedestal and spiralled upward inside the pedestal to the camera plug under the camera. When a cable failure occurred within the spiral it was impossible to make temporary repairs, and it was a major operation to change camera cables. The present cable has been entirely removed from within the camera pedestal so, in the case of failure, the cable may be replaced with a spare in a matter of minutes.

Figure 4 shows the television camera with the pedestal and cable terminating block disassembled.

Dialogue Equalization

In the usual audio broadcast studio the acoustical treatment of the walls and ceiling with the consequent optimum reverberation time is the result of careful planning in the original studio design. Great care is taken to insure that the audio broadcasts are made under as ideal

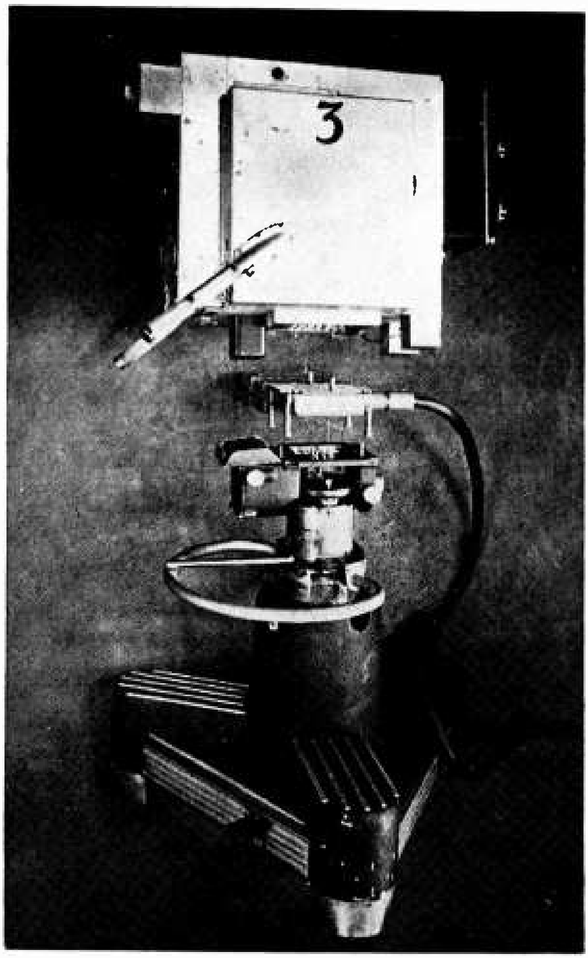

Figure 4—Cable terminal at camera.

conditions as possible. To maintain the desired reverberation time in the larger audio broadcast studios, it is not unusual to select the type of studio chairs so that there will be little change in the quality of the sound pickup whether or not an audience is present. However, in television studios of relatively small size such exact control of the acoustics is not feasible. The acoustical quality of the sound pickup is largely determined by the nature of the sets, the direction and distance of the microphone from the source of sound, and the characteristics of the microphones. Where the walls of the studio are to a great extent covered with sets the acoustical treatment of the walls will have only a limited effect upon the quality of the sound pickup. However, walls and ceilings designed for high sound absorption are desirable to provide the acoustical conditions to simulate outdoor conditions when necessary. This condition may be modified for interior scenes by an increase in reverberation resulting from the surrounding sets and scenery. In addition to improving the acoustical quality of the sound accompanying the picture, high absorption on the walls and ceiling helps to reduce the ambient noise level caused by the movement of personnel and equipment during operations. Where the quality of the sound pickup is noticeably different between sets, due to the resonance effects caused by the particular arrangement of sets, the use of filters to attenuate the lower audio frequencies has permitted a better balance to be obtained. These filters, or dialogue equalizers as they are usually called, have very little effect above 500 cycles with the loss gradually increasing as the frequency is decreased until at 60 cycles the loss is

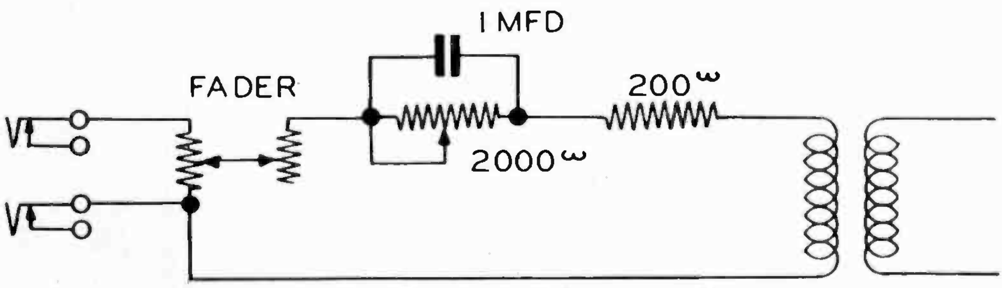

Figure 5—Circuit of dialogue equalizer.

approximately 6 decibels. The circuit for a dialogue equalizer is shown in Figure 5. The equalizers have proved to be of considerable help when balancing between an unusually heavy and low resonant male voice and a lighter and higher feminine voice. In the latest revision of the audio console the addition of dialogue equalizers for each microphone position has allowed the Audio Engineer considerable control over the quality of the pickup that otherwise would not be possible. The equalizers, being of the continuously adjustable type, permit the Audio Engineer to make reasonable changes in the amount of equalization employed while the microphone is being used on the air.

Audio Perspective

The received quality of a source of sound depends a great deal on the distance between the source and the listener. When switching between "long-shot" and "close-up" cameras in television, it is desirable to have the audio quality change at the same time—to create the aural illusion of a change in distance at the instant the effective visual change in distance is made. Specially designed units, called audio perspective filters, are used to create this illusion. The filter consists of a parallel coil and condenser in series with a variable resistor shunted across the fader circuit (see Figure 6). The perspective filter decreases both the high and low frequencies. Perspective may be obtained either by this method or by a change in audio level (a rough method) or by a combination of the two.

Transcription Turntables

There are numerous occasions when performers in the studio desired their accompanying music to be played at a faster or slower tempo than that obtainable from the standard turntables. This problem was solved by constructing an oscillator the frequency of which could be varied in one-half cycle steps from 58 to 64 cycles per second.

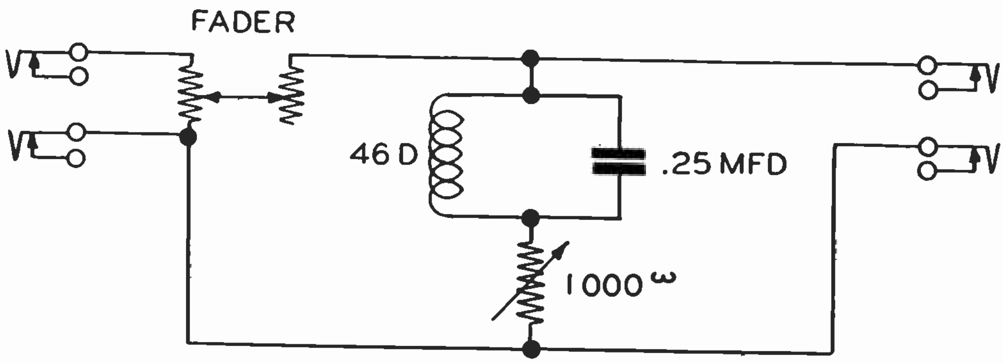

Figure 6—Circuit of audio perspective filter.

Employing an amplifier capable of delivering 100 watts at 110 volts, a turntable can be driven from this variable frequency oscillator. The desired speed (or frequency) is determined during rehearsal and must be set up by the transcription turntable operator at the proper time in the broadcast program.

The output is also utilized in the film studio, where additional outlets from the amplifiers are located. This permits playing recordings made to match silent films. Following a rehearsal, a high degree of synchronization over a fifteen minute period may be achieved.

Communication and Cue Systems

Several channels are used for communications and cues between the control room and studio.

(1) The studio microphone circuit used for the program sound is also utilized during rehearsals for conveying information to the Program Director in the control room.

(2) A circuit from the Program Director to a loudspeaker in the studio conveys instructions from the Director to the cast. During rehearsals the control room loudspeaker is cut off to prevent feedbacks when this circuit is being utilized.

A branch of the Program Director's circuit sends the directions to the Stage Manager on the studio floor.

The present circuit from the Program Director to the Stage Manager was developed to meet the demands of the greatly expanded television service from the one live-talent studio. Formerly the cueing

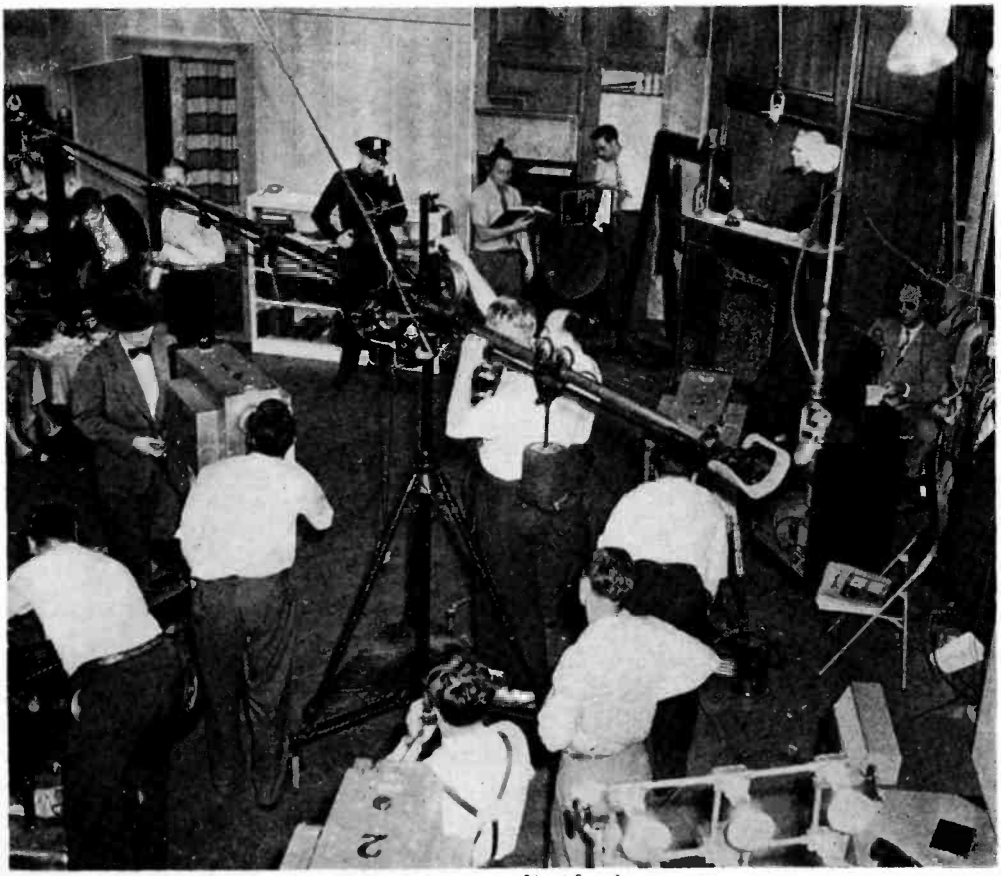

Figure 7—Television studio during a program.

system used by the Director in the control room to cue his Program Assistant on the studio floor was the usual type consisting of a microphone on the Director's desk, an amplifier, a long cable and heavy headset. The movement of the Program Assistant on the studio floor was greatly restricted by the long connecting cable. Often times the cable would become entangled with equipment or personnel and greatly hinder the rapid movement of cameras and lighting fixtures during operations. Figure 7 illustrates the amount of congestion encountered during the production of the more pretentious television shows. A special low-power transmitter[4] and a novel radio cue receiver[5] was developed which has proven to be highly satisfactory. In this system the output of a microphone on the Director's console modulates a radio transmitter mounted on the studio ceiling. This in turn is picked up by a small receiver carried by the Program Assistant. Incorporated within the receiver case is an electro-acoustic transducer from which sound is conveyed through a small flexible plastic tube to the ear of the Program Assistant. The antenna used by the receiver is a piece of small diameter wire located within the sound conducting plastic tube. Due to the relatively small diameter of the wire (in comparison

Figure 8—Stage Manager with the "Pocket Ear."

to the inside diameter of the plastic tube) no trouble is experienced with the transmission of sound to the earplug. This plastic tube and earplug, without the antenna, are similar to the ones previously described as now used by the Camera Engineers. Figure 8 shows the latest model of the radio cue receiver used in the live talent studio.

There is a cue system from the Technical Director to the Camera Operators. This is a two-way circuit for the number one Camera Operator. By means of this circuit it is possible for the Technical Director

[4] J. L. Hathaway and R. Kennedy, "The Radio Mike," *RCA REVIEW*, Vol. VIII, No. 2, pp. 251-258, June, 1947.
[5] J. L. Hathaway and W. Hotine, "The Pocket Ear," *RCA REVIEW*, Vol. VIII, No. 1, pp. 139-146, March, 1947.

to obtain information from the studio floor without utilizing the sound channel.

In this system, the output of a microphone on the Technical Director's desk is amplified and fed via the camera cable to a pair of telephone jacks on the rear of the television camera. The usual heavy headset with its cumbersome cord which formerly plugged into these jacks has been replaced by a more modern innovation. This consists of a combination plug and electro-acoustic transducer of very small size and a small flexible plastic tube on the end of which is located a tiny rubber ear plug. With the combination plug inserted into the jacks of the television cameras, the sound output of the transducer is

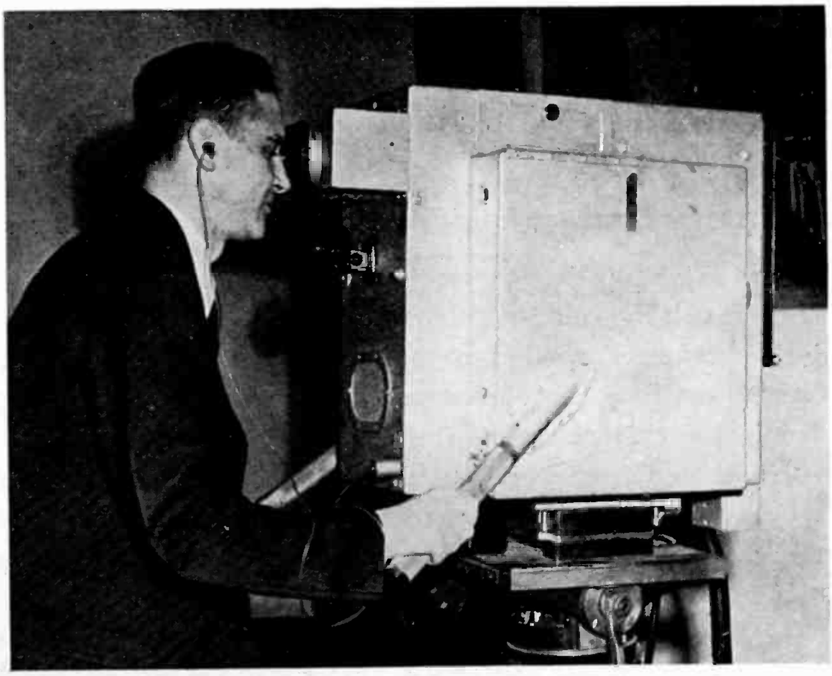

Figure 9—Camera Operators ear phone.

conveyed through the plastic tube to the ear of the Camera Engineer without being audible to others in the immediate vicinity. These small rubber earplugs are quickly replaceable and (for sanitary reasons) each person operating the cameras carries one or more of the ear plugs for his own use. Figure 9 shows the new type of earphone with its almost weightless sound conducting tube and earplug, which have served to reduce to a great extent the ear fatigue commonly encountered during television operations.

A fourth communication circuit connects the Video Control Engineer with the operator on the light bridge. Directions for adjustments and placements of lighting units are transmitted over this circuit.

In the design of future television studios it is expected that a radio cueing system, similar to the foregoing, will be used between the Video Control Engineer and the Lighting Engineer on the studio floor. The Lighting Engineer would then be better able to pace the show as it progresses and receive cues relative to lighting whenever necessary.

A fifth cue circuit permits the Audio Engineer to inform the Microphone Boom Operator when the microphone is in the picture and to pass instructions to the Boom Operator.

In order to correlate the studio programming with film and field portions of the show, there is a circuit between the Technical Directors of the live talent and film studios and the master control room.

Control Room

To permit a view of the action taking place on the studio floor, the control room is necessarily located one floor level above the studio floor and provided with a large glass window. The window was originally covered with a green cellulose material to reduce the transmission of light from the studio into the control room. When white kinescopes were installed in the video monitors, it was found desirable to replace the original material with a neutral density filter having a transmission factor of about four per cent. This matched the brightness of the view into the studio and the average brightness of the kinescopes in the control room.

It will be noted in Figure 10 that only three operating positions were used in the original control room. At that time it was believed that three men were adequate to handle the control room function, i.e., Director, Video Engineer, and Audio Engineer. As the number and complexities of the television programs increased, it became necessary to redesign and rearrange the control room equipment, redistributing the functions of the technical personnel for more efficient operation. Figures 11 and 12 show the present arrangement of the equipment and operating personnel utilizing the services of five persons, i.e., Program Director, Technical Director, Video Engineer, and two Audio Engineers (one of whom plays recordings).

The number of picture monitors has been increased from two to three. This allows one monitor to be used for "on-the-air", one for previewing forthcoming shots, and one for the exclusive use of the Video Engineer. Two cathode ray oscilloscopes have been provided for the Video Engineer so that horizontal and vertical video components may be observed without switching the sweep circuit. They are separated from the picture monitors and located immediately in front of the Video Engineer. Both oscilloscopes are switched in unison with

Figure 10—Control room—1936.

the Video Engineer's picture monitor. The Video Engineer by merely pushing the appropriate button can view the output of any camera simultaneously on both his picture monitor and oscilloscopes. This

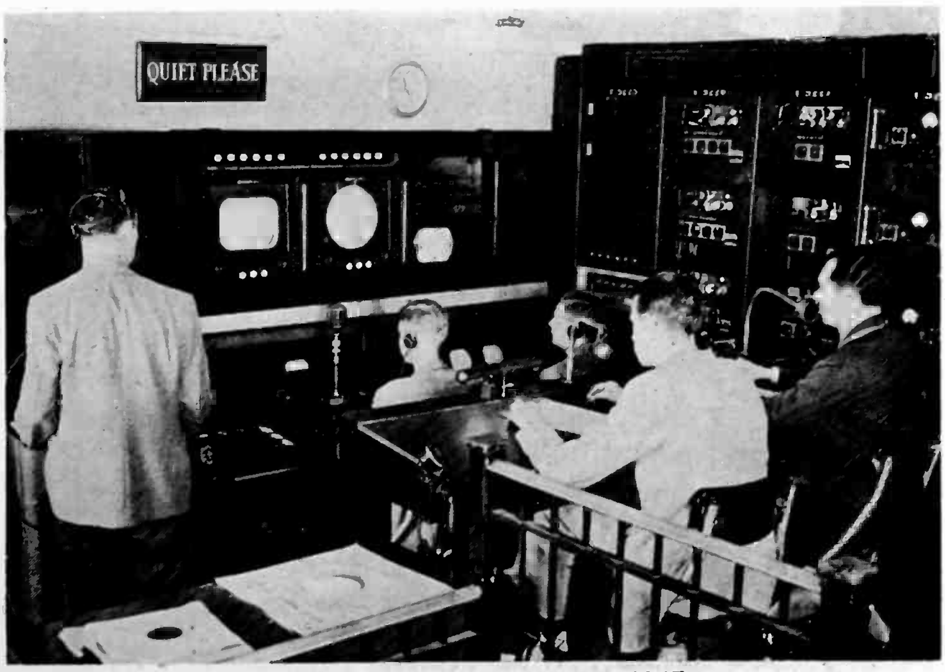

Figure 11—Control room—1947.

enables him to follow action quickly and maintain the correct picture content of the television cameras. He may also independently preview and adjust the controls for any camera.

The Program Director's console has been purposely simplified to allow him to concentrate on the production of the show and not be bothered by technical details. During both rehearsal and on-the-air production his microphone is constantly open to talk to his Program Assistant on the studio floor (via the radio transmitter and receiver previously described). To talk to the entire studio personnal over the studio loudspeaker system during rehearsals, it is only necessary for

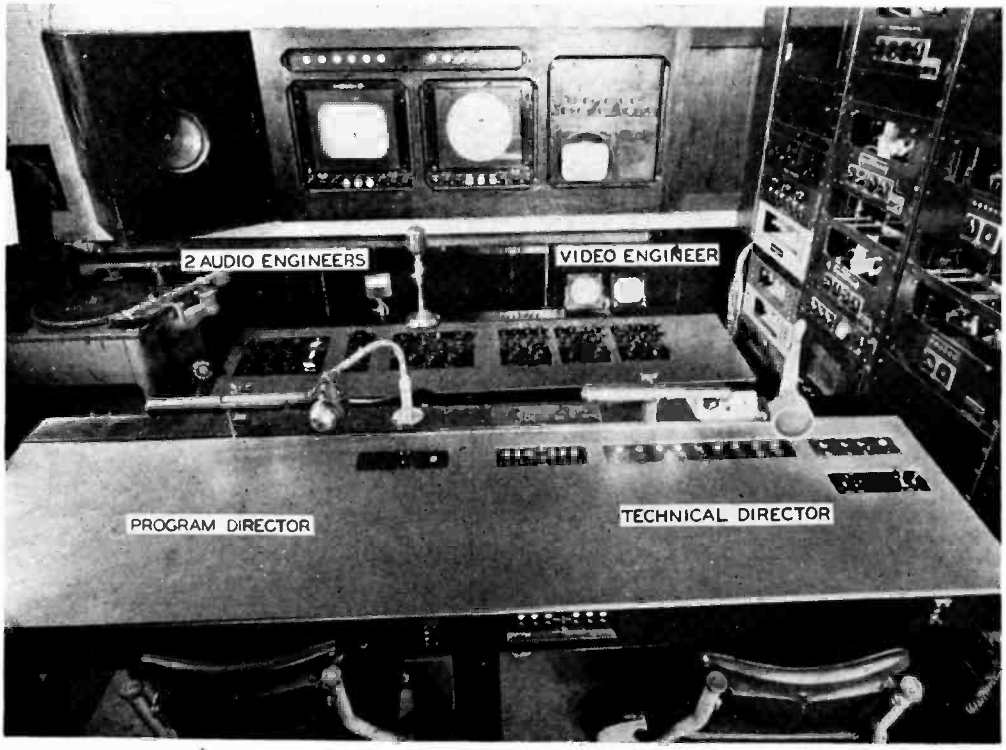

Figure 12—Control room consoles.

the Director to switch one key. However, when the studio is on-the-air, the studio loudspeaker system is automatically made non-operable. Thus, during the actual production of a television show, the director can give the necessary directions without the necessity of wearing a headset or operating a control.

FREQUENCY ALLOCATIONS*

By

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Summary—The subject of frequency allocations is discussed in general terms pointing out the various reasons why international agreement has become increasingly necessary. The background and scope of such agreement is then briefly reviewed, with particular reference to the recent International Radio Conference at Atlantic City, New Jersey. Emphasis is given to industry cooperation at this conference with reasons therefor. The RCA Radio Frequency Allocations Charts are described, from their beginning in 1941 to the post-conference edition now being distributed.

UNLESS radio stations are separated from each other by suitable spacing of their geographical locations and the frequencies on which they operate, interference which is mutually destructive of their services will result. Standardized practices with regard to the use of frequencies among stations of the same service, which are required to communicate with each other, are necessary in order that the communication or connection may be established and thereafter conducted efficiently. Agreement and regulation are the foundations upon which the entire radio structure is based. The importance of the allocation of frequencies and establishment of the conditions of their use is therefore measured by the total investment in radio and the value of the lives, property, interests and ideas which depend for advancement and protection upon the radio medium.

The science of frequency allocations, which at one time was very simple, has become exceedingly complex within the past few years. The first International Radio Conference in Berlin, 1906, dealt with only two frequencies, namely 500 and 1000 kilocycles and was concerned only with telegraphy between ship and shore. The London Conference of 1912 again dealt with only a small part of the radio frequency spectrum, between 150 and 1000 kilocycles, and was generally limited to ship-shore communications. The development of high frequency communications, hastened by World War I, expanded the use of the spectrum so materially that the Washington Conference of 1927 considered service allocations from 10 to 23,000 kilocycles. By the time of the Madrid Conference, 1932, this had been increased to 30,000 kilocycles,

* Decimal Classification: R007.1 × R084.

while the Cairo Conference of 1938 extended the allocations table to 200,000 kilocycles, the agreements on the higher bands being on a regional basis. At the Atlantic City Conference this year, due to the knowledge of the super-high and ultra-high frequency spectrum gained during World War II, the allocations table was further extended to 10,500,000 kilocycles.

As the spectrum has become more and more congested, every nation has wished to obtain for itself the possibility of acquiring maximum frequency rights in those parts of the spectrum where its main interests could best be accommodated. At Atlantic City, the maritime countries, the United States, United Kingdom, and the Scandinavian countries, sought the use of large portions of the spectrum for the maritime services. Countries lacking internal wire communications such as the Soviet Union, Brazil and China, desired many fixed point to point frequencies. Most of the smaller nations not interested particularly in the fixed or maritime services desired to use the major portion of the high frequency spectrum for broadcasting. This was particularly true in tropical areas where, due to high noise levels, the standard broadcast frequencies in the medium band do not give satisfactory radio reception.

Also, the Western Hemisphere, headed by the United States, has wanted to obtain the maximum possible allocation for the amateur service while others, particularly the European nations, which do not have the large number of amateurs present in this hemisphere, desired a minimum allocation to amateurs in order to provide more space in the spectrum for the broadcasting and fixed services.

The adjustment of these different viewpoints internationally was therefore a major undertaking and one which required concessions by all nations, with no one nation's plan completely fulfilled in the final result. A further consequence of these differences is that, wherever world-wide agreement is not essential, either because of propagational characteristics or because of the necessity for standardization of world-wide mobile communications, the allocations have now become largely regional in nature.

In discussing the results of the Atlantic City allocations, it is helpful to divide the spectrum into three portions: (1) that below 2850 kilocycles, (2) that between 2850 kilocycles and 30 megacycles, (3) that between 30 megacycles and 10,500 megacycles.

In the first portion, that below 2850 kilocycles, the most important development was the assignment of the band 90-110 kilocycles as the position in the spectrum for the ultimate long-distance navigational aid. Preliminary work on low frequency Loran has been started in the United States between 200 and 280 kilocycles but it was found impossible to get world-wide agreement on any small segment of that band. Accordingly, a small band centering on 100 kilocycles was selected as the best technical compromise from the standpoint of developing electronic aids for this purpose. Great opposition developed to the continuation of the present standard medium frequency Loran in Europe, particularly by the Scandinavian countries. It was argued by the United States that there should be no hiatus pending the development of low frequency Loran whereby those ships and aircraft depending upon electronic aids to navigation in trans-Atlantic voyages would have no aid at all. A compromise was effected therefore, whereby medium frequency Loran was accorded temporary recognition in the band 1900-2000 kilocycles until July 1, 1949, and provision made for a special administrative conference to be called if necessary to review this date of termination of Loran in the Eastern part of the North Atlantic. The use of Loran in the band 1800-2000 Kc in regions other than Europe was accepted on a permanent basis. It is only a question of time, however, before medium frequency Loran will be replaced.

Beyond the allocation for electronic aids, the points of major importance in the band 10-2850 kilocycles decided by the Conference were as follows:

1. The marine beacon band was widened by approximately 33⅓ per cent and was established on a world-wide basis.
2. The medium frequency broadcasting band was widened to 535-1605 kilocycles, thereby making the frequencies 540 and 1600 kilocycles available for U. S. broadcast stations after appropriate arrangements are concluded at the next NARBA Conference.
3. The frequency 2182 kilocycles was designated on a world wide basis for distress and calling purposes in the maritime mobile radiotelephone service.
4. Provision was made for both ship and coast station telegraphy in the 2-megacycle band in the Western Hemisphere. This action is regarded as of special future significance because it is doubtful that the maritime mobile service can continue to obtain adequate allocations in the vicinity of 500 kilocycles at future conferences.

In the high frequency portion of the spectrum, i.e., approximately from 2850 kilocycles to 30 megacycles, the aeronautical mobile service for the first time received exclusive bands and made substantial gains. The broadcasting bands were widened on an average of over 25 per cent.

The maritime mobile service maintained about the same spectrum space as had been previously allocated, while the fixed service lost practically all the space gained by the aeronautical and broadcasting service. Moreover, this loss does not represent by any means the actual loss to the fixed service. The reduction in spectrum space was most severe in those bands where the situation was already critical, namely in that portion of the frequency spectrum between 4 and 10 megacycles and further arrangements were made to share a considerable portion of the space allocated in the 4 megacycle band to the fixed service with tropical broadcasting. These losses to the fixed service must be made up in order to maintain traffic capacities. One way of accomplishing this is by devising more efficient means of frequency utilization such as multiplexing and intensive study and application of radio propagation data. Among other aids in solving the problem, an International Frequency Registration Board has been created which will endeavor to arrange on a sound engineering basis for maximum use of the frequency spectrum, taking advantage of the various propagation factors such as the cycle of sunspot activity.

The amateur service lost the band 1800 to 2000 kilocycles as was expected, but also lost 50 kilocycles in the 14 megacycle band. This was as a result of a compromise between the Western Hemisphere, holding out for an amateur allocation of 400 kilocycles in this band, and the European countries which advocated a band of 300 kilocycles only. However, this loss was compensated for by a new band 21,000 to 21,450 kilocycles which was assigned exclusively for amateurs on a world wide basis.

Above 30 megacycles and below 10,500 megacycles, the United States secured complete adoption of its proposals either on a world wide or regional basis. This includes also a major modification made in the United States proposal, steps toward which were taken just prior to the beginning of the Conference in the frequency range 1215 to 1660 megacycles in order to provide enlarged allocations for the aeronautical radio navigation service. This particular proposal was adopted for the Western Hemisphere and certain other areas as well. In addition, those countries not adopting the allocation at this time were urged to envisage the possibility of the future application of an integrated system of electronic aids to air navigation and traffic control in this frequency band of world-wide application.

All companies in the radio and electronics fields are, of course, deeply interested in all these developments regarding changes in allocations to the various services. The value of frequencies can be assessed only in terms of total investment; the continuation and growth of all

MARITIME MOBILE (MM)

SHIP STATIONS
1. Letter P indicates telephone, otherwise telegraph.
2. For telegraph, letters FAS indicate passenger ships and CAR indicate cargo ships.

COASTAL STATIONS

MARITIME NAVIGATION AIDS (MNA)

RADAR

GENERAL NAVIGATION AIDS (NA)

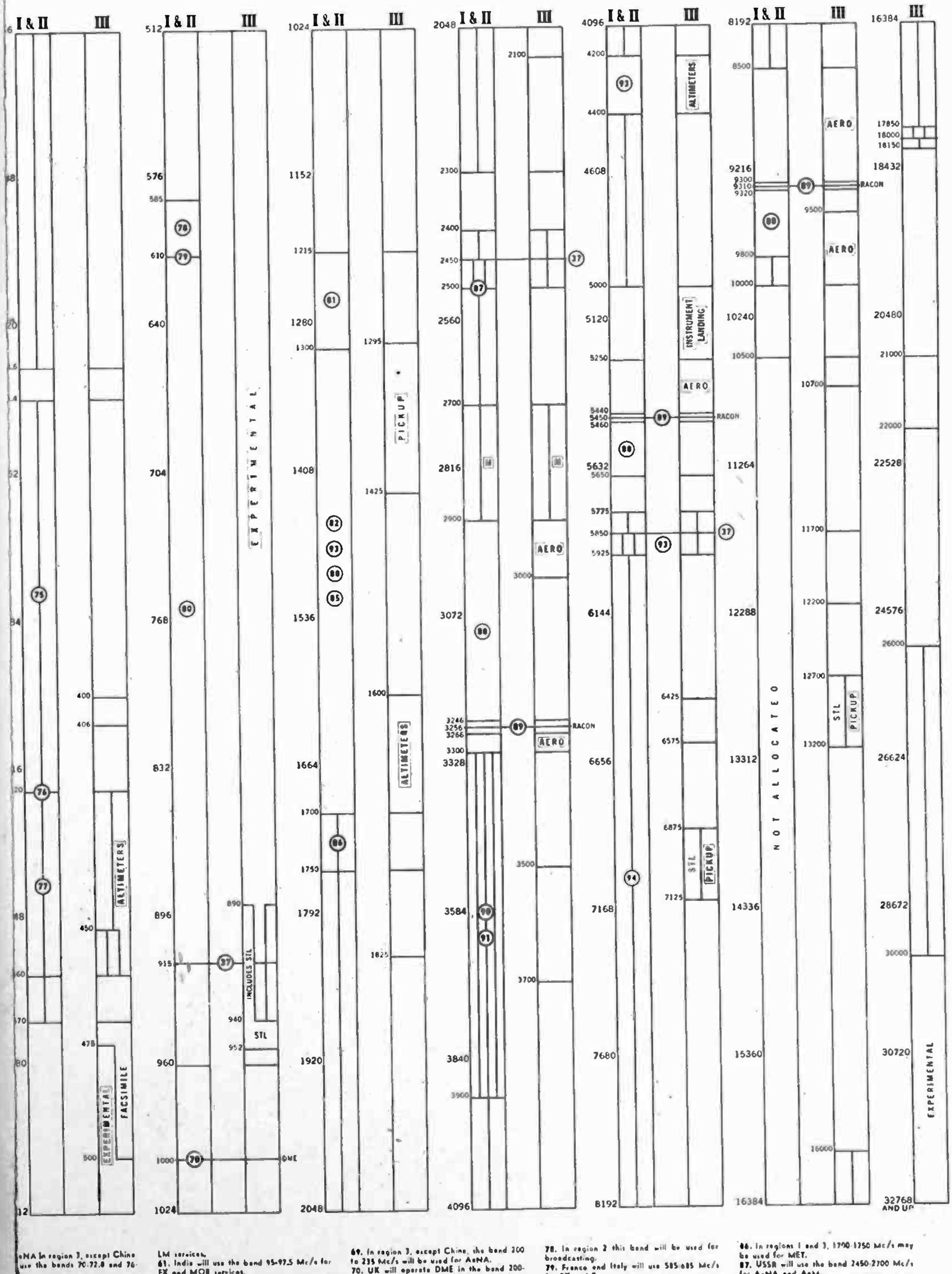
AERONAUTICAL MOBILE (AeM)
1. Letter R indicates "Route."
2. Letters OR indicate "Off-Route."

AIR NAVIGATION AIDS (AeNA)

RADIO SOUNDING
Letter M indicates meteorological aids (MET)

AIRPORT CONTROL

INDUSTRIAL SCIENTIFIC AND MEDICAL DEVICES (ISM)



...eNA in region 3, except China use the bands 70-72.8 and 76-...

...band on 75 Mc/s is ± 0.2 in ± 0.4 Mc/s in regions 2 and 3. 78-80 Mc/s will be used ex-...eNA in region 3, except China

...ing is excluded from the band ...n region 3, except China and ...Note 51).

...and New Zealand will use 85-90 ...A.

... use 76-108 Mc/s for broad-...

...se 85-90 Mc/s also for MNA

... use 88-108 Mc/s for B and FX

...UK and India will use 94.5-95 ...

... 95-100 Mc/s also for FX and LM services.

61. India will use the band 95-97.5 Mc/s for FX and MOB services.

62. South Africa will use the bands 100-108 Mc/s for B and 132-144 and 146-174 Mc/s for FX and MOB services.

63. 100-108 Mc/s will be used in Australia for AeM (OR) until required for B and in New Zealand for B and MOB services.

64. Australia and New Zealand will use the bands 132-144 and 148-156 Mc/s exclusively for AeM (OR) service.

65. In region 1, MET is admitted in the band 151-154 Mc/s.

66. France will use the band 162-174 Mc/s for broadcasting.

67. Broadcasting is admitted in 170-174 Mc/s in region 3, except Australia where 170-178 Mc/s is AeNA service.

68. UK will share 174-200 Mc/s with FX and use the band 200-216 Mc/s for AeNA services.

69. In region 3, except China, the band 200 to 235 Mc/s will be used for AeNA.

70. UK will operate DME in the band 200-235 Mc/s until world standardization at 1000 Mc/s has been accomplished.

71. South Africa will use 174-216 Mc/s also for FX and LM.

72. South Africa will use 220-225 Mc/s for amateur service.

73. USSR will use the band 216-260 Mc/s for NA.

74. Region 2 will operate DME in the band 220-221 Mc/s until January 1, 1952 in accordance with special arrangements.

75. MET may be operated in the band 400-420 Mc/s.

76. USSR will use the band 412-460 Mc/s for NA.

77. In the band 420-460 Mc/s the AeNA service has priority. Region 2 temporarily uses the band exclusively for altimeters and will later share AeNA, FX and MOB.

78. In region 2 this band will be used for broadcasting.

79. France and Italy will use 585-685 Mc/s for FX and B.

80. In region 2 the band 890-940 Mc/s will be used for FX and B and the band 940-960 Mc/s for FX services.

81. USSR will use 1215-1300 Mc/s for FX, primarily TV relays.

82. In region 1, except the USSR, the band 1300-1600 Mc/s will be used for FX and MOB services.

83. In region 2 the band 1300-1660 Mc/s is intended for development of an integrated world-wide system of electronic aids to air navigation and traffic control. The band 1660-1700 Mc/s will be used for MET in this region.

84. In region 2 and the UK the band 1300-1365 Mc/s is restricted to surveillance radar.

85. In region 3, the band 1300-1700 Mc/s will be shared between AeNA, FX and MOB services.

86. In regions 1 and 3, 1700-1750 Mc/s may be used for MET.

87. USSR will use the band 2450-2700 Mc/s for AeNA and AeM.

88. In the bands 2900-3300, 5250-5650 and 8500-9800 Mc/s shipborne radar in merchant ships are confined within 3000-3246, 5460-5650 and 9320-9500 Mc/s respectively.

89. The bands 3246-3266, 5440-5460 and 9300-9320 Mc/s are designated for Racons.

90. In region 1 amateurs are excluded from 3300-3900 Mc/s.

91. In region 2 the bands 3300-3500 Mc/s will be used for AM and 3500-3900 Mc/s for FX and MOB services.

92. China may use the band 4200-4400 Mc/s for FX on a non-interference basis.

93. In region 2 the band 5850-5925 Mc/s will be used for amateur service under the conditions of note 37.

94. USSR may use the band 6900-7050 Mc/s for MET service.

Prepared by the RCA Frequency Bureau

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All companies in the radio and electronics fields are, of course, deeply interested in all these developments regarding changes in allocations to the various services. The value of frequencies can be assessed only in terms of total investment; the continuation and growth of all

COLUMN I — REGION 1 ALLOCATION — EUROPE INCLUDING ALL USSR TERRITORY OUTSIDE EUROPE, MONGOLIAN PEOPLES' REPUBLIC, ASIA MINOR AND AFRICA.

COLUMN II — REGION 2 ALLOCATION — WESTERN HEMISPHERE INCLUDING HAWAII AND REGION 3 ALLOCATION — ASIA EXCLUDING USSR TERRITORY AND ASIA MINOR, AUSTRALIA, NEW ZEALAND AND OCEANIA.

COLUMN III — FEDERAL COMMUNICATIONS COMMISSION (UNITED STATES) ALLOCATION.

INDICATED FREQUENCIES ARE IN KILOCYCLES UP TO 25015. THEREAFTER IN MEGACYCLES.

GENERAL FIXED (FX)

LAND MOBILE (LM)

POLICE

BROADCASTING (B) Letter T indicates Tropical Broadcast

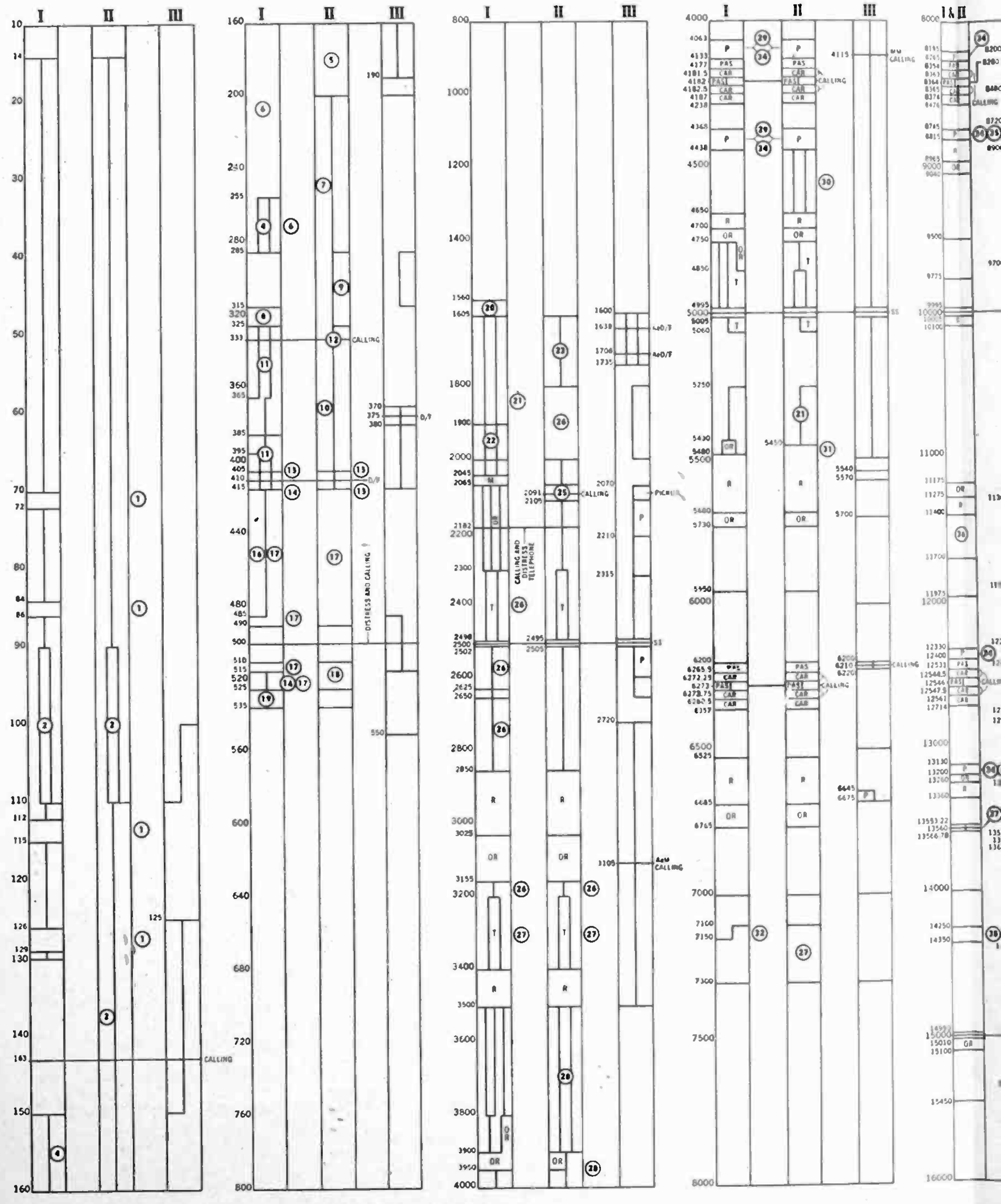
INTERNATIONAL BROADCASTING

TELEVISION BROADCASTING (TV)

FM BROADCAS[...] Facsimile por[...]

EDUCATIONAL BROADCASTING

CITIZENS' OR CIVIL RADIO



ALLOCATION NOTES

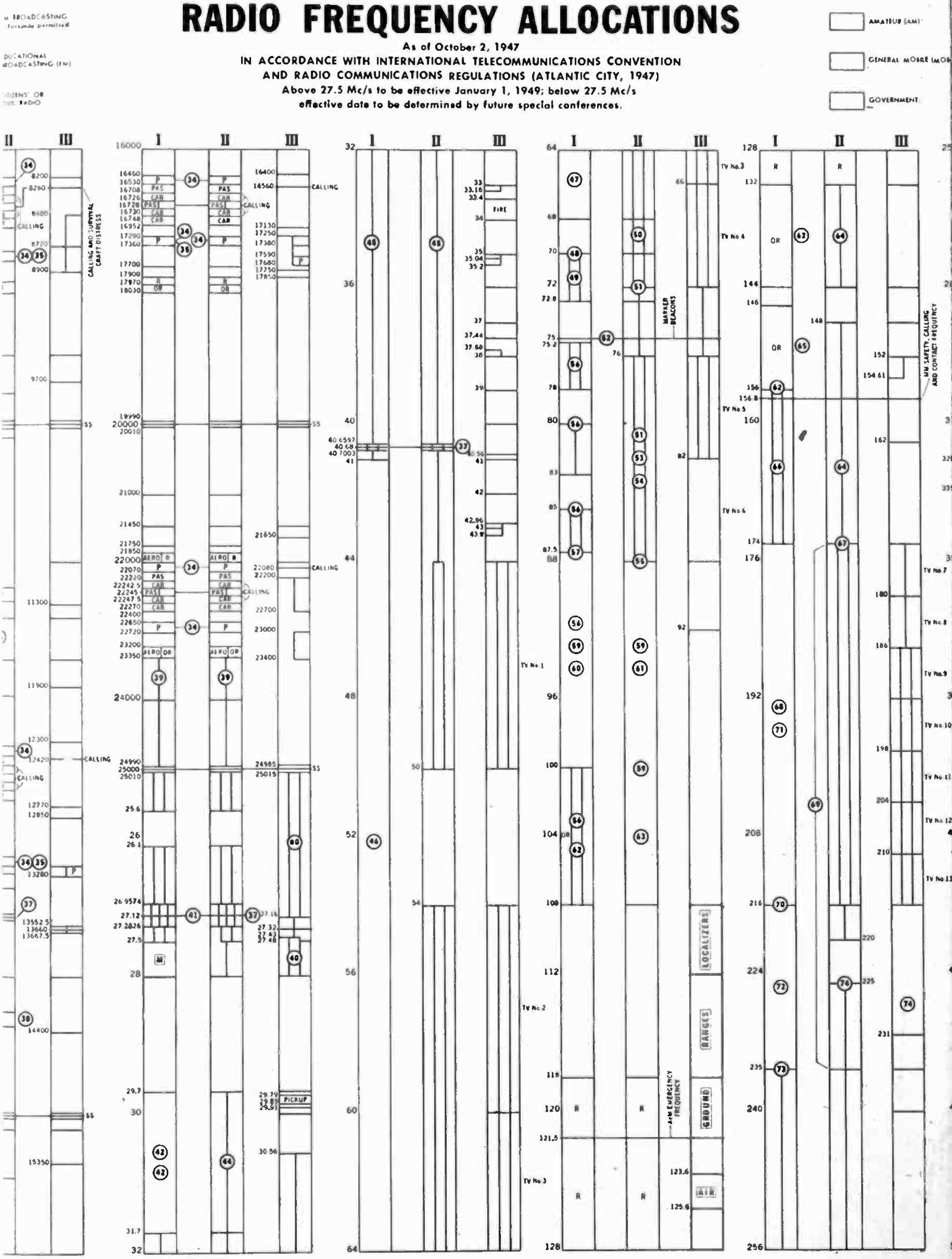
1. In region 1, Australia and New Zealand, the bands 70-72, 84-86, 112-115 and 126-129 kc/s are reserved for cw NA.
2. Development of long distance NA authorized in band 90-110 kc/s to be exclusively allocated whenever such aid is adopted internationally. If pulse aid is employed, emissions must be confined within the band, and must cause no harmful interference outside the band to stations operating pursuant to the regulations.
3. In areas outside region 1, FX may operate provided that it does not cause harmful interference to ship telegraphy in the North Atlantic and Mediterranean.
4. MM service must not cause harmful interference to the broadcasting service within the boundaries of the national territories in which the broadcasting stations are situated.
5. Priority will be given the aeronautical fixed service in northern areas which are subject to auroral disturbances.
6. South Africa will use the band 160-200 kc/s for FX and 200-285 kc/s for AeNA and AeM services.
7. AeNA has priority in region 2, China, India and Pakistan.
8. USSR will use the band 315-325 kc/s for MNA by special arrangement.
9. MNA has priority in the band 285-325 kc/s.
10. AeNA has priority except in New Zealand.
11. Primarily aeronautical; other services must not interfere.
12. 333 kc/s is aircraft calling frequency in regions 1 and 3.
13. No harmful interference to be caused to direction-finding (D/F) on 410 kc/s.
14. USSR may use this band for AeNA on the basis of no harmful interference to existing MNA services.
15. AeNA has priority over AeM in region 2.
16. MM service has priority.
17. Limited to telegraphy.
18. MM service has priority in region 3.
19. South Africa will use this band for mobile service.
20. The band 1560-1605 kc/s is shared in the USSR with fixed service subject to mutual non-interference with B in Europe.
21. In the band 1715-2000 kc/s Austria, UK, Ireland, The Netherlands, Switzerland and South Africa may assign up to 200 kc/s for the amateur service.
22. Operation of loran chain (Iceland-Faroes-Hebrides) is authorized in band 1900-2000 kc/s until July 1st, 1949 provided all practicable means are taken to minimize harmful interference to other services in the same or adjacent frequency bands. Operation beyond that date is contingent upon action to be taken as prescribed in the radio regulations.
23. AeNA admitted in this band in region 2.
24. Loran systems are operated in this band. AM, FX or MOB services except aeronautical may be authorized on condition of no harmful interference to NA service.
25. In region 2 the band 2088.5-2093.5 kc/s is used for calling.
26. AeM (R) service not permitted in this band.
27. AeM (R) service not permitted in the band 3200-3230 kc/s and no AeM in the band 3230-3400 kc/s.
28. In region 2 the band 3500-4000 kc/s will be shared between AM, FX and MOB, except AeM (R).
29. In the bands 4063-4133 and 4408-4418 kc/s, FX service not exceeding 5 kw peak power may be operated, under conditions that minimize the possibilities of harmful interference to MM service, if situated at least 600 kilometers from coast.
30. AeM (OR) admitted in the band 4438-4650 kc/s in region 2.
31. In region 1 the allocation of the band 5250-5480 kc/s is the same as represented in column I.
32. Broadcasting has priority in this band. However, in South Africa the band will be used for amateur service.
33. In region 3 this band will be used as represented in col. 1, except in China and New Zealand the entire band and in Australia and Netherlands East Indies the band 7100-7150 kc/s may be used for amateur service.
34. For telegraph use of the MM telephony bands see Chapter III, art. 9, Section IV, paragraph 11 of the radio regulations.
35. In the USSR the bands 8615-8815, 12925-13200 and 17160-17360 kc/s will be used for

RADIO FREQUENCY ALLOCATIONS

As of October 2, 1947

IN ACCORDANCE WITH INTERNATIONAL TELECOMMUNICATIONS CONVENTION AND RADIO COMMUNICATIONS REGULATIONS (ATLANTIC CITY, 1947)

Above 27.5 Mc/s to be effective January 1, 1949; below 27.5 Mc/s effective date to be determined by future special conferences.



FX service on a shared basis with MM service.

36. In the USSR the band 11400-11450 kc/s will be used for AeM (OR) service shared with FX service.

37. The frequencies expressed in Mc/s. 13.56, 27.12, 40.68, 915 ± 25 Mc/s in region 2 only, 2450 ± 50 Mc/s and 5850 ± 75 Mc/s in region 2, Australia, New Zealand, South Africa and the UK only; are designated for industrial, scientific, and medical (ISM) purposes. Emissions confined within the limits of the bands indicated, enjoy priority over other services operated in these bands.

38. The USSR uses the band 14250-14350 kc/s also for FX.

39. MM inter-ship telegraphy admitted in the band 23350-24000 kc/s.

40. Remote pickup broadcast and aeronautical flight test and schools accommodated.

41. In region 2, Australia, New Zealand and South Africa, the band 26.96-27.23 Mc/s will be used for AM service.

42. The USSR will use the band 29.7-30 Mc/s additionally for AeM shared with AeNA service.

43. The USSR will use the band 30-31.7 Mc/s for NA.

44. Australia will use the band 29.7-31.7 Mc/s for AeNA.

45. In region I the AeNA service may be accommodated in the band 31.7-41 Mc/s, and in region 3 in the band 31.7-44 Mc/s.

46. South Africa will use 41-44 Mc/s for AeNA, FX and MOB; 44-50 and 54-68 for FX, MOB and B, and 50-54 Mc/s for amateurs.

47. United Kingdom (UK) will use the band 66.5-68 Mc/s for FX and LM under local agreement with France.

48. The USSR will use the band 68-72 Mc/s for broadcasting.

49. USSR and France will use band 72-72.8 Mc/s for amateurs.

50. The band 68-70 Mc/s will be used ex-

clusively for A

51. India will

85 Mc/s for

52. The guard

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54. Broadcast

80-87 Mc/s

India (see als

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Mc/s for M

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58. China wil

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59. France,

Mc/s for ME

60. UK will

RADIO CORPORATION OF AMERICA

RCA LABORATORIES DIVISION • RCA VICTOR DIVISION

RCA INTERNATIONAL DIVISION

RCA COMMUNICATIONS, INC. • RADIOMARINE CORPORATION OF AMERICA

NATIONAL BROADCASTING COMPANY

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the components of the radio and electronic industry depends on frequency availability. Frequencies are the life's blood of radio operating and broadcasting companies who must be assured that the proper frequencies to carry on their essential communication and broadcasting work are provided. Manufacturing organizations also have to follow these changes carefully in order that they can supply to the operating organizations, in this country and in foreign countries, equipment which will meet and take maximum advantage of the provisions of the Radio Regulations and the allocations table. Experimental and research work must make certain that future developments will lead to the provision of equipment in the proper frequency bands. The Radio Corporation of America is concerned with all these phases and was represented, together with other companies in the radio and electronics fields, by a group of experts at Atlantic City, as it has been at previous conferences, to the end that these developments and changes may be guided so far as possible in the interests of the nation and its radio industry, and so that the policies of all operating and manufacturing companies may be geared to the future course of radio as charted in the international allocations.

RADIO FREQUENCY ALLOCATIONS CHART

For its own use in both the manufacturing and the operating branches, the Radio Corporation of America has felt that a large colored chart suitable for framing was essential for ready reference. Realizing that such a chart would fill a similar need for government officials and others interested in radio communications, it published its initial coordinated color-coded Radio Frequency Allocations Chart and distributed it widely to domestic and international radio communication and manufacturing interests in 1941. This chart represented the frequency allocation plan adopted at Cairo in 1938. It was arranged in double adjacent columns, one representing the general international plan and the other the European regional plan. Departures from the basic allocations plans were explained by numbered footnotes referenced to the color coded portion of the chart. The vertical columns were arranged so that the exact numerical frequency values noted in the margin beside the columns at 2000 cycles and above, were in harmonic relation horizontally across the chart. This general scheme has been followed in subsequent charts.

The second edition of the Radio Frequency Allocations Chart was prepared and distributed on July 1, 1945. This was a three column vertically grouped arrangement covering the radio spectrum from 10 to 32,000 kilocycles with two columns vertically grouped from 32,000 to 256,000 kilocycles and a single vertical column above 256,000 kilocycles, in order to reflect the Cairo allocations, world-wide and regional, together with the U. S. plans extending beyond the Cairo allocations. Since many of the users of the chart had already had the old chart framed, it was felt desirable to maintain the size but change the spacing between the columns and scale somewhat in order to include all of the additional information and still make it possible to put the new chart in the frames in use.

As a result of the many changes made in frequency allocations at Atlantic City, a new Radio Frequency Allocations Chart has been prepared and distributed.† This chart is illustrated, solely for recognition purposes, in black and white, reduced size, opposite page 740. The actual chart, however, is a complete seven-color chart 40 inches by 25 inches on heavy paper suitable for framing. While this chart is similar in size and in the general method of presentation to the previous RCA charts, it is, of course, considerably more complex due to the present complicated structure of the international and domestic frequency allocations, as well as to the tendency for more and more regional allocations as previously mentioned.

In order to make this new chart of maximum value, an effort has been made to have it show the allocations for every important country in the world; therefore, in general, a three column arrangement has been adopted: the first column for the European region which also includes the territory of the USSR outside of Europe, the Outer Mongolian Peoples Republic, Asia Minor and Africa; the second column for allocations for the Western Hemisphere, Southern Asia, Australia and Oceania; and the third column for the Federal Communications Commission allocation for the United States. As before, the size of the chart has been maintained the same as that of previous charts so that it may be placed in the frames now in use but it has again been found necessary to change somewhat the spacing between the columns as well as the scale in order to include all the additional information desired. Many more notes are included than heretofore in order to clarify all exceptions to the main chart but as each note is referred to by a figure in the chart itself, it is felt that it is not too complicated for ready reference. It is believed that this chart, issued soon after many radical changes in the allocations table had been decided upon at the Atlantic City Conference, will be of great interest and value to many radio and electronic engineers.

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† The RCA REVIEW Department of RCA Laboratories Division has available a very limited number of these charts. Anyone whose work involves frequency allocations and who feels that this chart would serve as a useful reference source in this work may obtain a copy, as long as the supply lasts, by writing to RCA REVIEW, RCA Laboratories Division, Princeton, N. J.

AN ULTRA-HIGH-FREQUENCY LOW-PASS FILTER OF COAXIAL CONSTRUCTION*

By

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Summary—The design of an ultra-high-frequency low-pass filter of coaxial construction is described. The design equations and design criteria are derived and are applied to two coaxial filters whose cut-off frequencies are 800 megacycles and 1800 megacycles, respectively, and whose mechanical details and transmission characteristics are discussed.

During the course of an investigation of modes of oscillation in cavity magnetron blocks by means of cold-resonance measurements, it was found that the harmonics in the output of the test oscillator excited modes belonging to multiplets other than the fundamental one. These cold-resonance tests were conducted for modes existing in the range from 600 to 1500 megacycles and troublesome oscillator harmonics were identified up to 10,000 megacycles. It was evident that a filter with low-pass characteristics could be used to eliminate this difficulty without modification of the oscillator or the other elements of the testing apparatus and a simple coaxial filter that was successfully utilized will be described.

It was required that the low-pass filter, which was to be designed for this application, have a minimum number of pass bands above the cut-off frequency. Since the phase shift and input impedance variation over the low-pass band were not of primary importance, these aspects are not considered here. To be most useful, a filter for experimental work at these frequencies should be fitted with coaxial connectors permitting rapid connection to coaxial cables, and should be completely closed to prevent radiation and coupling with other apparatus. It is always desirable that such units be of simple and rugged mechanical construction.

After consideration of various filter designs which might be satisfactory, it was decided that the most direct approach to the problem was to construct a constant-K, T-type, low-pass filter in which the lumped series inductances and shunt capacitances used in low-pass filters were to be replaced by lengths of coaxial transmission line whose distributed parameters would closely simulate the lumped parameters.

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* Decimal Classification: R310 × R386.2.

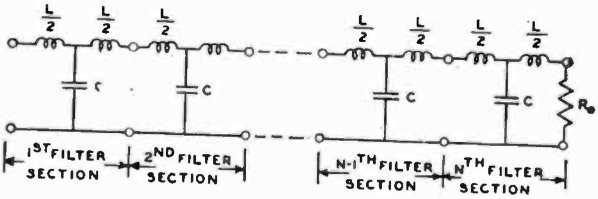

Fig. 1—Lumped parameter constant-K T-type low-pass filter with N sections.

The design of the filter was carried out in the following way: It is known[1] that for the low-pass, constant-K, T-type filter shown in Figure 1,

$$C = \frac{1}{\pi f_c R_o} \text{ farads} \quad (1) \qquad L = \frac{R_o}{\pi f_c} \text{ henries} \quad (2)$$

where f_c is the cut-off frequency in cycles per second and R_o is the terminating resistance in ohms.

In the ultra-high-frequency coaxial filter being described, a section of coaxial line having a large characteristic impedance with respect to R_o is used as the inductance element and a section of coaxial line having a small characteristic impedance with respect to R_o is used as the capacitance element. The conditions for which a short length of coaxial line will appear to be a series inductance or a shunt capacitance are derived in the Appendix and are listed there in Equations (5-a), (6-a), and (10-a).

The geometry used for the coaxial filters is shown in Figure 2. Expressions for the characteristic impedance of the inductance and capacitance sections are

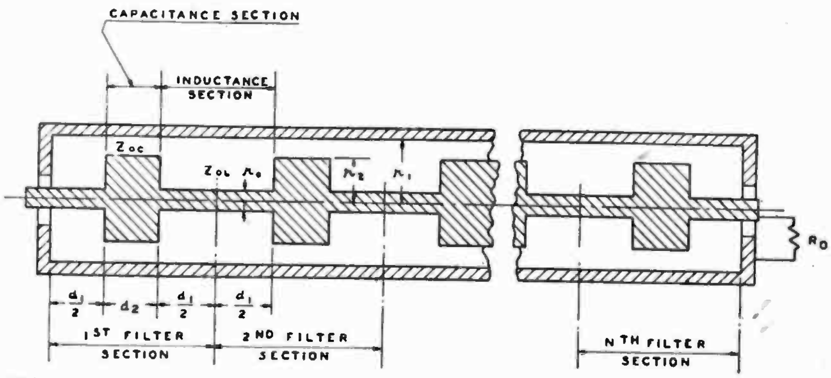

Fig. 2—Detail of UHF low-pass coaxial filter with N sections.

[1] Ware and Reed, COMMUNICATION CIRCUITS, John Wiley and Sons, New York, N. Y., 1944 (Equations 9-14, 9-15, pp. 126-127.)

$$Z_{oL} = 138 \operatorname{Log}_{10} \frac{r_1}{r_o} \text{ (ohms)} \quad (3), \qquad Z_{oc} = 138 \operatorname{Log}_{10} \frac{r_1}{r_2} \text{ (ohms)}. \quad (4)$$

The total inductance of the inductance section of length, d_1, is approximately

$$L \cong \frac{Z_{oL}\, d_1}{v} \qquad (5)$$

and, neglecting fringing, the total capacitance of the capacitance section of length, d_2, is approximately

$$C \cong \frac{d_2}{Z_{oc}\, v} \qquad (6)$$

where v is the velocity of light. Equations (5) and (6) are suitable representations for lumped inductance and capacitance if

$$\tan \frac{2\pi f\, L_o\, d_1}{Z_{oL}} \approx \frac{2\pi f\, L_o\, d_1}{Z_{oL}} \qquad (7)$$

and

$$\tan 2\pi f\, Z_{oc}\, C_o d_2 \approx 2\pi f\, Z_{oc}\, C_o d_2 \qquad (8)$$

as specified in Equation (6-a). L_o and C_o are the inductance and capacitance per unit length of line, respectively.

Substituting Equations (5) and (6) into Equations (1) and (2), the following design relationships are obtained:

$$d_1 = \frac{R_o\, \lambda_o}{138\, \pi \operatorname{Log}_{10} \frac{r_1}{r_o}} \quad (9) \qquad d_2 = \frac{138\, \lambda_c}{\pi\, R_o} \operatorname{Log}_{10} \frac{r_1}{r_2} \quad (10)$$

where d_1 and d_2 are in centimeters and λ_c is the cut-off wavelength in centimeters. It is evident that Equations (9) and (10) may be combined to yield the general relationship

$$\lambda_c = \pi \sqrt{d_1\, d_2 \frac{\operatorname{Log}_{10} \frac{r_1}{r_o}}{\operatorname{Log}_{10} \frac{r_1}{r_2}}} \qquad (11)$$

in terms of the lengths, d_1 and d_2, and the radii, r_o, r_1, and r_2 which are interrelated by Equations (9) and (10).

The design relations—Equations (9), (10), and (11)—for a practical ultra-high-frequency coaxial filter will be valid if the following criteria are observed:

$$d_2 < d_1 << \lambda_c/8 \quad (12) \qquad 2\pi r_1 < \lambda_c \quad (13)$$

$$r_1 - r_2 < d_2/4 \quad (14) \qquad r_1 - r_o < d_1/4 \quad (15)$$

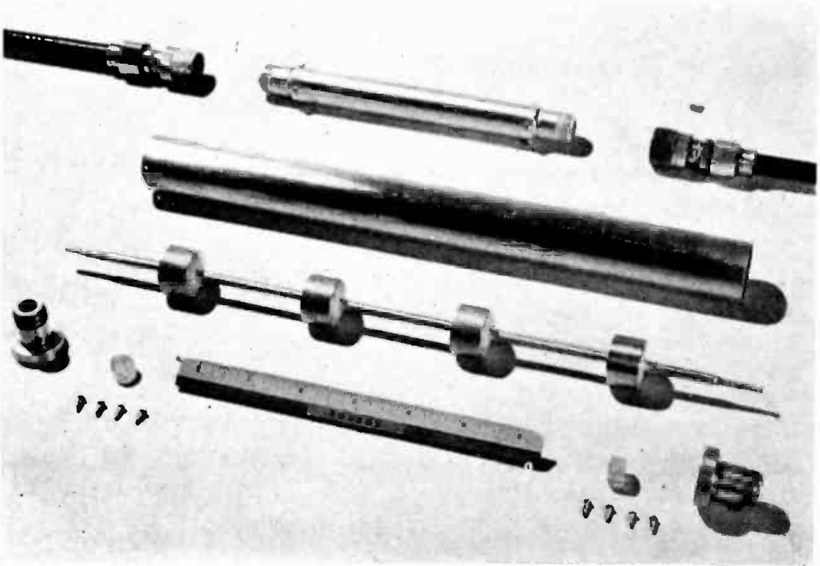

Fig. 3—Photograph of the 800-megacycle and 1800-megacycle coaxial low-pass filters. The 800-megacycle filter is shown dismantled.

Equation (12) is deduced from Equations (7) and (8). Equation (13) is specified so that the design of the filter will be made for the fundamental mode of any of the component sections of coaxial line. Equations (14) and (15) are intended to minimize end effects and represent an average of information which has been yielded by numerous field plots.

Two of the filters which were constructed were designed using Equations (9) and (10) for cut-off frequencies of 800 and 1800 megacycles and are pictured in Figure 3 with the 800-megacycle filter shown dismantled. The center structure was made by soldering discs at proper intervals on a center rod. Army-Navy coaxial cable connectors were used at both ends of the structure and served to hold and position the center conductor in the enclosing cylinder. All of the metal components were made of brass and were silver plated.

The details of these filters are tabulated as follows, with the dimensions in inches:

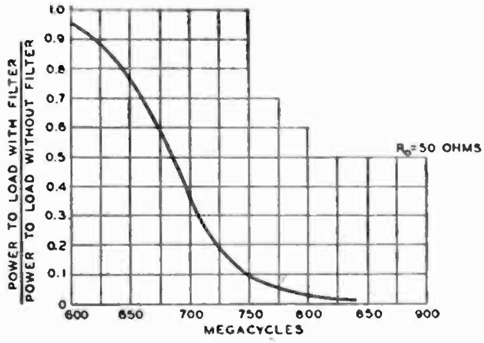

Fig. 4—Transmission curve of 800-megacycle coaxial filter in the vicinity of cutoff.

f_c	Number of Sections	R_o	r_o	r_1	r_2	d_1	d_2
800	4	50	0.0937	0.500	0.450	2.34	0.583
1800	4	50	0.0468	0.3125	0.275	0.95	0.312

Measurements were made with and without the filters between an oscillator and a 50-ohm load using 50-ohm coaxial cable for frequencies in the range of 500 to 3750 megacycles. The ratio of the power to the load with and without the filter for both filters is shown in Figures 4 and 5 in the vicinity of cut-off. Although the actual power transfer through the filter is a function of the impedance reflected back to the oscillator and so will vary with different oscillators and filters, the cut-off frequency of the filter will not be appreciably affected by the over-all circuit and it is this cut-off region which is of interest. The 800-megacycle filter design adheres closely to Equations (12-15) and Figure 4 shows its experimentally observed cut-off to be closely in accord with the predicted value. The 1800-megacycle filter's experi-

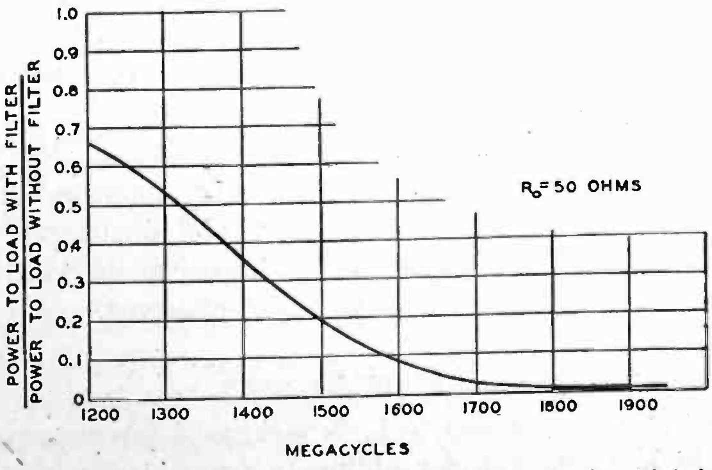

Fig. 5—Transmission curve of 1800-megacycle filter in the vicinity of cutoff.

mentally observed cut-off in Figure 5 does not agree exactly with the predicted cut-off since it is evident from the dimensions listed in the table that Equation (12) is exceeded to such an extent that the angle in Equation (7) is not an accurate approximation of the tangent of the angle. The effective cut-off of the 1800-megacycle filter is seen in Figure 5 to be in the neighborhood of 1650 megacycles and represents an over-all error of about eight per cent.

There are frequencies above cut-off where pass bands will result due to resonances on the part of either the capacitance or the inductance sections. At 2690 megacycles, for example, the 800-megacycle filter was found to have an extremely sharp pass band in which considerable transmission was observed. This is precisely that frequency at which the inductance section length, d_1, is equal to a half wavelength and the filter becomes a line composed of a series of resonant sections connected by short lengths of low characteristic impedance line. Resonance on the part of the capacitance section will also produce a pass band. These pass bands may be predicted and if, in general, d_1 and d_2 are made very small compared to the cut-off wavelength, the resonant pass bands will be far removed from the main pass band of the filter. It is possible to decrease the transmission in these pass bands by including sections in the structure with different d_1 and d_2 which are designed to cut-off at the correct frequency.

Since the coaxial low-pass filter will decrease in size as the cut-off frequency is increased, the upper limit of design and usage will be determined by mechanical considerations which will become important above 5000 megacycles. It is particularly important at these higher frequencies that the connectors do not introduce reflections of large magnitude and that the termination be made as close to R_o in value as possible. In the specific application of the filters being described, a termination made up of a small 50-ohm resistor which was bent into the shape of a loop was found to function satisfactorily, but at higher frequencies the termination must be carefully engineered to preserve the cut-off characteristic of the filter.

When installed into the cold-resonance equipment for which they were designed, the low-pass coaxial filters were found to be completely satisfactory with respect to performance and during the course of numerous experiments, indications due to higher modes getting through the high-pass bands of the filters were not observed.

ACKNOWLEDGMENT

The authors are indebted to L. S. Nergaard for numerous suggestions and to L. E. Norton for aiding in transmission measurements.

APPENDIX

A. Consider the use of a short length of coaxial transmission line, having zero dissipation, as a series-lumped inductance in series with a terminating impedance, Z_L. Let the characteristic impedance of this line be Z_o and let it be terminated by Z_L such that the input impedance, Z_1, may be written

$$Z_1 = Z_o \frac{Z_L \cos \beta l + j Z_o \sin \beta l}{Z_o \cos \beta l + j Z_L \sin \beta l} \qquad (1\text{-a})$$

where the length of the line is l and

$$Z_o = \sqrt{\frac{L_o}{C_o}} \qquad (2\text{-a}) \qquad\qquad \beta l = \omega \sqrt{L_o C_o}\, l \qquad (3\text{-a})$$

L_o and C_o are the inductance and capacitance per unit length of line, respectively. Equation (1-a) may be rewritten

$$Z_1 = \frac{Z_L + jZ_o \tan \beta l}{1 + j \dfrac{Z_L}{Z_o} \tan \beta l} \qquad (4\text{-a})$$

If the length, l, and the ratio, $\dfrac{Z_L}{Z_o}$, are such that

$$\left| \frac{Z_L}{Z_o} \tan \beta l \right| << 1 \qquad (5\text{-a})$$

and $\quad \tan \beta l \approx \beta l \quad (6\text{-a}) \qquad$ then $\quad Z_1 = Z_L + j\omega L_o l \quad (7\text{-a})$

which shows that the short length of line will appear to be a series inductance if (5-a) and (6-a) are adhered to.

B. Consider the use of a short length of coaxial transmission line, having zero dissipation, as a lumped capacitance in shunt with a terminating impedance Z_L. Let the characteristic impedance of this line be Z_o and let it be terminated by Z_L such that the input admittance, Y_1, may be written

$$Y_1 = \frac{1}{Z_o} \frac{Z_o \cos \beta l + j Z_L \sin \beta l}{Z_L \cos \beta l + j Z_o \sin \beta l} \tag{8-a}$$

$$= \frac{\dfrac{1}{Z_L} + j \dfrac{1}{Z_o} \tan \beta l}{1 + j \dfrac{Z_o}{Z_L} \tan \beta l} \tag{9-a}$$

If the length, l, and the ratio, $\dfrac{Z_o}{Z_L}$, are such that

$$\left| \frac{Z_o}{Z_L} \tan \beta l \right| << 1 \tag{10-a}$$

and $\tan \beta l \approx \beta l$ (6-a) then $Y_1 = \dfrac{1}{Z_L} + j\omega C_o l$ (11-a)

which shows that the short length of line will appear to be a shunt capacitance if (10-a) and (6-a) are adhered to.

If the radius of the outer conductor, r_a, is much larger than the radius of the inner conductor, r_b, fringing may not be neglected and it is convenient to include an approximate fringing factor such that

$$Y_1 \cong \frac{1}{Z_L} + j\omega \left[1 + \frac{r_a - r_b}{l} \right] C_o l \tag{12-a}$$

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RECENT DEVELOPMENTS IN RADIOTELEGRAPH TRANSMITTERS FOR SHORE STATIONS*

BY

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Summary—This paper describes three new types of radiotelegraph transmitters designed for the maritime mobile services. One type delivers 20 kilowatts of antenna power with push button selection of two frequencies in the 350-500 kilocycle band. A second type, otherwise similar to the 20-kilowatt transmitter, provides an antenna power of 50 kilowatts. A third type for the 2-18 megacycle band is rated at 15 kilowatts and includes two complete radio-frequency units for push button selection of two channels from a common power supply.

INTRODUCTION

DURING the past ten or twelve years relatively little material has been published on the development and design of high-power radiotelegraph transmitters for the maritime mobile services. Such transmitters, with a power output of 20 to 50 kilowatts in the 350-500 kilocycle medium-frequency band, or 15 kilowatts in the 2-18 megacycle high-frequency band, may be considered the backbone of shore to ship communication. From the safety at sea aspect they are a very important factor in the transmission of distress signals, hurrican warnings, and weather reports, and for the overall coordination of rescue operations. With more and more aircraft flying over the oceans, powerful shore-based stations become an essential part of air-sea communication networks for safety purposes. Less spectacular, but of great economic value, the transmitters serve thousands of vessels as the link between shore and ship for business and personal message traffic.

GENERAL FEATURES

A well coordinated group of medium and high frequency transmitters has been designed in connection with a project for the U. S. Coast Guard, which prepared rigid specifications for modern equipment to meet the special needs of their safety services. Particular attention was directed to achieve technical performance which would anticipate rigorous future requirements for such characteristics as frequency stability,

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* Decimal Classification: R355 × R423.2.

harmonic reduction, rapid frequency shift and the shaping of the keyed signals. In addition the mechanical design conforms to modern practices in presenting a pleasing appearance.

To the engineer accustomed to dealing with very-high-frequency or microwave equipment, the transmitters are rather large in a physical as well as an electrical sense. For example, the 20-kilowatt or 50-kilowatt medium frequency equipment is 34 feet long and 8 feet high, Figure 1, and is made up of six frames, 4 × 4 × 7 feet, which are attached to the front enclosure. The frames are welded steel channels protected against corrosion with zinc coating which is applied by hot spraying. The front enclosure is of sheet steel construction with a door three feet wide in front of each frame. The doors are of rigid steel construction with concealed hardware and an interleaved lining of sound insulation. The viewing ports are of safety glass. Double doors at each end of the enclosure provide access to a four foot passageway to the rear of the transmitter. Large heavy components, such as the main rectifier

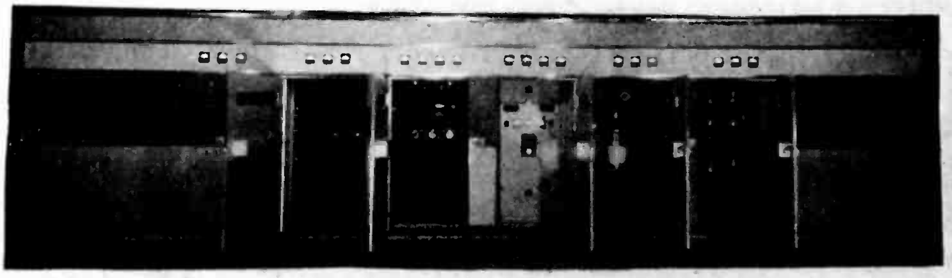

Fig. 1—Medium frequency transmitter.

plate transformers and variometers, are separated from the frames and located at the rear of the transmitter.

Loading inductors which tune the antenna to resonance are inherently large and must have low losses when carrying currents up to 125 amperes at 70 kilovolts and 350 kilocycles. The antenna, usually a 300-foot insulated tower, is only a small fraction of the operating wavelength and must be tuned both to 500 kilocycles and some lower frequency down to 350 kilocycles. The medium frequency transmitter includes continuously variable dual tuning circuits from the crystal stage to the antenna circuit with relays to effect the frequency shift, instantaneous choice of either frequency being provided.

The 20-kilowatt medium frequency transmitter has the following tube complement:

6AC7 crystal oscillator
807 class-A buffer amplifier
807 class-C intermediate amplifier

6AC7 audio oscillator
807 (2) audio amplifiers
8000 (2) class-B modulators

813 (2) class-C plate-modulated amplifiers
9C22 power amplifier
813 (2) hum suppressors
807 keying tube
869-B (6) main rectifier
5R4GY regulating diode
VR-105 voltage regulator

The 50-kilowatt transmitter uses a similar arrangement of tubes except that four 813 class-C plate-modulated amplifiers and two 9C22 power amplifiers are required. The audio oscillator-modulator portion of the circuit provides A2 emission (modulated continuous waves) and is disconnected for A1 emission (continuous waves).

The 15-kilowatt high frequency transmitter is a dual design comprising two complete radio frequency units and a common power supply unit. Each radio frequency unit may be adjusted to any frequency between 2 and 18 megacycles and provides A1 emission only. Two dual high-frequency equipments are housed in an enclosure which matches in size and appearance that of a medium-frequency transmitter. With this arrangement there are available four pretuned high frequency channels and two power supply units, with simultaneous transmission on two channels if desired. Normally one pair of channels (above approximately 8 megacycles) are set up for daytime transmission and the other pair (below 8 megacycles) for night service.

Each 15-kilowatt radio frequency unit uses the following tube layout:

6AC7 Crystal Oscillator
807 Buffer
807 Radio Frequency Amplifier
807 (2) Double Stages
813 (4) Push-Pull Intermediate Power Amplifiers
807 Keying Tube
VR-75 Voltage Regulator
VR-105 Voltage Regulator
673 (6) Main Rectifier
889-RA (2) Power Amplifiers

The electrical components used in the medium and high frequency transmitters were selected to conform to the following requirements. All small transformers and reactors are hermetically sealed. Large transformers and reactors are oil filled. Mica and paper dielectric capacitors and wire wound resistors conform to Joint Army-Navy specifications. Vacuum capacitors are used, some fixed and some variable, where the capacitance required is less than 250 micromicrofarads for voltages from 15 to 50 kilovolts. For higher capacitance values, 1000 to 4000 micromicrofarads, pressurized nitrogen filled capacitors are used. Insulation of the ceramic or glass-bound mica variety is employed for tube sockets, radio frequency coils, and switches.

CONTROL CONSOLE

The control console, Figure 2, contains controls for the switching functions for two dual high frequency transmitters and one medium frequency transmitter. All the switching functions can be controlled at the transmitter or at the control console with a parallel arrangement of momentary contact push button switches, including Start-Stop, Standby-Operate, Frequency Change, Emission Change, and Keying. Additional switches are provided on the console to transfer the above controls from the console to switches at a remote point through a land line or radio link. An audio amplifier, monitoring speaker and a volume indicator are provided for monitoring a tone converter used in connection with the remote control. Indicator lights are associated with all the switching function controls at the transmitter and at the console, and an alarm bell sounds a warning in the event of an overload or undervoltage or insufficient cooling air to the power amplifier tubes. The keying circuits are arranged for keying the

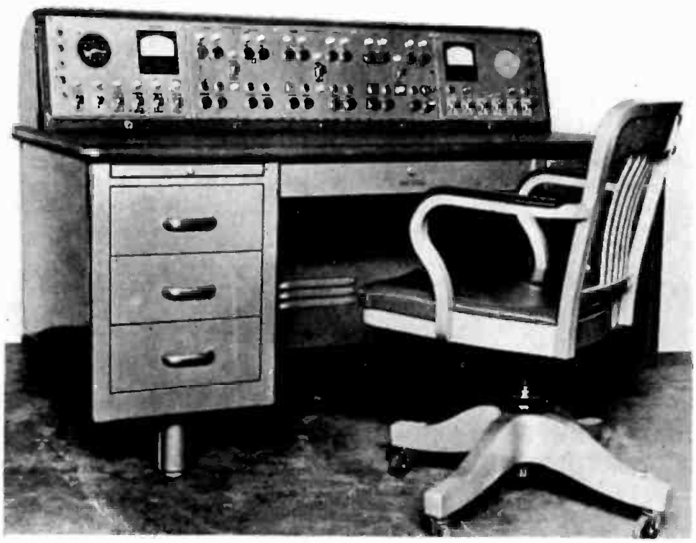

Fig. 2—Control console.

transmitters individually or collectively from the console or the remote point.

The control console is mounted on a double pedestal flat top steel desk. The controls are symmetrically arranged on three satin chrome finished brass panels which are hinged for convenient servicing. The entire rear of the console and all terminals are enclosed in a one-piece streamlined hinged cover. A 48-volt auxiliary rectifier and an amplifier are located in a steel cabinet which is built into the rear of the desk between the pedestals. All wiring to the console enters through the bottom of the pedestals and is fully concealed.

Rectifiers

All rectifiers, except the main rectifier, employ selenium. These rectifiers provide outputs ranging from 5 amperes at 48 volts for the control circuits to 1 ampere at 1900 volts as plate supply for the intermediate amplifier. The selenium stack for the 48-volt rectifier consists of 12 cells, 4⅜ inches in diameter. The 1900-volt selenium unit consists of ninety-six cells, 1¾ inches in diameter, in each of the six legs of a three phase full wave rectifier. One hundred percent voltage control is provided in these rectifiers to maintain close adjustment of the operating potentials for the various stages, as well as to compensate for line voltage variations and aging of the selenium. This control is effected through the use of two-gang variable transformers connected in open delta operating from a 3-phase 110-volt secondary line. Each transmitter has a 400-volt bias rectifier and a 400-volt plate supply rectifier.

The main rectifiers for all transmitters use mercury vapor tubes connected three phase full wave. The filament transformers are arranged to maintain out-of-phase relationship between the filament and plate voltages for maximum efficiency. Careful consideration was given to the ripple filters associated with these rectifiers in order to insure a regular keying pulse shape and to minimize keying transients.[1]

Main Rectifier Ratings and Filters

	Output Rating	*Filter Reactor*	*Filter Capacitor*
#1	16,000 Volts 7.2 Amperes	1 Henry	30 Microfarads
#2	16,000 Volts 3.6 Amperes	2 Henries	15 Microfarads
#3	7,500 Volts 4.0 Amperes	1 Henry	30 Microfarads

Rectifier #1 is used for the 50-kilowatt medium frequency transmitter and rectifier #2 for the 20-kilowatt set. Rectifier #3 supplies the 15-kilowatt high-frequency transmitter.

A surge limiting circuit is employed to restrict the load on the rectifier tubes when the a-c plate voltage is first applied. This restriction is accomplished by the insertion of a resistor in series with the filter capacitors. The resistor is removed from the circuit by a time delay relay after approximately 5 seconds. Bleeder resistors are not used on the output of the rectifier due to the large power loss if the bleeder is to be effective. However, an additional contact on the anti-surge relay applies a grounding resistor to the output of the rectifier when the a-c plate voltage is removed. Properly interlocked circuits

[1] R. Lee, "Radiotelegraph Keying Transients", *Proc. I.R.E.*, Vol. 22, No. 2, February, 1934.

remove this grounding resistor an instant before the plate voltage is applied.

Overload Relays

The transmitters operate on 440 volts, 3 phase, 60 cycles, and are safeguarded against damage by triple overload protection. Alternating current operated relays and contactors are avoided to eliminate acoustical hum. The start-stop contactors are alternating current operated, but are of the permanent magnet type requiring no energy for hold-in. Three-pole magnetic circuit breakers are used on the input circuits to each blower motor and rectifier. These circuit breakers are mounted on the front panel for convenience as manually operated branch circuit switches. An overcurrent relay is used in each power amplifier plate return circuit. These relays have separate time delay adjustments for pick-up and drop-out. A notching relay is used in conjunction with the overcurrent relays to accept three overloads before locking out.

Branch circuits are also protected with 3-pole magnetic circuit breakers. Suitable relays provide overcurrent and undervoltage protection at the critical circuit positions. Fuses are used only to give protection against excessive heat in the oscillator ovens.

High Frequency Oscillator and Amplifiers

The high frequency oscillator circuit is of the tuned-grid tuned-plate type using a crystal unit as the grid circuit tank and a balanced rotor type air capacitor with its associated inductor to tune the plate circuit. The oscillator tube is of the 6AC7 type using 3 milliamperes at 105 volts d-c on the plate. Heater and plate voltages are regulated to stabilize the oscillator against variable line voltage. The rectified grid current is less than 1 milliampere, thus insuring negligible effect of crystal current on the frequency of the oscillator. Provision for a total of ten crystals is made in the oscillator oven, and front panel switching for selection of any of the ten frequencies is provided. Exact adjustment of the oscillator frequency is made by use of individual variable trimmer capacitors in parallel with each crystal. A frequency coverage of 2 to 4.5 megacycles is provided in three-band ranges by front panel switching.

Over a range of ambient temperatures of 10 to 50 degrees Centigrade, the frequency stability is within $\pm$ 0.0004 per cent. Over an initial two-hour period the frequency drift of the oscillator is less than 0.0005 per cent. Crystals are ground to a tolerance of $\pm$ 0.003 per cent at the crystal-oven temperature. The effect of humidity results in a frequency change of less than 0.0002 per cent.

The buffer amplifier for the high frequency transmitter is a class-A amplifier following the oscillator stage, and uses a type 807 tube which supplies approximately 0.5 watt to drive the succeeding stage. A frequency monitoring jack is coupled to the buffer tank to supply approximately 300 millivolts to a 52-ohm line. Two 807 doubler stages and an 807 amplifier stage follow the buffer to drive four type 813 tetrodes connected in push-pull-parallel as intermediate power amplifiers. Two type 889-RA triodes are used in the power amplifier operating at 7500 volts plate potential. The power amplifier grid and the intermediate power amplifier plate utilize separate tuned tank circuits which are coupled together with an untuned link. The power amplifiers are cross neutralized with variable vacuum capacitors controlled from the front panel.

The push-pull power amplifier tank circuit is tuned with variable vacuum capacitors having a range of 20 to 250 micromicrofarads on each side of the tank. The antenna matching network comprises a series of air capacitors and a tuning inductance. A portion of the antenna tuning inductance is coupled to the power amplifier tank through a variable inductor. This inductor is a variable shunt across a portion of the coupling inductor to obtain the desired coupling which is adjustable from the front panel. The output circuit is designed to deliver 15 kilowatts to a 52 ohm coaxial cable type RG-20/U. The entire range of 2 to 18 megacycles requires six sets of power amplifier inductors which are proportioned to suit the bands in which they operate. Figure 3 shows a front view of one radio frequency unit with the power supply unit on the right. A rear view of the radio frequency unit is shown in Figure 4.

Mica capacitors are used to keep direct current blocked from the vacuum capacitors. This is necessary to prevent damage to tubes and circuits in the event of a gaseous arc-over within the vacuum capacitors. In general, mica capacitors are avoided wherever it is convenient to use an air dielectric type due to the wide range of currents encountered over the frequency band.

Medium Frequency Oscillator and Amplifiers

The oscillator circuit of the medium frequency transmitter (350-500 kilocycles) is untuned and is a Colpitts type using the crystal as the tank circuit connected between the plate and grid of the type 6AC7 oscillator tube. The basic circuit is shown in Figure 5. The capacitor C-4 in series with the crystal allows precise alignment of the crystal to its nominal frequency. Capacitor C-1 and the combination of capacitors C-2 and C-3 act as a radio frequency voltage divider between the plate and grid circuits of the oscillator.

The inductance L and capacitor C-5 act as a harmonic trap in the plate circuit and are tuned to a frequency to present high impedance to second harmonic frequencies of the crystals over the range of 350 to 500 kilocycles. Output from the oscillator is taken from the junction of capacitors C-2 and C-3. The filament voltage and the plate voltage are regulated in order to secure stable operation. The plate voltage applied to the tube is 105 volts d-c and the plate current is approximately 1.7 milliamperes. The rectified grid current is in the range of 30 to 50 microamperes, depending on the activity of the crystal.

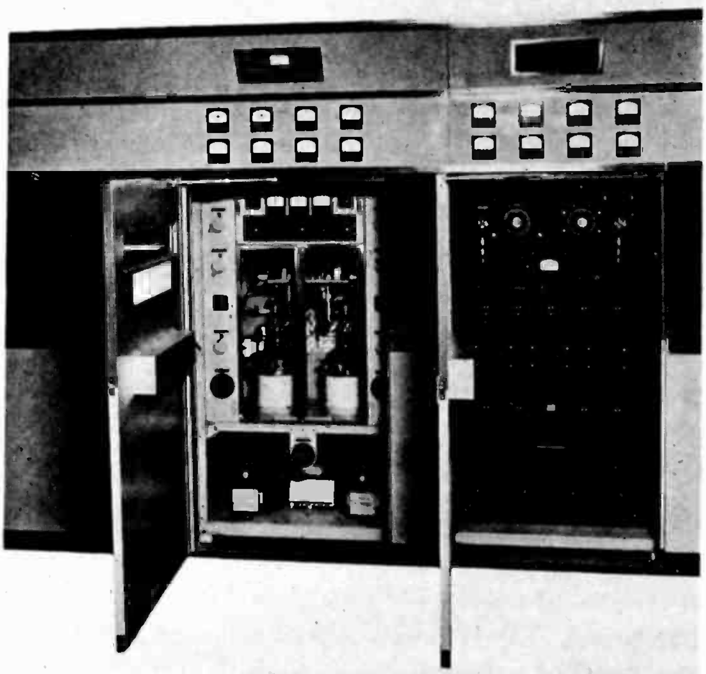

Fig. 3—Front view of one radio frequency unit with power supply unit on right.

The crystal unit of the CR-7-T type consists of a CT cut quartz plate, silver plated and wire mounted in a small hermetically sealed metal container. Positions for six crystals are provided in the oscillator oven, and the choice of crystal for operation being made by a switch on the front panel of the exciter unit. Any one of the six crystals may be selected for use on either of the two frequency channels provided in this transmitter. The crystals are ground to a frequency tolerance of ±0.004 per cent at the oven temperature. The crystal trimmer capacitors have adequate range to permit adjustment of frequency to within ±0.0005 per cent of the reference frequency used. The fre-

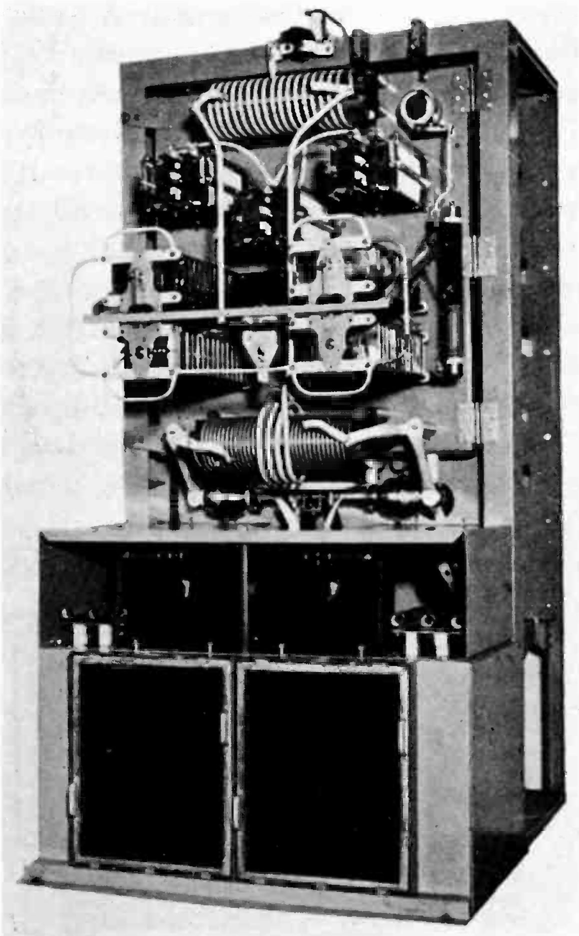

Fig. 4—Rear view of radio frequency unit.

quency stability of this oscillator is comparable to that of the oscillator used in the high frequency transmitter.

Figure 6 shows the construction of the oscillator oven. The low frequency oscillator temperature is maintained at 70 degrees Centigrade and the high frequency oven temperature at 60 degrees Centigrade with approximately 30 watts of heater power at normal ambient temperature. The thermostat is of the mercury thermometer type

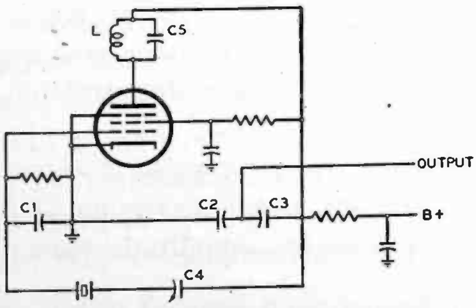

Fig. 5—Oscillator circuit.

with a differential of 0.1 degrees Centigrade. A heater using approximately 25 milliwatts is applied to the thermostat[2]. A potentiometer is used for control of the thermostat heater power to calibrate the oven temperature. The thermostat contacts handle 6 milliwatts at 6.3 volts, 60 cycles, operating a relay to power the oven heaters. The temperature of the inner oven is measured by a bi-metallic coil thermometer with the indicator scale located on the front panel.

The medium frequency transmitter uses one type 9C22 triode as the power amplifier for the 20-kilowatt transmitter. Two 9C22 tubes are used in the 50-kilowatt transmitter. The power output on A2 emission is approximately 70 per cent of A1 rating. For A2 emission, the power amplifier is operated class-B and the driver stage is plate modulated. Since the same adjustment of the power amplifier and

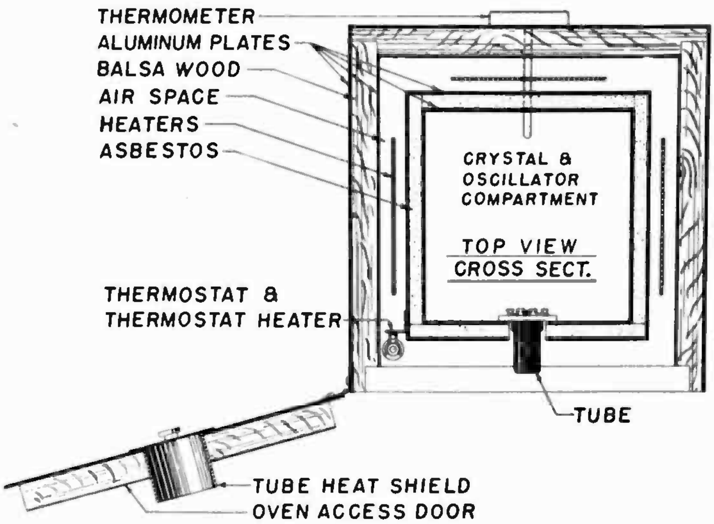

Fig. 6—Oscillator oven.

antenna tuning and coupling must be used for both A1 and A2 emission at any particular frequency, it is necessary to maintain the proper operating conditions for the required output. This is accomplished by separate controls for the plate voltage of the driver stage. The Rice system of neutralization is used in the power amplifier stage. A regulating diode 5R4GY is used in the power amplifier grid circuit to suppress parasitic oscillations. A resistor and capacitor are used in conjunction with the diode. The capacitor functions to shape the leading edge of the keying pulse.

The rise and fall of the keying pulse, Figure 7, is confined to a minimum of 0.3 milliseconds and a maximum of 3 milliseconds for a pulse length of 25 milliseconds. Amplitude variation is restricted to

[2] J. K. Clapp, "Notes on the Design of Temperature Control Units", *General Radio Experimenter*, Vol. 19, No. 3, August, 1944.

±10 per cent. Delay circuits are applied to the keyer tube for some control of the pulse shape. The shape control is somewhat reduced if the keyed stage is followed by several amplifier or multiplier stages, in which case it is necessary to key two stages. The amplifier stages which are not keyed are biased to cut off. The regulation of the grid bias voltage also has some influence on the keying pulse shape. The oscillator is allowed to operate with the "key up". The buffer stage is keyed in order to minimize the shielding necessary to prevent radiation in the "key up" condition.

Much has been written on the treatment of parasitic oscillations[3]. The low power stages are readily purged through the use of suppressor resistors in the grids and the familiar choke and resistor combination in the plate circuits. The power amplifier stage required special consideration given to the length of leads in the grid and plate circuits. Damping circuits are essential in the neutralizing circuits to suppress feedback on the harmonic frequencies. Damping is also necessary in

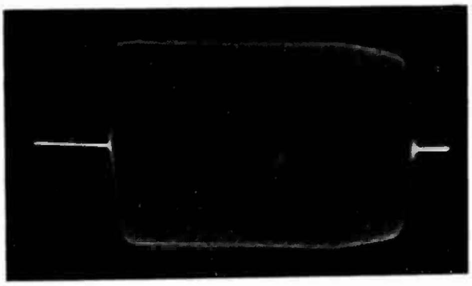

Fig. 7—Keying pulse.

the power amplifier grid tank to eliminate a tendency of the power amplifier to self-oscillate at the fundamental frequency even with neutralization properly adjusted. The tendency to self-oscillate is shown by the appearance of irregular tails on the keying pulse. When the grid circuit is sufficiently damped, the neutralizing adjustment is less critical.

Litz wound variometers, Figure 8, provide separate adjustment of power amplifier tuning, antenna tuning and antenna coupling in each channel. The variometers are located to the rear of the transmitter frames and the controls are extended to the front panel through the use of synchros geared down 50 to 1. Special contactors were designed for switching channels in the power amplifier and antenna circuits for the medium frequency transmitter. These contactors handle radio frequency potentials up to 75 kilovolts crest at 125 amperes.

Each medium frequency transmitter includes six variometers which

[3] G. W. Fyler, "Parasites and Instability in Radio Transmitters", *Proc. I.R.E.*, Vol. 23, No. 9, pp. 985-1012, September, 1935.

make up the power amplifier and antenna tuning elements for two instantly selected channels. The variometers are wound with 5 conductors in parallel per turn. Each Litzendraht conductor consists of 128 strands of #36 wire. The variometers for the 50-kilowatt antenna circuit are wound with ten conductors in parallel.

One of the main rectifier filter capacitors in the 350-500 kilocycle transmitter is used as a radio-frequency plate by-pass. This capacitor is isolated from the remaining capacitors by additional current limiting resistors which remain in the circuit at all times. The resistors serve a dual purpose in that they limit the peak current drawn from the capacitors in the event of an overload, and they also confine the radio frequency current to the single capacitor.

Considerable hum can appear on the carrier due to the 8 kilowatt filament of the 9C22 power amplifier tube when operating at the low grid current required for class-B operation. A circuit was developed wherein a sample of hum voltage is obtained from the power amplifier

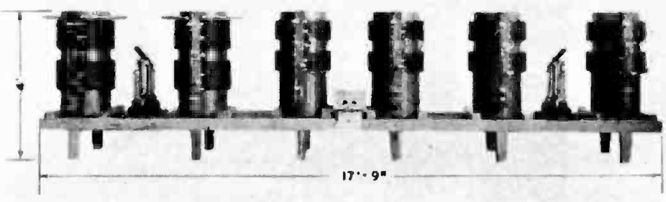

Fig. 8—Litz wound variometers.

cathode resistor. This voltage, properly phased, is fed into two type 813 tetrodes which modulate the intermediate power amplifier at the hum frequencies. The hum level, before suppression was applied, amounted to —20 decibels as referred to 100 per cent modulation. After suppression was applied, the hum level was reduced to —40 decibels. Hum suppression is applied for both A1 and A2 emission and, since there are no resonant circuits tuned to the carrier frequency for the hum pickup voltage, the circuit performs well regardless of the carrier frequency selected.

The 350-500 kilocycle transmitter circuits are designed to match antennas whose characteristics vary from 3-j455 to 6-j152 at 350 kilocycles and 6-j255 to 12-j77 at 500 kilocycles. A 4000-micromicrofarad pressurized capacitor is used in series with the antenna where the antenna capacitance approaches 4000 micromicrofarads at the higher frequencies. The antenna is maintained at d-c ground potential to prevent the accumulation of high static potentials. Antenna current metering is effected through a current transformer in the ground lead of the antenna coupling variometers. The secondary voltage of the transformer is rectified and measured with a d-c meter.

The antenna coupling variometer provides a smooth stepless control of coupling by combining portions of the antenna and power amplifier tuning inductances in a single unit. This arrangement allows coupling adjustments to be made with a minimum effect on the power amplifier and antenna tuning.

ANTENNA CONSIDERATIONS

It is necessary to provide tower lighting circuits for the 300-foot insulated tower which is used as a radiator in the 350-500 kilocycle

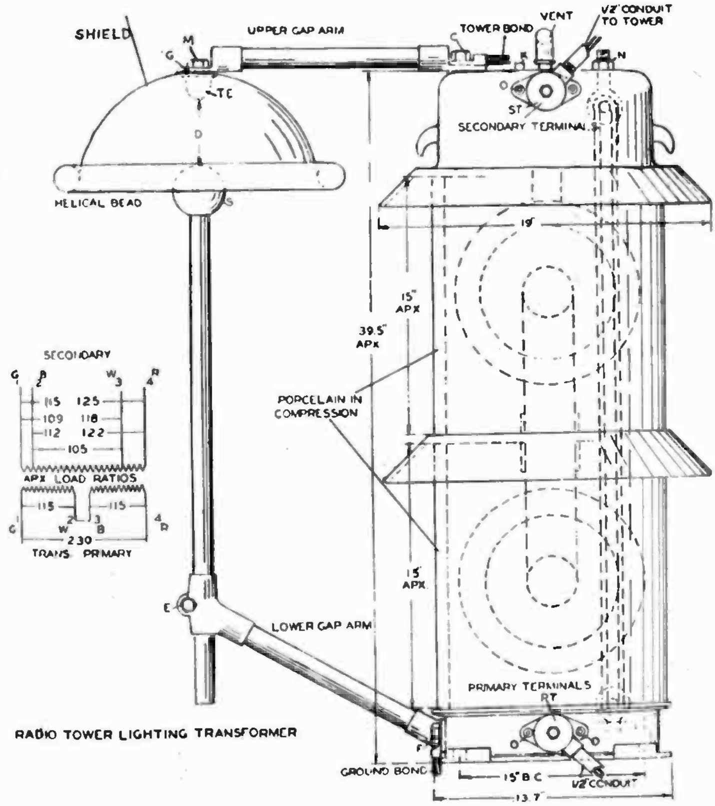

Fig. 9—Tower lighting transformer.

band. With the 50-kilowatt transmitter operating at 350 kilocycles, the radio frequency voltage present on the antenna base insulator is about 75 kilovolts. A radio-frequency choke to carry the tower lighting load of 1500 watts at 115 volts becomes unwieldy and inefficient for medium frequencies. Use is, therefore, made of an oil-insulated tower-lighting transformer shown in Figure 9. The transformer is provided with a protective gap and rain shield. With rain precipitation of 0.2 inch per minute, the radio frequency arc-over voltage is approximately 100 kilovolts. This may be compared with 275 kilovolts arc-over under precipitation at 60 cycles. The transformer introduces an additional capacitance of 65 micromicrofarads. Since two-thirds of the tower lighting load is a flashing load, the transformer includes a compensating impedance with two opposed windings to maintain the voltage regulation within 3 per cent. The flasher mechanism and the compensating impedance are mounted on the tower and, hence, are at radio frequency potential.

In order to test several transmitters of the types described in this paper, suitable artificial antennas capable of dissipating considerable power are necessary. A circulating pump delivering 15 gallons of water per minute, in conjunction with a fan cooled radiator, provides for a total dissipation of 60 kilowatts. The resistive element for the 350-500 kilocycle artificial antenna consists of Tophet "A" Ribbons, #29 gauge by 3/32 inch. This material has a resistance of 0.499 ohms per foot. Various values ranging from 3 to 12 ohms are mounted in a cylindrical container 6 inches in diameter and 12 inches high. Water is circulated through this container to carry off the heat. A tap switch is immersed in the water for selecting the required resistance, with the switch control brought out through a watertight gland.

The capacitive element for the 350-500 kilocycle artificial antenna comprises a series-parallel group of pressurized nitrogen-filled capacitors having a capacity range of 1000 to 4000 micromicrofarads and capable of carrying 125 amperes.

For the 2-18 megacycle band the resistors for the artificial antenna are of the open grid type rated at 275 watts in air. These resistors are capable of dissipating 15 kilowatts when mounted in a cylindrical circulating water container similar to that used for the medium frequency transmitter. A corrective network is used in conjunction with the resistive element to compensate for the inherent reactance of the resistor in order to present a 52-ohm termination for the transmitter under test.

RCA TECHNICAL PAPERS†

Third Quarter, 1947

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Electronics (August) 1947
Broadcast News (September) 1947

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NOTE—Omissions or errors in these listings will be corrected in the yearly index.

Correction:

On page 526 of the September 1947 issue, a typographical error appeared in the second line beneath Figure 18. The equation as printed was: $n_1/n_2 < 1$. This should have read: $n_2/n_1 < 1$.

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JOSEPH F. MCDONALD completed a course in mechanical and electrical design at the Elizabeth Vocational School in 1926. He was employed by the Weston Electrical Instrument Company for 3 years and then served 3 years as radio service manager for a chain of stores in New Jersey. In 1936-1937 he completed an evening course at the RCA Institutes serving a short time as assistant instructor. From 1933 to 1936 he was studio engineer with WBNX. Since 1936 he has been engaged in the development and design of transmitters for the Radiomarine Corporation of America. Mr. McDonald is an Associate Member of the I.R.E.

ADOLPH R. MORGAN received his B.S. degree in Electrical Engineering from the University of California in 1931. In the same year he joined the RCA Manufacturing Company in their Student Training Activities and in 1933 became a research engineer. In 1942 he transferred to RCA Laboratories Division in Princeton, N. J., where he has been engaged in recording and acoustic research problems. Mr. Morgan is a Member of the Acoustical Society of America, Tau Beta Pi, Eta Kappa Nu, Sigma Xi and Phi Beta Kappa.

HARRY F. OLSON received the B.S. degree in 1924, the M.S. degree in 1925, the Ph.D. degree in 1928 and the E.E. degree in 1932 from the University of Iowa. From 1928 to 1930 he was in the research department of Radio Corporation of America; from 1930 to 1932, in the engineering department of RCA Photophone; from 1932 to 1941, in the research division of RCA Manufacturing Company; since 1941, with RCA Laboratories Division. Dr. Olson is a member of Tau Beta Pi, Sigma Xi and the American Physical Society, an Associate Member of the Institute of Radio Engineers and a Fellow of the Acoustical Society of America.

JOHN PRESTON received his education in England, attended Manchester Scientific Institute taking courses in Mechanical Engineering. He then joined the engineering staff of the Bolton Power and Electrical Company. He became a member of Radio Corporation of America in 1929 and was associated with the Research Department of RCA Victor from 1930 to 1935; the Research Department of RCA Manufacturing Company from 1935 to 1941, and RCA Laboratories Division from 1941 to date. His work has been in the field of Acoustics since joining Radio Corporation of America. Mr. Preston is a member of the Acoustical Society of America and of Sigma Xi.

HAROLD J. SCHRADER graduated from the University of Nebraska. In 1923 he entered into the electrical contract business in Lincoln, Nebraska, and continued until 1925 when he joined the General Electric Co., at Schenectady, N. Y. In 1930 he became a member of the Factory Test Equipment Group of the RCA Manufacturing Co. and in 1943 he was assigned to the Combined Research Group of the Naval Research Laboratory in Washington, D. C. Returning to Camden in 1945 he joined the Teleran Engineering Section as Supervising Engineer. Mr. Schrader is a Senior Member of the I.R.E.

CARL I. SHULMAN received the B.S. and M.S. degrees in electrical engineering in 1938 and 1939 from the Massachusetts Institute of Technology. From June 1939, to December, 1940, he was employed by the Submarine Signal Company of Boston, and from December 1940 to date has been with RCA Laboratories Division at Princeton, N. J., working on ultra-high-frequency tube problems. He is now also a part-time graduate student in the Physics Department of Princeton University. Mr. Shulman is a Member of the American Physical Society and Sigma Xi.

PHILIP F. SILING received the Ph.D. degree in E.E., from Yale University in 1917. He was employed by the American Telephone and Telegraph Company from 1917 to 1929, International Telephone and Telegraph Corporation until 1933, and as Assistant Deputy Administrator of the National Recovery Administration, in charge of codes of the electrical manufacturing industry. From 1935 until 1945 he served in the Federal Communications Commission. During World War II he was active in the work of the Board of War Communications. In 1945 he joined RCA as Engineer-in-Charge of the RCA Frequency Bureau. Mr. Siling is a Senior Member of the Institute of Radio Engineers, and a Member of Sigma Xi.

S. H. SIMPSON JR. received the B.S. degree in Electrical Engineering from Texas A & M College in 1928, when he joined RCA working in the Technical and Test Department. From 1930 to 1933 he was with RCA Communications, Inc. in Riverhead, L. I., transferring in 1933 to the New York central office as Manager of Program and Radiophoto Services. From 1941 to 1945 he served in the armed forces, first in the Censorship Bureau for International Communications, and later as Deputy Chief of the Communications Branch O.S.S., of the E.T.O., also heading a special intelligence unit. In October 1945 he returned to RCA Communications. Mr. Simpson is a Member of the I.R.E. and A.I.E.E.

ROBERT W. K. SMITH received his B.S. degree in Physics from Oregon State College in 1933. He was with Warner Bros. as sound technician and junior engineer for seven years. Following a year with Bendix Aviation as radio engineer, he joined Sperry Gyroscope Company in 1941 as an engineer on microwave applications for air navigation and communications. He is now with RCA Victor Division in Camden, N. J., as a supervisory engineer, in the Teleran Engineering Section.

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