Proceedings of
The Radio Club of America, Inc.

Founded 1909

Volume 28, No. 2

1951

SURVEY OF RADIO-FREQUENCY TRANSMISSION LINES AND WAVE GUIDES

By

E. S. Winlund

THE RADIO CLUB OF AMERICA

11 West 42nd Street ★ ★ ★ New York City

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By E. S. Winlund*

Submitted originally as a seminar paper at the Massachusetts Institute of Technology.

Scope

This paper gives the essential technical data from various published material on the principles of radio-frequency transmission lines and wave guides appearing between 1919 and 1936, followed by a bibliography covering later years. During this time the art grew to a mature state. The material was prepared and presented as a seminar paper at the Massachusetts Institute of Technology.

General Development Prior to 1925

The first successful effort toward evaluation of the magnetic and electric fields about a conductor in terms of the current through it and the constants of the materials involved appears to begin with Maxwell's well-known equations:

$$\begin{aligned} \text{curl } E &= -\mu i p H \\ \text{curl } H &= (4\pi\lambda + K i p) E \\ \text{div } E &= 0 \\ \text{div } H &= 0 \end{aligned} \tag{1}$$

where E is the electromotive force, H the magnetic force, μ the permeability, λ the conductivity, and K the specific inductive capacity of the medium. Indeed there is abundant reference by transmission line mathematicians to his original paper[1] of 1881, but apparently no reference to papers before that time. Soon after that Poynting[2] in 1885 and Rayleigh[3] in 1886 discussed the elementary mechanism of the field about a single conductor and treated mathematically the inductance and resistance, respectively, of single straight conductors. Hertz[4] in 1889 contributed and Heaviside[5] in 1893 remodelled Maxwell's equations and derived a submarine cable formula which was no doubt the first published recognition of coaxial lines as such. It is generally conceded that Heaviside's and Kelvin's work mark the beginning of transmission line theory.

Gustav Mie[6] in 1900 introduced the first transmission line equations; they were for a cylindrical coaxial line, restricted frequency range, and a quite large "radius ratio". He showed, among other things, that no radiation passes into space from such a system.

The term "radius ratio" hereafter will be used to denote the ratio of the inner diameter of the outer tube to the outer diameter of the inner tube or wire.

In 1907 Abraham and Foppl[7] introduced the first complete solution of propagation along parallel wires of infinite conductivity in a medium of zero conductivity, beginning with the Maxwell equations. In 1909 Nicholson[8] derived formulas for the resistance and reactance of parallel wires for a wide range of frequency, but still limited to a large radius ratio. In the same year Russell[9] published limited discussions on effective resistance and inductance of "concentric mains", but applied it to audio frequencies.

Then some ten years passed with only very sparse work on skin effect and telephone carrier before Steinmetz[10] in 1919 introduced a complete formulation of circuit theory, concluding erroneously, that there is prodigious radiation from a transmission line. He stated that the radiation resistance of a one or two wire line is a thousand times greater than its ohmic resistance at high frequencies and suggested that as a consequence transmission lines would become increasingly useless as frequency was increased. In France, Ravut[11] published his "line equations" but they were limited still to quite restricted parameters.

In 1921 J. R. Carson of the A. T. and T. published three articles contributing vastly to the art. His first[12] introduced a more general solution for variable parameters, taking

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* Westinghouse Electric Corporation

[1] For all numbered references see bibliography beginning on page 47.

into account fluctuations of wire size, and spacing and splicing irregularities. His second[13] began with the Maxwell equations, assuming a small propagation factor. Current distribution within the wires was solved by consideration of the tangential magnetic force and the perpendicular magnetic induction at the surface of the wires. Formulas for a-c resistance and proximity effect were developed and equations were finally obtained for the impedance of the wires and the induction between them. His third[14] article concerned itself with the controversy then existing as to the radiation from transmission lines. He pointed out that Steinmetz and Poynting alike incorrectly agree that the magnetic field is propagated at right angles to the direction of propagation of the current, and with the velocity of light. Carson agreed with Heaviside that such is not true, and used Poincare's application of the Lorentz equations to show that the magnitude of radiation for a 1-foot wire spacing and a frequency of 1 megacycle is only

$$0.0051\left(1-\frac{\sin\frac{4\pi l}{\lambda}}{\frac{4\pi l}{\lambda}}\right)\text{ watts per ampere}$$

where l = length of the transmission line, λ = wavelength, and that since lines are frequently transposed, the radiation is probably much less. He hastened to show that this fact should not interfere with his previous calculations.

The first available literature describing practical tests at radio frequencies appears to be that of Murdoch, Hayes and Webster[15] in 1922, wherein characteristic impedance and attenuation measurements were made up to 135 kilocycles on an ordinary telephone wire as an MIT thesis problem. They concluded that, contrary to a few existing opinions, attenuation increases with frequency.

In 1932 Espenschied[16] pointed out the points of similarity between radio and wire transmission, making quantitative comparisons of power, fidelity, economy, interference, and singing. For his thesis at MIT Manneback[17] in 1923 wrote a comprehensive mathematical recapitulation of discussions on radiation from transmission lines, embracing the work of Maxwell, Abraham, Steinmetz, Carson, Mie and Hondros, and introducing the concept of "transient" and "stationary" radiation. He concluded that radiation has a negligible effect upon electromagnetic waves compared to the losses by heat dissipation in the wires. Also of interest are Carson's[18] treatment in 1924 of transients on long loaded lines, developing approximate but engineering formulas by consideration of the buildup envelope, and his account[19] of a recapitulatory mathematical investigation of guided waves and radiation from transmission lines, giving the detailed results without the corresponding detailed mathematical steps.

During the period 1920 to 1925 (roughly), authors of literature concerning the transmission line began to notice important differences in behavior between coaxial and open wire lines. Articles were written about either one type or the other, or each type was treated separately and distinctly; generalized articles became fewer and fewer. Investigation of the development of the two types will be completed separately henceforth, beginning with the then popular open line.

Open-Wire Line Development, 1925 to 1936

Open-wire lines may be classified by the number of conductors they employ. One-wire lines, of course, transmit their energy by means of a potential difference with respect to some other conductor, such as ground. They were very popular for antenna feeding at one time and are still used for that purpose, to some extent.

Two-wire lines include twisted pair and transposed and untransposed parallel wires, the last two having spacings between one-quarter inch and three feet. Most of the historical development of open lines in general has concerned itself with the question "Is radiation negligible?" and "Are losses (attenuation) low enough for practical use?" Two-wire lines carrying currents exactly out of phase in respective lines are supposed to have zero radiation compared to the single-wire line, but that will be reserved for further consideration in this section.

Three-wire lines have been used very little at radio frequencies and no literature was found concerning them. There are, of course, extensive treatments of them at audio and power frequencies for three-phase power transmission.

Four-wire lines are usually built with a rhombic (or square as a special case) configuration, the same principles governing the cancellation of radiation as for two-wire lines. It was considered at one time that the amount of radiation could be considerably reduced by the use of four or more (even in number) wires in proper juxtaposition.

Since authors quite often covered in the literature several of the above types in a single article, development of all types of open-wire lines will be considered together and chronologically:

Sallie P. Mead[20] in 1925 independently derived the equations for axial electric force and tangential magnetic force in non-magnetic tubular conductors with parallel return, and extended Carson's[13] equations in order to apply to the case of parallel tubular conductors. Carson[21] in 1926 derived mathematical formulas for wave propagation in overhead wires with ground return, taking into account the effect of inductive disturbances in and from neighboring transmitting systems. A current density correction is applied to Rudenberg's 1925 paper (a reference given in Carson's paper).

In 1927 Levin[22] published a simplification of parallel-wire theory without loss of accuracy, yet taking into account the finite conductivity of the earth.

Carson and Hoyt[23] in 1927 presented a formal theory for the general propagation of periodic currents over all types of transmission lines, including both grounded and all-metallic, taking into account the mathematical effect of the field of force on the conductor. Their paper also included a synthetic treatment of transmission line theory for physical grasp of the subject, and a discourse on the wave antenna. Carson[24] again in 1928 compared then known rigorous and approximate theories for transmission lines, pointing out the legitimacy of certain approximations for particular problems. The conflicts of Maxwell's equations with elementary engineering transmission theory were studied. Later the same year he[25] published an academic non-mathematical discussion of the problems, assumptions and new possible attacks of transmission line theory, reviewing its then existing status thoroughly.

In 1928 Green[26] published a very complete elementary theory of high-frequency energy transmission along feeders, in which the following well-known relations were derived. Basing his analyses on Heaviside's "Electromagnetic Theory" and defining "feeder" to mean either parallel-wire or concentric lines, he first treated the "infinitely long feeder of negligible resistance and leakage", concluding that for perfect conductivities there would be no attenuation or distortion. Under "relations of current, voltage, and energy in the wave sheet", he derived the relation that

$$\text{Total energy} = LI^2/2 + CE^2/2 \qquad (3)$$

and

$$LI^2 = CE^2, \qquad (4)$$

as it should be in an electromagnetic wave. Likewise,

$$R_o = E/I = \sqrt{L/C} = 1/CV \qquad (5)$$

and

$$V = 1/\sqrt{LC} = 3 \times 10^{10}/\sqrt{\mu K} \qquad (6)$$

wherein L = Inductance of feeder in henries per unit length, go and return.

C = Capacity between wires in farads per unit length.

μ = Magnetic permeability of dielectric between conductors.

K = Specific inductive capacity of same.

V = Velocity of propagation of electromagnetic disturbances along the feeder.

E = Input electromotive force.

I = Input current.

R_o = Surge (characteristic) impedance.

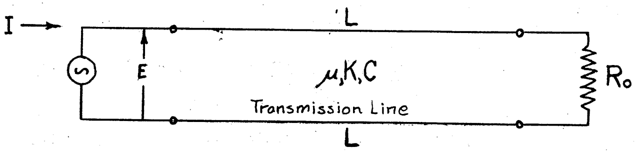


FIGURE 1—TRANSMISSION LINE TERMINATED IN PURE RESISTANCE.

Figure 1 - TRANSMISSION LINE TERMINATED IN PURE RESISTANCE.

The case of the "infinite feeder with sinusoidal E.M.F. applied to the input terminals" was considered briefly, and attention was then turned to the "steady conditions in a finite feeder with a sinusoidal E.M.F. at the input end and various terminal loads." In the case of a terminal load equal to a pure resistance, as in Figure 1, the proper termination for the feeder is given by (5) as $R_0 = \sqrt{L/C}$. It absorbs all energy, reflecting none back to the input end. Green took up the case of the "terminal load not equal to R_0" very extensively:

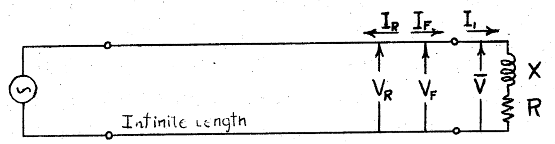


Figure 2 - TRANSMISSION LINE OF INFINITE LENGTH

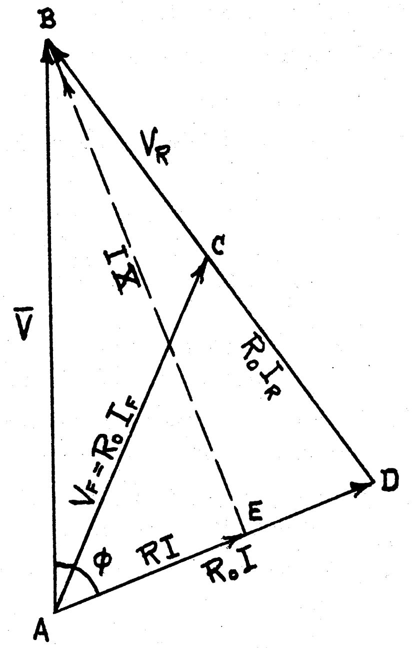


Figure 3 - LINE TERMINATION VECTOR DIAGRAM.

Considering the output end of a transmission line of infinite length, as shown in Figure 2, and letting

R = Resistance of the terminal load.

X = Reactance of the terminal load.

V and I = Amplitude of voltage and current in the load in the directions of Figure 2.

V_x and I_x = Resultant voltage and current at any point.

V_F and I_F = Amplitude of voltage and current in the reflected wave.

we have
$$R_0 = \frac{V_F}{I_F} = \frac{V_R}{I_R} \qquad (7)$$

and V_F and V_R are respectively in phase with I_F and I_R.

The vector sum of V_F and V_R at the load is

$$\overline{V} = \overline{V}_F + \overline{V}_R \, . \qquad (8)$$

Likewise the vector difference of currents at the load is

$$\overline{I} = \overline{I}_F - \overline{I}_R$$

Expressing (8) graphically in Figure 3, and in addition noting that

$$V_F = R_0 \cdot I_F,$$

so that V_F is in phase with I_F,

a current triangle ACD may be constructed with the voltage triangle ABC by extending CD = BC and in line with BC. Then $V_R = R_0 \, I_R$, so that V_R and I_R are in phase, also vectorially AD = AC − DC = $R_0(\overline{I}_F - \overline{I}_R) = R_0\overline{I}$ in phase and magnitude, and therefore ϕ is the phase lag of $\overline{I}$ behind $\overline{V}$. Then knowing R and X of Figure 2 and assuming a value of I, RI may be laid off on $R_0 I$ and the normal to it, EB, will be the reactance drop XI. Obviously, the entire vector diagram may be reproduced knowing R, X, V and R_0. Likewise, conditions existing at other points along the line at distance x from the load may be expressed graphically by rotating AC counter-clockwise about C through the angle $360x/\lambda$ degrees, (where λ is the wavelength in the feeder) at the same time rotating DCB in the opposite direction by the same angle, one being incident and the other reflected.

The foregoing applies to a single incident wave and a single reflected wave; Mr. Green included (in his own figure 6) an interesting chart of vector diagrams for each of several points spaced an eighth wavelength, showing the distribution of V, I ϕ, and equivalent impedance along the line. Shorter discussions on the vector diagrams for the special case of the pure resistance load for open and shorted terminals, and a note on the effect of attenuation were also included.

Nancarrow[27] the same year published a different point of view of the same problem, using the regular transmission line equations. He used (and derived in an appendix) the expressions:

$$V = I_r\ (Z_o \sinh P_x + Z_a \cosh P_x) \qquad (10)$$

and $$I = I_r\ (Z_a \sinh P_x + Z_o \cosh P_x)/Z_o, \qquad (11)$$

wherein V = Potential difference at a distance x from load

I = Current at a distance X from the load

Z_a = Impedance of the load

Z_o = Iterative or surge impedance of the line

P = Transmission (propagation) constant of the line

I_r = Current at the load.

For the audio frequencies the characteristic impedance is given by

$$Z_o = \sqrt{\frac{R + j\omega L}{S + j\omega C}} \qquad (12)$$

wherein R, L, S, and C are the resistance, inductance, leakance, and capacitance, respectively, per unit length of line, and ω is, of course, $2\pi f$.

But at radio frequencies, R and S are negligible compared with ωL and ωC, so that

$$Z_o = \sqrt{L/C} = \text{pure resistance.} \qquad (5a)$$

Then $$P = \sqrt{(R + j\omega L)\ (S + j\omega C)} = \alpha + j\beta \qquad (13)$$

where α is the attenuation constant and β is the "velocity" (phase) constant. The value of α usually must be obtained by direct measurement, but β is simply

$$\beta = \omega/v = 2\pi f/v = 2\pi/\lambda, \qquad (14)$$

where v is the velocity of propagation along the line, (essentially that of light) and λ is the wavelength in the conductor. Then from the equation

$$Z = \frac{V}{I} = Z_o \frac{Z_o \sinh Px + Z_a \cosh Px}{Z_a \sinh Px + Z_o \cosh Px} \qquad (15)$$

for negligible α we get

$$Z = Z_o \frac{(Z_o + Z_a)e^{j\frac{2\pi x}{\lambda}} + (Z_a - Z_o)e^{j\frac{2\pi x}{\lambda}}}{(Z_o + Z_a)e^{j\frac{2\pi x}{\lambda}} - (Z_a - Z_o)e^{j\frac{2\pi x}{\lambda}}}, \qquad (16)$$

x being the distance from the load to the disturbance.

Nancarrow constructed another vector diagram analogous to Figure 3, showing that the maximum value of Z is

$$Z_R = Z_o \frac{(Z_o + Z_a) + (Z_a - Z_o)}{(Z_o + Z_a) - (Z_a - Z_o)}, \qquad (17)$$

and the minimum value is

$$Z_R = Z_o \frac{(Z_o + Z_a) - (Z_a - Z_o)}{(Z_o + Z_a) + (Z_a - Z_o)}, \qquad (18)$$

from which $Z_o = \sqrt{(Z_{R\ max})\ (Z_{R\ min})}$. (19)

He also pointed out that a line terminated by a resistance (or a short or open circuit) behaves as a resistance at points one quarter of a wave-length from the load. The expressions for Z_o as a function of radius ratio (for the concenctric line) and spacing/wire-diameter ratio (for the open-wire line) was derived and curves given, but a more complete set is reproduced in those respective sections of this report.

Moser[28] in 1928 gave a short treatise on all types of feeders, stating that heating was the principal limitation to the length of a transmission line at radio frequencies. He showed that the efficiencies of 100 and 1000 meter lines could be 97% and 77% respectively. Everitt and Byrne[29] published in 1929 one of the first defenses of the single-wire transmission line, showing that it is efficient only when terminated correctly. Experiment and theory were invoked to show that feeder radiation "will be small and the feeder will act efficiently." Bird[30] in 1930 gave a method for obtaining the correct terminal impedance from three current readings on the existing feeder, based upon equation (19), and in the same issue Green[31] gave vector diagrams relating the voltage and current at any other point with the voltage and current at a given point, derived in the same manner that he derived figure 3.

Roosenstein[32] in his "transmission of high frequency electrical wave energy" of 1930, gave a theoretical and practical investigation of "Lecher wire" transmission characteristics. Transmission efficiency, output impedance, and methods for obtaining Z, A, and symmetry were discussed. Tanesesco[33] in 1930 gave a grapho-analytic method of obtaining the variation of effective impedance of a transmission line at high frequency by super-position of the direct and reflected waves, not unlike Green's[26].

In a 1931 article primarily concerned with "beam arrays", T. Walmsley[34] gave a qualitative comparison of the relative merits of the concentric and parallel-wire lines, stating that open-wire lines have no more loss than concentric, and classifying the losses. He also suggested the use of stranded conductors, larger cross section, good balance, close spacing, good insulation, and proper impedance match. A second article[35] dealt mathematically with unbalanced current in parallel lines, applicable to lines many wavelengths long. He stated that the amount of unbalance current in the open-wire depends upon the type of array it feeds, and that the short lines have almost as much radiation resistance as the long ones, - - that is, the radiation resistance does not limit the length of a transmission line, the greatest loss being high-frequency ohmic. Uda[36], the same year, published in English a description and mathematical examination of an open-wire line having a two-centimeter conductor spacing and operating at a two-meter wavelength.

Roosenstein[37] again in June 1931 described his damping measurements on high-frequency lines at the Telefunken Laboratories in Nauen, based upon the "sharpness" of voltage nodes and the characteristic impedance, and discussed the matching of aerials and feeders. He used the formula

$$X = -Z_0 \cot \frac{2\pi l}{\lambda} \qquad (20)$$

to get Z_0 by measuring X (as a capacitive reactance) for an eighth wavelength of line. His calculated value of 500 ohms agreed fairly well with the measured 527 ohms at twenty-seven meters wavelength. He also suggested the use of quarter-wave lines to match an antenna to its feeders, using the trigonometric loss-free variation of equations (10,11):

$$E_1 = E_2 \cos \frac{2\pi l}{\lambda} + jI_2Z_0 \sin \frac{2\pi l}{\lambda} \qquad (21)$$

$$I_1 = I_2 \cos \frac{2\pi l}{\lambda} + j\frac{E_2}{Z_0} \sin \frac{2\pi l}{\lambda} \qquad (22)$$

which for a quarter-wave line becomes

$$E_1 = jI_2Z_0 \qquad (21a)$$

and $$I_1 = j\frac{E_2}{Z_0} \qquad (22a)$$

When loaded at the free end (subscript 2) with R_2, then $E_2 = I_2R_2$ and at the input end

$$R_1 = E_1/I_1 = Z_0^2/R_2, \text{ or}$$

$$Z_0 = \sqrt{R_1R_2} \qquad (23)$$

which gives the characteristic impedance of a quarter-wave line to be operated as a matching transformer between impedances R_1 and R_2. Together with Baumann[38] a few months later he published further results of his damping measurement investigation. In the same year his mathematical paper[39] on the conduction of high-frequency oscillatory energy, taking dissymmetry into account, appeared, and was immediately followed by still another[40] concerned with the symmetry of two and four-wire transmission lines. In the last paper he showed that not only a necessary but sufficient condition that radiation be small, is that the apparatus associated with it be symmetrical to ground. Means for testing and correction of unsymmetrical radiation from feeders was included.

During 1931 two Russian articles appeared. One by Tatrinov[41] merely described how to get the characteristic impedance of a line by nodal voltage measurements and equation (19). The other by Uger[42] derived formulas based upon the Poynting[2] method for obtaining power, radiation resistance, and distribution function as functions of $2\pi l/\lambda$ for the one-wire feeder.

A book by McIlwain and Brainerd[43], "High-Frequency Alternating Currents", also appeared in 1931 and includes a chapter on transmission lines.

Creamer[44] in 1932 published a method intended for telephone workmen, obtaining the equation for the propagation constant (equation 13) without resorting to calculus. Posthumus[45] the same year showed how to insert one correcting impedance into the line to eliminate standing waves on it.

In July 1932 Sterba and Feldman[46] published a very comprehensive treatise comparing open-wire and concentric lines as a basis for later articles leading up to the construction of the first long-distance coaxial cable from New

York to Philadelphia. The portions of the article dealing with coaxial cables and their relative merit are discussed on page 13. The elementary relations (5, 12 and 19) are stated, and a discussion is launched on the properties and limitations of open-wire lines:

With sufficient accuracy, the characteristic impedance Z_o of a two-wire open line is

$$Z_o = 276 \log_{10} \frac{2D}{d} \text{ ohms.} \qquad (23.5)$$

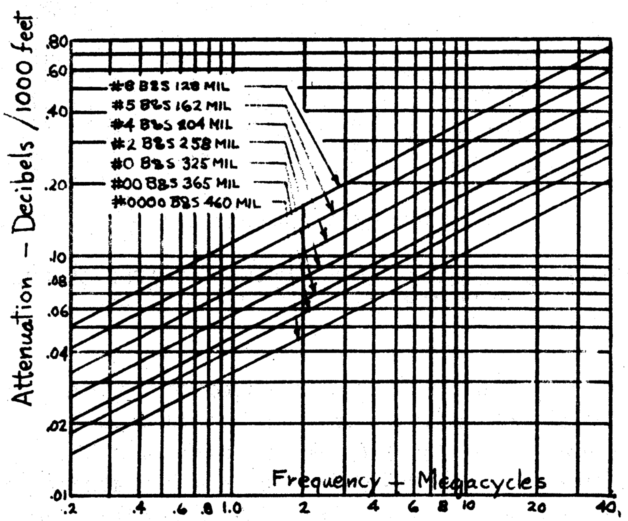


Figure 4 - CALCULATED ATTENUATION BY COPPER LOSSES IN TWO WIRE 600 OHM LINES

The attenuation is shown graphically in Figure 4. Here D is the conductor spacing and d the wire diameter. Figure 5 shows the proximity effect to be negligible for the ordinarily used construction; references are given from which the curve was obtained.

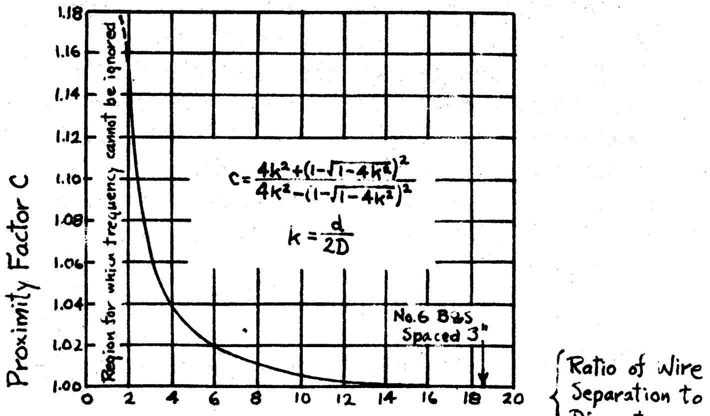


Figure 5 - INCREASE IN COPPER LOSS IN BALANCED TWO WIRE LINE DUE TO PROXIMITY EFFECT.

When the line spacing is less than one-tenth wavelength and its length is more than twenty times the spacing, the power radiated by a two-wire line terminated in its characteristic impedance is given approximately by

$$\frac{P}{I^2} = 160 \left(\frac{\pi D}{\lambda}\right)^2 \text{ watts per square ampere} \qquad (24)$$

in which I is the r-m-s current in the line and D/λ is the line spacing in wavelengths; radiated power does not involve the line length. The radiated power evidently is about twice that radiated by a doublet antenna of length equal to the line spacing (from doublet radiation formulas)! Considering line terminations of correct characteristic impedance and length D/λ each, the total radiation will be about twice that given by P/I^2 of (24) and is plotted in Figure 6. Therefore radiated power is negligible compared to the transmitted power.

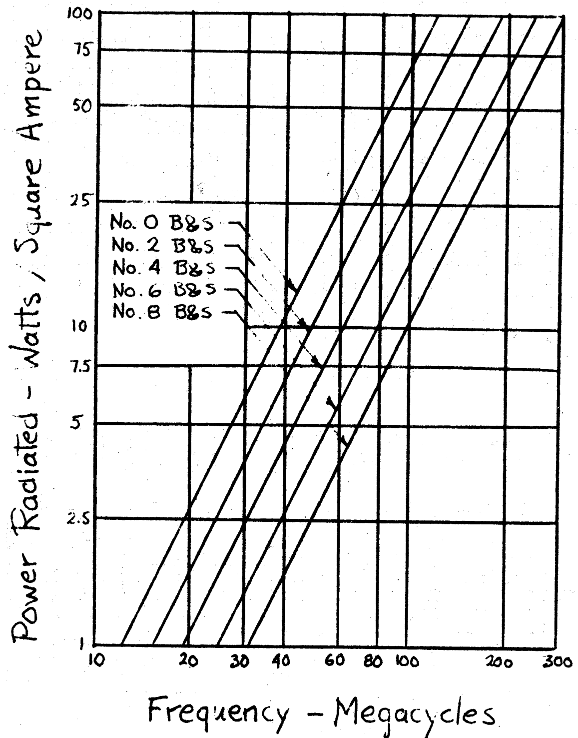


Figure 6 - CALCULATED POWER RADIATED BY 600 OHM OPEN LINES TERMINATED IN THEIR CHARACTERISTIC IMPEDANCE.

Unbalance current in a line may be considered the residual after the balanced components are removed, in which case the two wires act in parallel as a single-wire antenna. Ignoring "mutual interactions" between lines, the unbalanced power radiated is approximately given by

$$\frac{P}{I^2} = 30 \left[0.5772 + ln\,(2L) - \sin^2(L)\ \left(1 - \frac{\sin H}{H}\right) - Ci(2L) - 2Ci(H) + Ci(\sqrt{L^2 + H^2} - L) + Ci(\sqrt{L^2 + H^2} + L) \right] \text{ watts per ampere}^2, \qquad (25)$$

in which I = r-m-s value of maximum current

$H = 4\pi h/\lambda$

h = height of wires above ground

$L = 2\pi l/\lambda$, and

l = length of line

with lengths being in consistent units; the result is plotted in Figure 7. Examination of the curves shows that a thirty-percent unbalance radiates about the same amount of power as the balanced currents in the line. Sterba and Feldman state that it is their experience that unbalance radiation is very much larger than this.

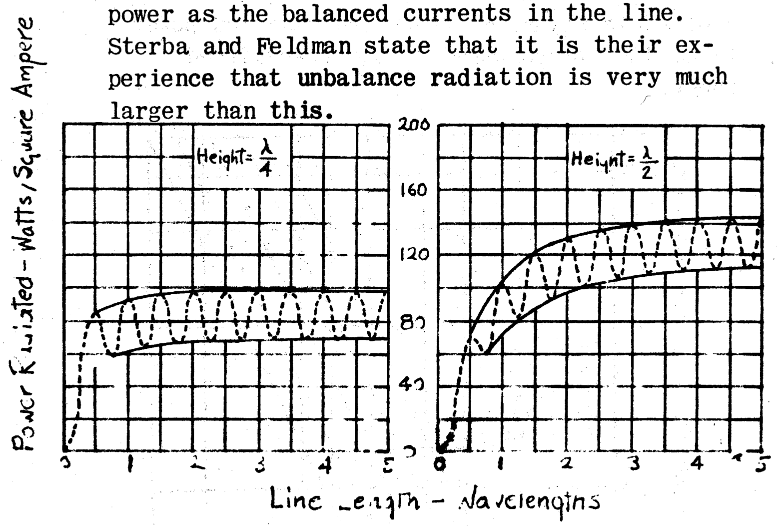


Figure 7 - APPROXIMATE POWER RADIATED BY AN UNBALANCE CURRENT OF ONE AMPERE.

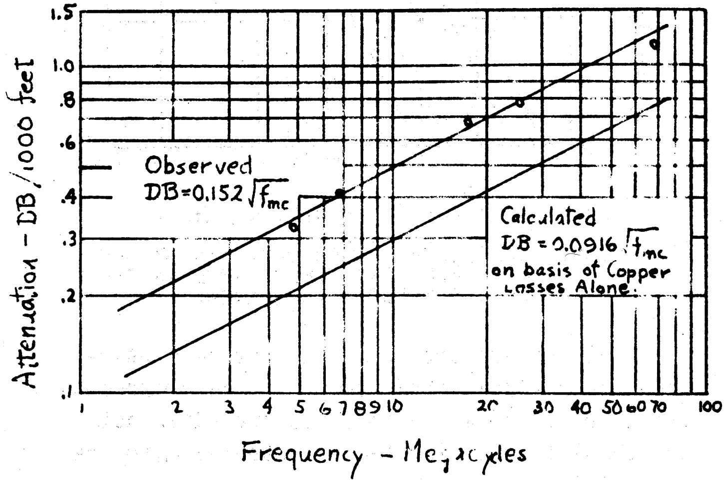


Figure 8 - ATTENUATION IN A 600 OHM OPEN LINE OF 0.162" COPPER CONDUCTORS.

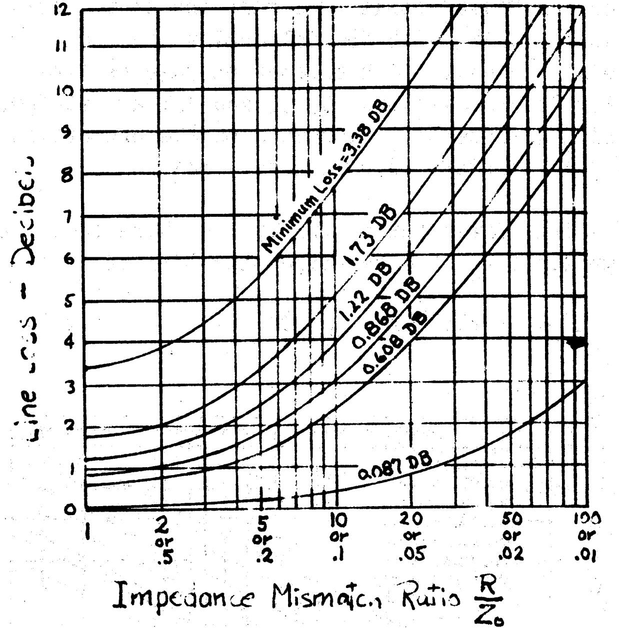


Figure 9 - LINE LOSS FOR SEVERAL VALUES OF LOSS AT IMPERFECT IMPEDANCE MATCH.

Radiation resistance “according to Carson's[19] investigation” is a term to be added to the impedance of the line at its terminals or points of discontinuity; it does not appear in the propagation constant. “On this basis the power radiated. . . is small compared to the transmitted power”. Figure 8 shows the poor agreement between observed impedances and those calculated on the basis of 1830 emu copper resistivity; an explanation is offered for the discrepancy. Figure 9 shows the effect of mismatching impedances, assuming a line length of an integral number of quarter wavelengths, avoiding complex impedances. Schemes presented for matching impedances will be found on page 42 of this paper. Location of the current nodes and poles is facilitated by the use of such instruments as shown in Figures 10 and 11. Instrument A of Figure 11 gives no indication of the line unbalance, whereas instrument B does. Measurements of average current at each end of the line give loss in decibels from the definition

$$\text{Loss} = 20 \log_{10} \frac{I_s}{I_r} \text{ decibels} \qquad (26)$$

and since

$$\frac{I_s}{I_r} = e^{\frac{Rl}{2Z_o}} \qquad (27)$$

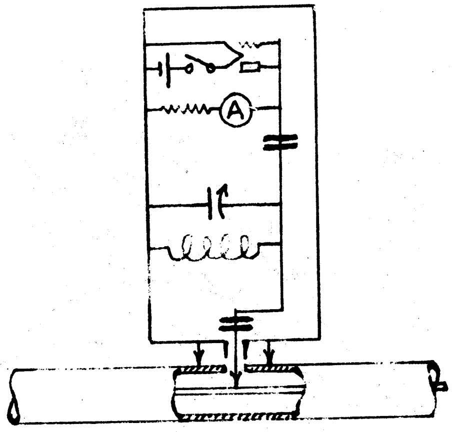


Figure 10 - VOLTAGE MEASUREMENT ON CONCENTRIC LINES.

we find

$$\text{Loss} = 4.343\,\frac{R}{Z_o},\ \text{db/unit length of line} \qquad (28)$$

which gives the resistance per unit length of line, R, with accuracy as good as Z_o. Care must be taken to extract only a very small portion of the power from the line during measurements in order to avoid reflection.

A resistance-substitution method is described for obtaining the r-f resistance and attenuation constant of a quarter-wave-multiple line either shorted or open at the far end, using the equations

$$Z_1 = Z_o \tanh\left[\alpha\frac{n\lambda}{4}\right]$$

$$Z_2 = Z_o \coth\left[\alpha\frac{n\lambda}{4}\right] \qquad (29)$$

$$\text{and} \quad \alpha = 4.34\left[\frac{R}{Z_o} + GZ_o\right] \text{ db per foot} \qquad (30)$$

to obtain with 1.5% maximum error

$$Z_1 = \frac{Rn\lambda}{8}\left[1 + \frac{GZ_o^2}{R}\right] \text{ ohms} \qquad (31)$$

for n even and shorted termination, or n odd and open termination, or the alternative

$$Z_2 = \frac{8Z_o^2}{Rn\lambda}\left[1 - \frac{GZ_o^2}{R}\right] \text{ ohms} \qquad (32)$$

for n odd and shorted termination, or n even and open termination, where

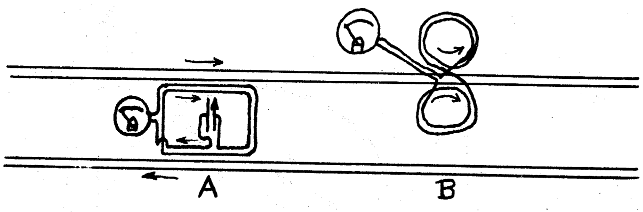


Figure 11
CURRENT MEASUREMENT ON OPEN-WIRE LINES.

n = length in integral number of wavelengths,

R = resistance in ohms per foot,

and G = conductance in mhos per foot.

When G is negligible, the impedance measurement gives resistance directly. The method brought to light the fact that small irregularities such as elbows, bends, etc., (for concentric lines) may cause a variation of impedance of five to ten percent, whereas the characteristic is very smooth for the carefully constructed line.

Feldman[47] in an article appearing in the Bell Laboratories Record in the same year, gave an excellent comparison between power and radio-frequency lines covering non-mathematically much of the ground of the previous article[46]. The same photographs appeared.

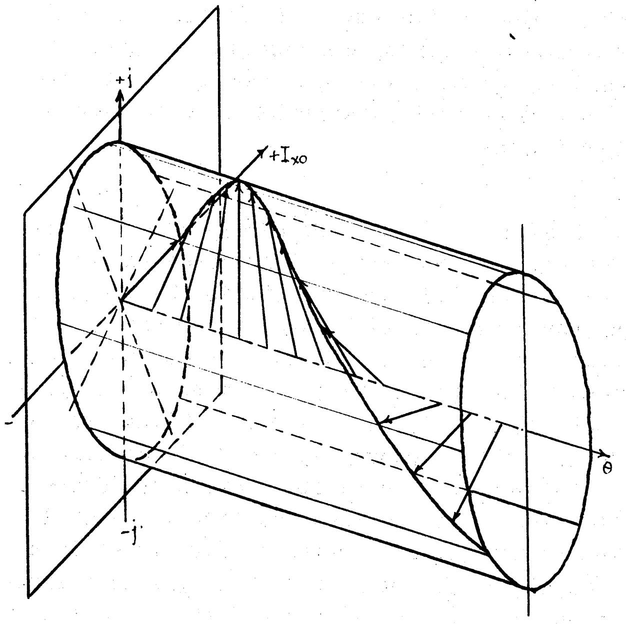


Figure 12 - ELLIPTICAL CYLINDER DIAGRAM OF THE TRANSMISSION LINE.

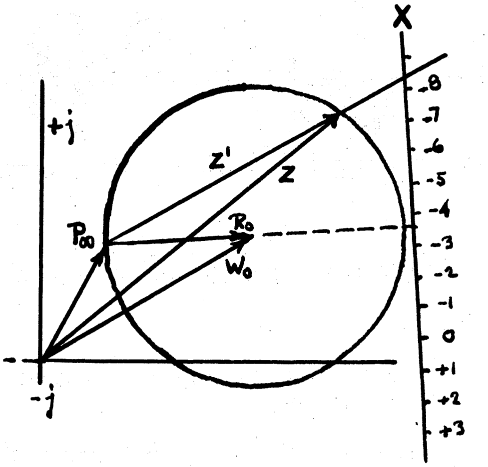


Figure 13 - IMPEDANCE AT INPUT AND LOCATION OF LOAD IMPEDANCE.

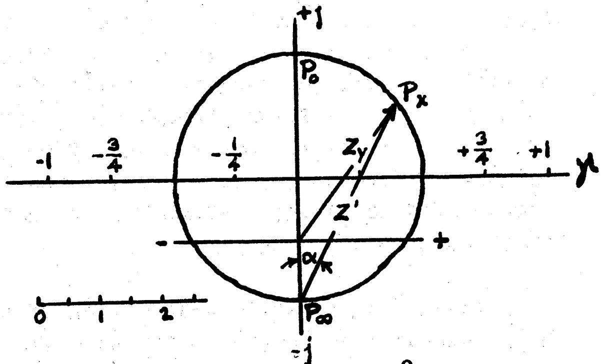


Figure 15 - INPUT IMPEDANCE FOR 45° LENGTH OF LINE AND VARIABLE RESISTIVE COMPONENT.

Fischer[48] also published an article in 1932 concerning active and reactive power in r-f transmission lines, and Bahnemann[49] gave theoretical equations and results of experiments on energy transmission at very high frequencies.

Hans Roder[50], then of the General Electric Company, in February 1933 reviewed the art of transmission line representation by means of vector diagrams, with especial emphasis upon practical graphical solution of common problems. After preliminary statement of formulas such as (10, 11, 12, 5 and 14), Figure 12 is derived. The vectors represent either current or voltage and the elliptic cylinder degenerates into a circle when the line is matched. The author gives a rather incomplete explanation of the use of Figure 13 to get line input impedance Z in terms of the distance of the load impedance Z_1 along the transmission line, but gives several references. Figure 14 gives the input impedance of a line of variable

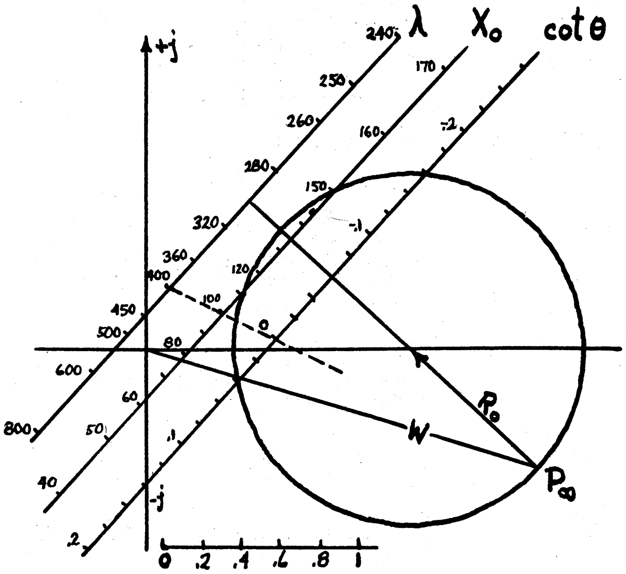


Figure 14
INPUT IMPEDANCE AND ELECTRICAL LINE-LENGTH

length which at its end is loaded by an impedance $Z_0(2 - j0.6)$, and Figure 15 is the solution of the same impedance but with "variable resistive component", electrical line length of 45°, R_{xo} variable, $X_{xo} = Z_0/2$. Only positive values of γ have meaning.

A similar article by Hikosaburo[51] in the same issue explained in particular the double-hump phenomenon. He also gave in detail the use of the three-dimensional ellipse diagram giving reciprocal bridge current (the bridge across a length of line) as a vector, as a function of distance along the line.

Strutt[52] in March 1933 reviewed the five then existing methods of measuring attenuation of lines and added a sixth. His method consisted of varying the frequency of input and use of readings of high-impedance voltmeters bridging each end of the line in the formula:

$$\frac{E_1}{E_2} = \cosh \alpha l \qquad (33)$$

for the frequencies at which the line is an even number of half-wavelengths long.

Another book available as reference material on radio-frequency transmission lines is

M. G. Malti's "Electric Circuit Analysis"[53], which takes up the subject as its Chapter 19.

A very complete discussion of crosstalk on open-wire lines was made public by Chapman[54] in April 1934. His article covered all types of open-wire lines and crosstalk mitigation in general. A purely mathematical investigation by Wise[55] of propagation of high-frequency currents in ground-return circuits, correcting Carson's formula, appeared in the same month.

C. W. Finnigan[56] in 1934 chose as a thesis problem the determination of the efficiency of two-wire feeders. He found the loss of a line to be given by

$$\left[\frac{Z_r}{Z_o} + \frac{Z_o}{Z_r}\right] \frac{RL}{2Z_o}, \qquad (34)$$

in which Z_o = the characteristic impedance,
R = high-frequency loop resistance per unit length of line,
Z_r = impedance at receiving end,
and L = length of line.

For instance a 600-ohm line working into an eighty-ohm load will have over seven times the loss that it would have had working matched. The impedance at the receiving end of a line one-quarter wavelength long is found to be

$$Z_r = Z_o \frac{I_x}{I_r}, \qquad (35)$$

in which I_x = current at the input end,
and I_r = current at receiving end.

This relation is very practically used to measure impedances of other circuits, by merely employing a quarter wavelength of line of known characteristic impedance. Efficiency of a line, defined as the ratio of output to input power, may be then obtained from current and impedance readings. Finnigan's work assumed no inductive nor capacitive components in the impedances used.

Gerth[57] in December 1934 published an article describing a method for the prevention of radiation from lines by the insertion of special coupling coils between the wires, blocking in-phase currents and passing those out of phase.

Clark[58] in April 1935 published another "background" article (on the New York-Philadelphia coaxial cable) concerned with the relative merits of coaxial and open-wire circuits for wide-band transmission. Attenuation may be as low as one decibel for the open line compared to six for the small coaxial at one megacycle. But crosstalk between wires and interference to and from other systems make open lines undesirable. The use of the latter at the present state of the art is permissable for television but not for carrier telephony.

Hunter and Booth[59] in the same month quantitatively investigated the mutual inductance between two adjacent r-f circuits. Because of skin and proximity effects, the effective mutual inductance is a complex quantity whose real and imaginary coefficients change with frequency. Measurements and computations were carried to one megacycle, in anticipation of the new carrier-cable requirements. A second paper by Hoyt and Mead[60] covered the same territory but is mathematically more complete. Physical theory, mathematical theory and calculations, and production and properties of electric field intensities and voltages were reviewed.

L. M. Craft[61] in December 1935 described the use of the quarter-wave multiple transmission line (transformer) between antenna and exciter, and gave a convenient recapitulation of popular line theory. The following formulas are useful:

Conductor resistance

$$R = \sqrt{\frac{\rho f}{r}} \times 10^{-9} \text{ ohms/cm.} \qquad (36)$$

Proximity correction

$$C = \frac{D}{\sqrt{D^2 - 4r^2}}. \qquad (37)$$

In these ρ = resistivity in emu,
f = frequency in cps,
r = conductor radius in cm.,
and D = separation, center to center, in cm.

Characteristic Impedance

$$Z_o = \sqrt{\frac{Z}{Y}} \text{ ohms} \qquad (38)$$

Propagation Constant

$$\gamma = \sqrt{ZY} \text{ nepers} \tag{39}$$

Impedance per unit length

$$Z = R + j\omega L \text{ ohms} \tag{40}$$

Admittance per unit length

$$Y = G + j\omega C \text{ mhos} = j\omega C \text{ mhos} \tag{41}$$

for negligible leakage conductance. For a low-loss line R is negligible compared to ωL and the angle of Z is approximately $\pi/2$, so that

$$Z = \omega L \angle \frac{\pi}{2} - \frac{R}{\omega L} \text{ ohms } \angle \text{radians} \tag{42}$$

Replacing the angle by its tangent (for small angles),

$$Z_o = \sqrt{\frac{L}{C}} \angle \frac{-R}{\omega L} \text{ ohms } \angle \text{radians} \tag{43}$$

which is almost a pure resistance. Also,

$$\gamma = \omega\sqrt{LC} \angle \frac{\pi}{2} - \frac{R}{2\omega L} \text{ nepers } \angle \text{radians} \tag{44}$$

$$\gamma = \frac{R}{2Z_0} + j\omega\sqrt{LC} \text{ nepers} \tag{45}$$

and if the line is terminated in its characteristic impedance Z_o, then

$$\gamma = \frac{R}{2Z_o} \text{ nepers} \tag{46}$$

$$= 4.34 \frac{R}{Z_o} \text{ decibels.} \tag{47}$$

Phase shift

$$\beta = \omega\sqrt{LC} \text{ radians.} \tag{48}$$

Velocity of propagation

$$V = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}}. \tag{49}$$

Neglecting proximity and the effect of dielectric constant upon capacitance,

$$Z_o = \sqrt{\frac{L}{C}} = 276 \log_{10} \frac{D}{r} \text{ ohms.} \tag{50}$$

For the coaxial line,

$$Z_0 = 138 \log_{10} \frac{R_D}{r}, \tag{51}$$

in which R_D = the inner radius of the outer conductor. The hyperbolic line equations are

Sending end current

$$I_s = I_r \cosh \gamma l + \frac{E_r}{Z_o} \sinh \gamma l \tag{52}$$

Sending end emf

$$E_s = E_r \cosh \gamma l + I_r Z_o \sinh \gamma l \tag{53}$$

where I_r = load current

E_r = load voltage

l = line length.

Terminating the line of characteristic impedance R_o (pure resistance) by the complex impedance $Z_r = E_r/I_r$,

Input power

$$P_i = aR_oI_r^2 \left[\cosh 2\alpha l + \frac{a^2+b^2+1}{2a} \sinh 2\alpha l\right] \tag{54}$$

in which b = 0, if the line is terminated in a resistance. If it is terminated in its characteristic impedance, then

Line Loss $L = 8.68\alpha l$ decibels. (55)

Output power $P_o = I_r^2R_r = aI_r^2R_o$, (56)

where

$$a = \frac{R_r}{R_o},$$

$$b = \frac{X_r}{R_o},$$

and R_r and X_r are the load resistance and reactance respectively. Then the line loss

$$L = 10 \log_{10} \frac{P_i}{P_o}, \text{ decibels} \tag{57}$$

and the efficiency

$$E = 100 \frac{P_o}{P_i} \tag{58}$$

McNary[62] edited the Engineering Handbook of the National Association of Broadcasters, in which there appear the simple relations between conductor size, spacing and impedance, plotted in Figure 16. A. Rosen[63] in January 1936 gave a very brief procedure wherein the primary constants of a uniform line may be obtained from the propagation constants "without recourse to trigonometric tables, and suitable for use with a slide rule, and valid for all frequencies and loaded and unloaded lines".

Weber and Kulman[64] in March published a graphical and mathematical proof of the theory of transient phenomena. The solution was obtained in the form of a succession of traveling damped sine waves "whose reflections diminish in order until steady state is reached".

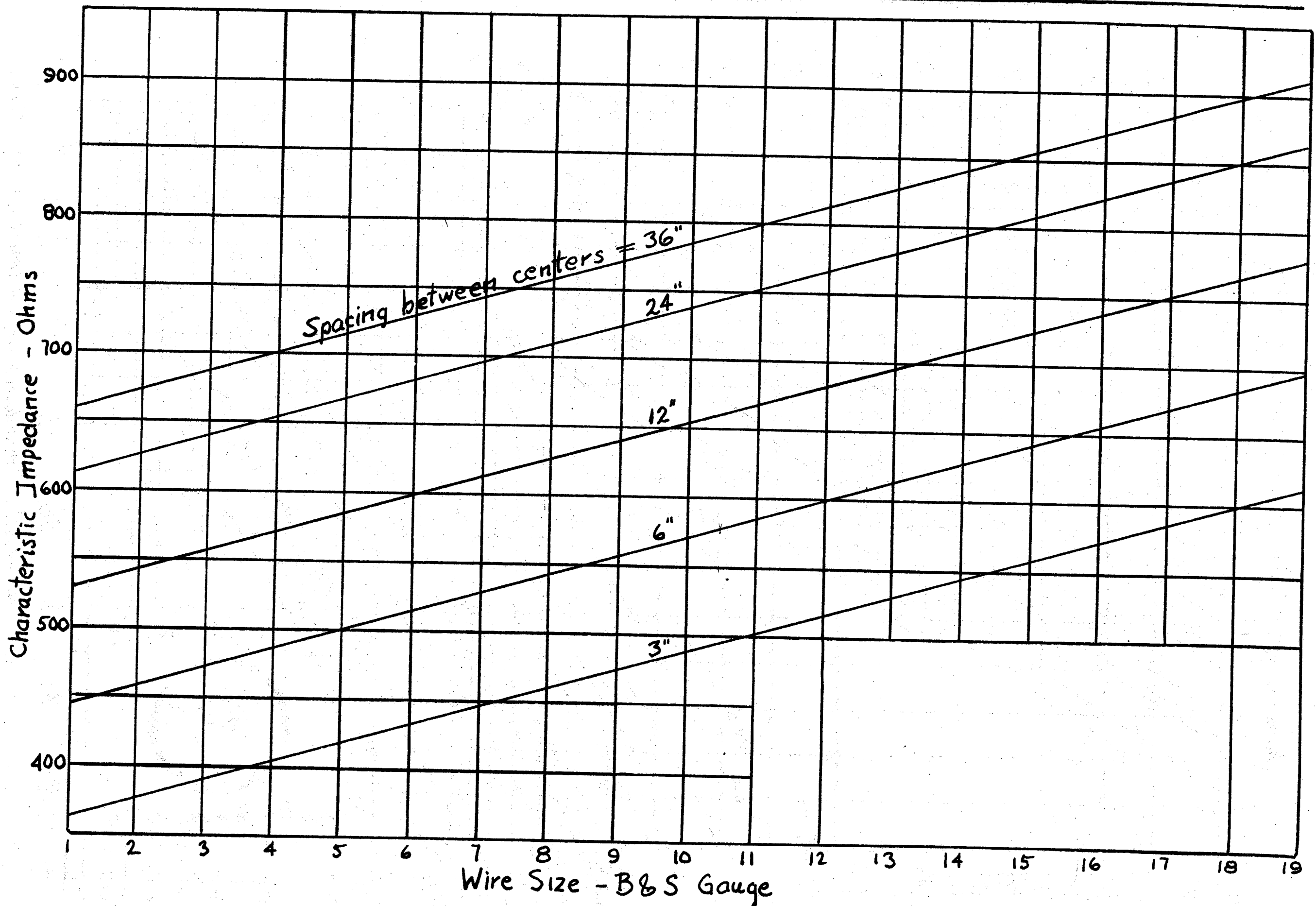


Figure 16 - CHARACTERISTIC IMPEDANCE OF TWO WIRE OPEN LINE.

Gordon and Sperling[65] briefly reviewed a simplified line theory in April 1936, developing equations (38, 39, 40, 41 and 43) and using (50) to plot a chart (Figure 17) in practical form. In Sperling's part II[66] the causes of attenuation are briefly discussed and curves are plotted for equations (36) and (47). The efficiency curve of Figure 18 is also included from the equation

$$\text{Efficiency} = (100 - \frac{2L}{\lambda}). \qquad (59)$$

Teachman[67] in December 1936 published a chart giving the impedance of open-wire lines of the range from 330 to 880 ohms in terms of the wire size (No. 8 to 16) and spacing (1" to 20") and pointed out that equation (50) does not hold for a spacing less than ten times the diameter of the conductors. His chart is practical, but nearly all its information is given by Gordon and Sperling in Figure 17.

Schelkunoff[68] in October 1936 introduced a modified Sommerfeld's integral equation and gave applications, specifically covering radiation from parallel wires.

Cable Development, 1925 to 1936

Although the word "cable" includes any grouping of two or more into a unit, regardless of whether a metal sheath is used, it is advisable for our present purpose to limit the meaning to one or more shielded conductors. Comparatively little work has been done in the multi-conductor field at high frequencies, but such articles have been included chronologically amid the work on the coaxial line.

Carson[69] in 1929 published a short article in which he derived the formulas for an underground wire close to the surface, concluding that return impedance is independent of wire depth.

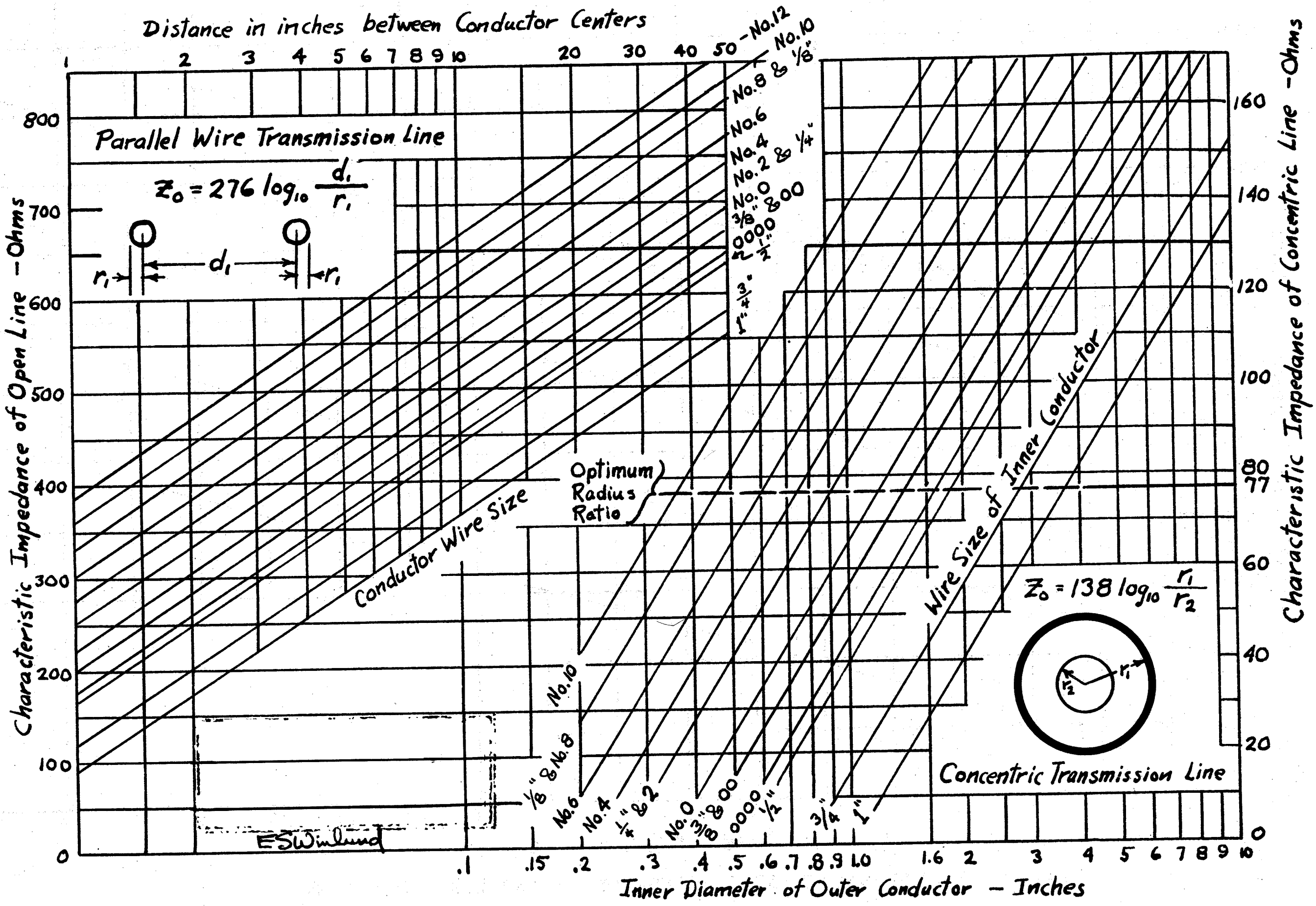


Figure 17 - CHARACTERISTIC IMPEDANCE OF OPEN WIRE AND CONCENTRIC LINES.

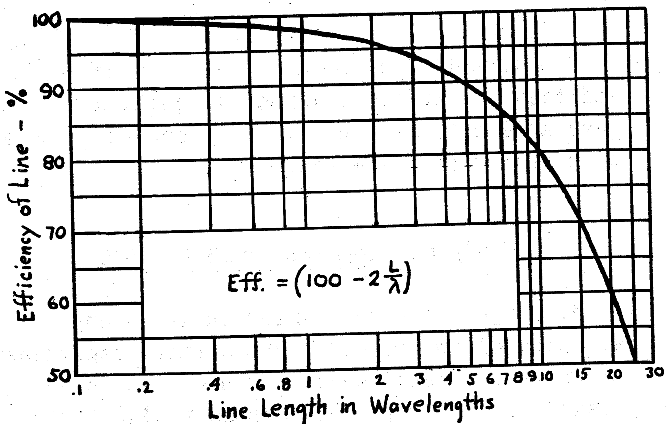


Figure 18 - EFFICIENCY OF CONCENTRIC OR OPEN WIRE TRANSMISSION LINE

Takagishi, Iso, and Kawazoe[70] in the same year found efficiencies of 82 to 86% for a coaxial line fifty-five meters long supplying an antenna array. The 82% was for a #3 B. & S. copper wire in a galvanized iron pipe of 3.8 centimeters inside diameter, and the 86% was for the same inner conductor in a copper pipe of 4.0 centimeters inside diameter.

The next work covering the subject of coaxial lines as such appears in the previously recorded work of Sterba and Feldman[46] in 1932. Equations (12, 13, and 50) are stated, and in addition it is pointed out that the propagation constant

$$P = \alpha + j\beta = \frac{R}{2Z_o} + \frac{GZ_o}{2} + \frac{j2\pi}{\lambda}, \qquad (60)$$

where R is the resistance and G the conductance, each per unit length and at wavelength λ. The capacity per unit length is given as

$$C = \frac{1}{2 \ln\frac{b}{a}} \text{ esu}, \qquad (61)$$

the inductance as

$$L = 2 \ln \frac{b}{a} \text{ emu} \qquad (62)$$

and the impedance as

$$Z_o = 138 \log_{10} \frac{b}{a} \text{ ohms}, \qquad (63)$$

where a = outer radius of inner conductor, and
b = inner radius of outer conductor.

The high-frequency resistance obtained by Russell* is

$$R = \sqrt{\rho\mu f}\left(\frac{1}{a} + \frac{1}{b}\right) 10^{-9} \text{ ohms per cm.,} \qquad (64)$$

where ρ = resistivity in emu (=1730 for pure copper),
μ = magnetic permeability,
and f = frequency in cps.

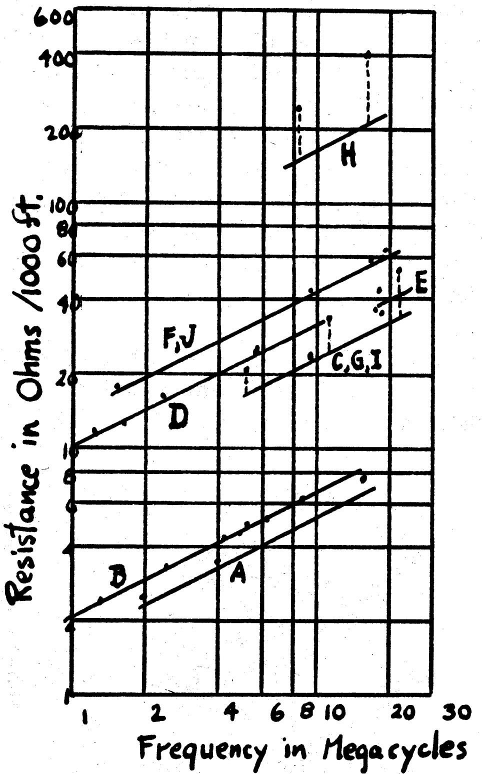


Figure 19 - RADIO FREQUENCY RESISTANCE - CONCENTRIC LINE.

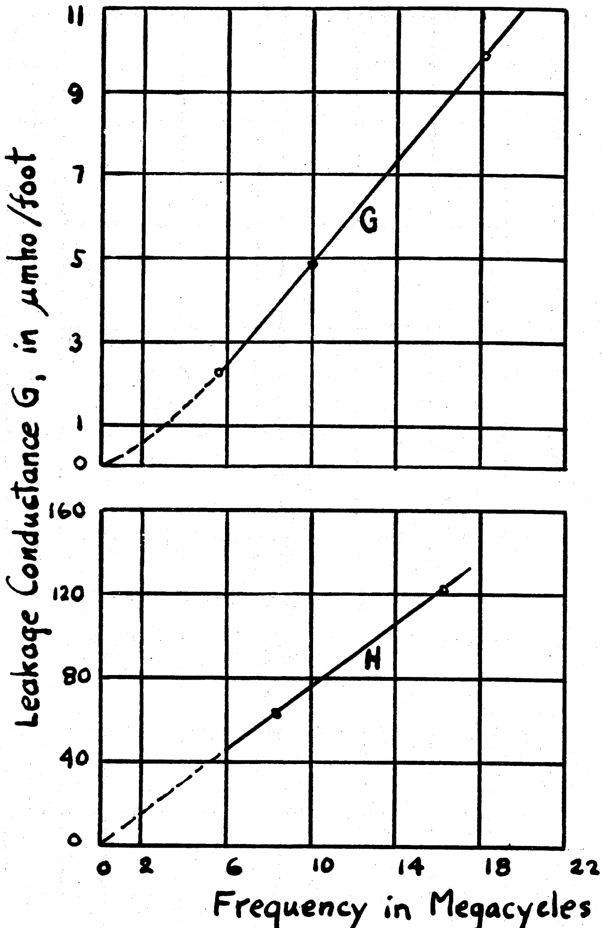


Figure 20 - CALCULATED CONCENTRIC LINE RADIO FREQUENCY LEAKAGE CONDUCTANCE.

This is a more accurate form than equation (36). The straight curves of Figure 19 represent this equation and the dots are observed data on various lines whose dimensions are given in the article. The equation agrees very well except for G and H; G was completely filled with porcelain insulators and H was completely rubber-insulated. Figure 20 shows the tremendous effect found, and of course expected, upon the leakage conductance. By minimizing the expression for the real part of the propagation constant, namely,

$$\alpha = \frac{\sqrt{\rho\mu f}}{276}\,\frac{\left(\frac{1}{a} + \frac{1}{b}\right)}{\log_{10}\frac{b}{a}}\,10^{-9}. \qquad (65)$$

the optimum radius ratio is found to be

$$\frac{b}{a} = 3.6 \qquad (66)$$

Equation (65) is shown graphically in Figure 21. The numerical loss in copper concentric lines of optimum radius ratio 3.6, is

$$\frac{0.128\sqrt{f_{mc}}}{b_{in.}} \text{ db per 1000 ft.,} \qquad (67)$$

which is plotted for f_{mc} = 20 mc in Figure 22. Summarizing, "attenuation in well-constructed lines is proportional to the square root of frequency, inversely proportional to diameters (at optimum radius ratio), and proportional to the square root of resistivity." Pictures of the concentric lines involved in their investigation appear in Figures 23, 24 and 25.

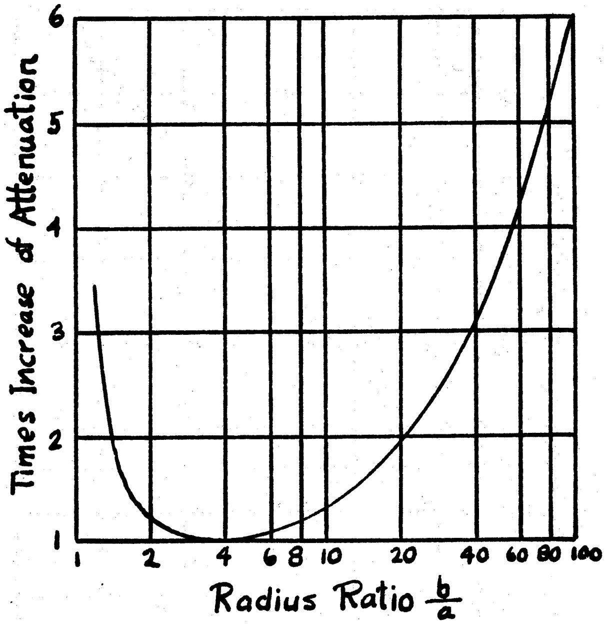


Figure 21 - PENALTY OF VARIATION FROM OPTIMUM RADIUS RATIO.

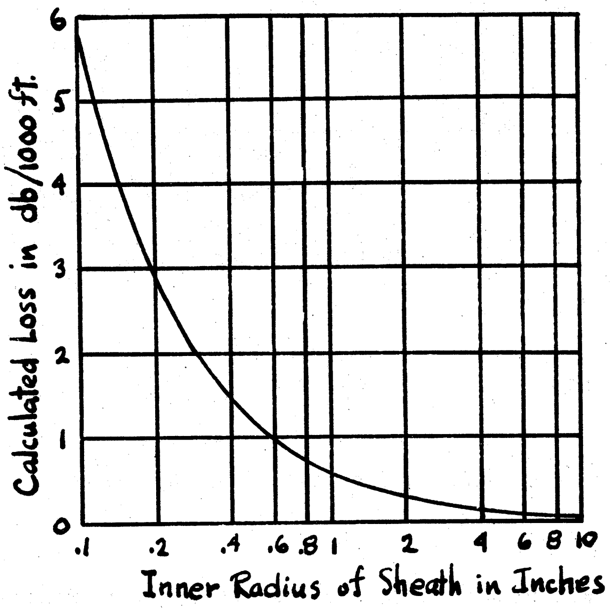


Figure 22 - CALCULATED 20 MEGACYCLE LOSS IN COPPER LINES OF OPTIMUM RADIUS RATIO.

L. V. King[71] in February 1933 published a mathematical treatise of shielding containers of extreme shapes at high frequencies. He discussed thickness of material and magnetic permeability, and suggested the use of iron for r-f shielding.

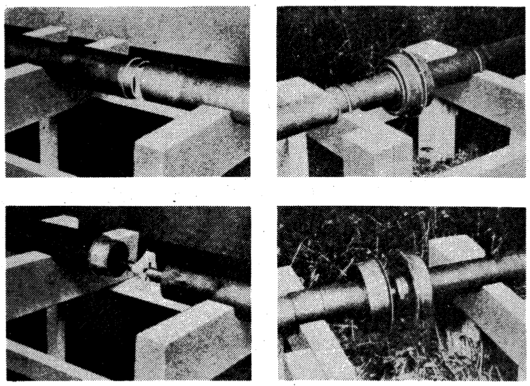


Figure 23 - CONCENTRIC LINE LOCK-JOINTS.

Wainman[72] in August of the same year published a description of one of the early amateur concentric lines employing a No. 2 copper wire in a three-quarter inch steel conduit. Whitmer[73] the following month, published an entirely mathematical paper concerned with the radiation resistance of coaxial lines, pointing out that the radiation resistance is 0.086 ohm compared to 3.02 ohm for the parallel-wire line. In each case the line was half-wavelength and the conductor spacing one-twentieth wavelength.

Schelkunoff[74] in October of 1934 published the very comprehensive mathematical development which laid the groundwork for the New York-

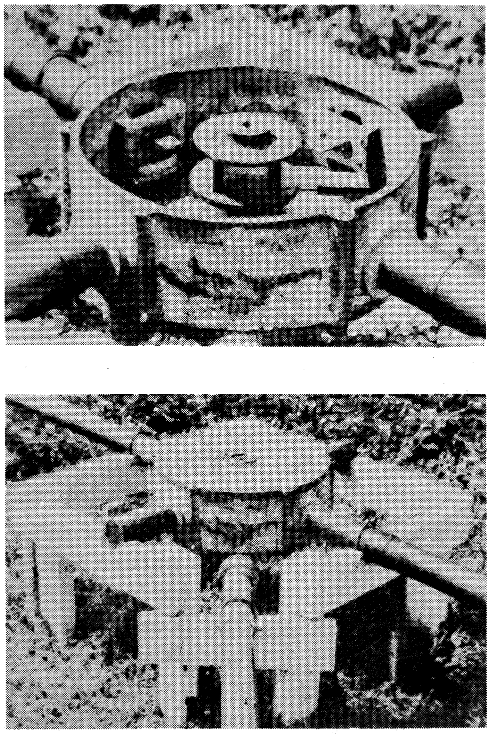


Figure 24 - CONCENTRIC LINE SELECTOR SWITCH.

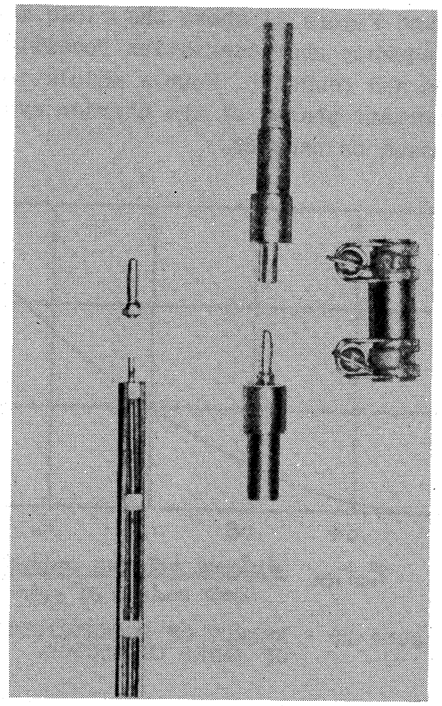


Figure 25 - FLEXIBLE CONCENTRIC LINES AND CONNECTORS.

Philadelphia coaxial cable, starting from elementary electromagnetic theory. Included in his paper are "perfectly conducting coaxial cylinders, imperfectly conducting coaxial cylinders, emf intensities in dielectrics, potential difference between two coaxial cylinders, external inductance, conversion of field equations into line equations, current distribution in cylindrical conductors, surface impedance of a solid wire, and of hollow cylindrical shells, the complex Poynting vector, surface impedance of tubular conductors, internal impedance of laminated conductors, terminal impedance for coaxial pairs, cylindrical waves and shields, cylindrical waves in dielectrics and metals, power losses in shields, and resistance of nearly coaxial tubular conductors".

Espenschied and Strieby[75] followed this with an article describing the results of their experiments with two new coaxial cables of 0.5" and 2.5" diameters; they suggested the use of coaxial cable not only for 200-channel carrier telephony, but wide-band television as well. Repeaters are required every ten miles and a photograph of one of them together with its circuit and characteristics appears in the article. The authors state that "For a given

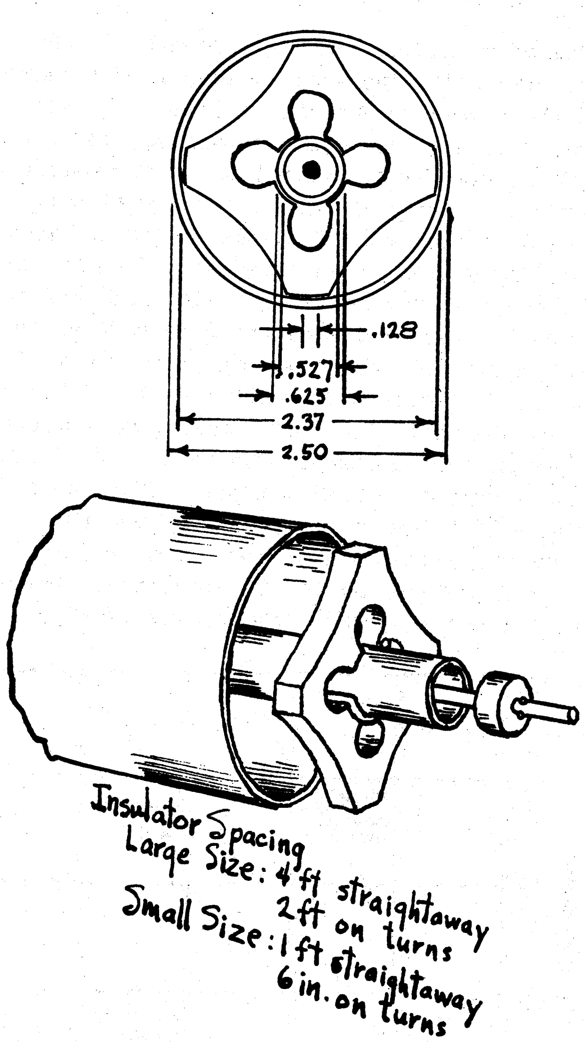


Figure 26 - PHOENXIVILLE COAXIAL CABLE DETAIL.

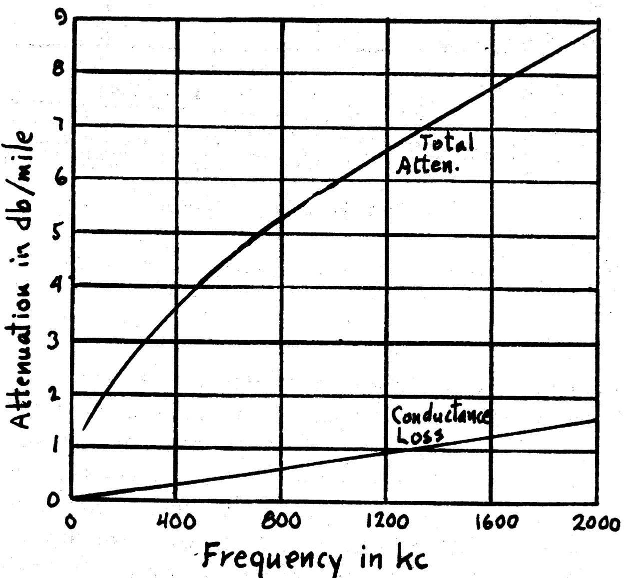


Figure 26a - ATTENUATION OF FLEXIBLE COAXIAL CABLE.

size of conductor and given length of line, the band width increases nearly as the square of the number of repeater points", and also that "For a given repeater spacing, the band width increases approximately as the square of conductor diameter". The experimental work included tests with flexible cables and also with the double rigid structure of Figure 26. The outer conductor is made air tight and dried with nitrogen. Measurements were carried to as high as 20 M.C. The frequency characteristic of the 0.3" inside diameter flexible cable is shown in Figure 26a, and comparisons between types of line appear in Figure 27. An effect of off-center central conductor is shown by Figure 28 to be rather severe. Since the elimination of crosstalk is one of the primary reasons for the use of the concentric line by the telephone company, the curves of Figure 29 were obtained showing pickup to be very small. It is pointed out that it is more economical not to use the lower five or ten percent of the frequency range than to use a heavier outer conductor for adequate shielding at low frequencies. Sixty-cycle current for operating the repeaters is sent along the line, however. Figure 30 shows the temperature regulation of the line and Figure 31 shows the close matching of frequency characteristics possible between line and repeater. Double modulation and other important phases of the carrier system are discussed on page 33.

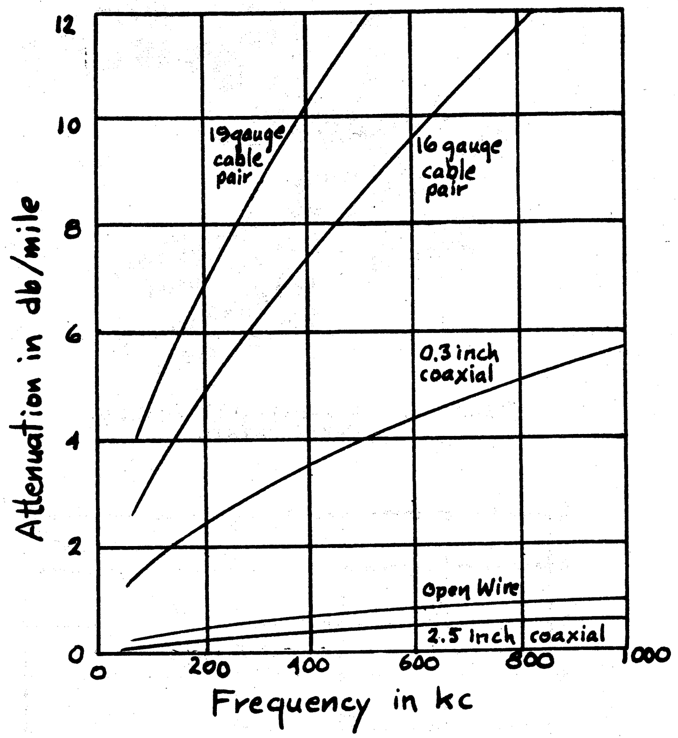


Figure 27 - ATTENUATION OF SEVERAL TYPES OF LINE.

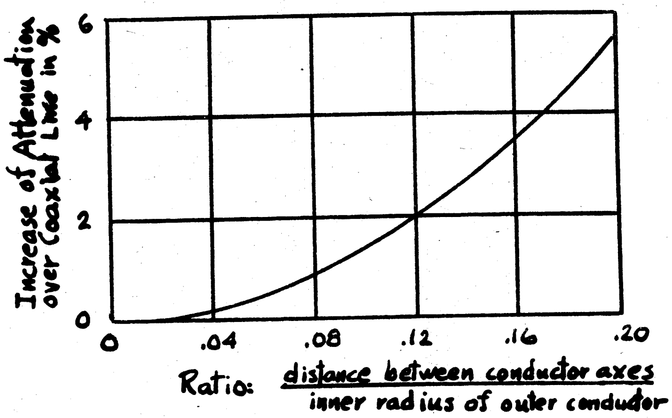


Figure 28 - EFFECT OF ECCENTRICITY OF INNER CONDUCTOR.

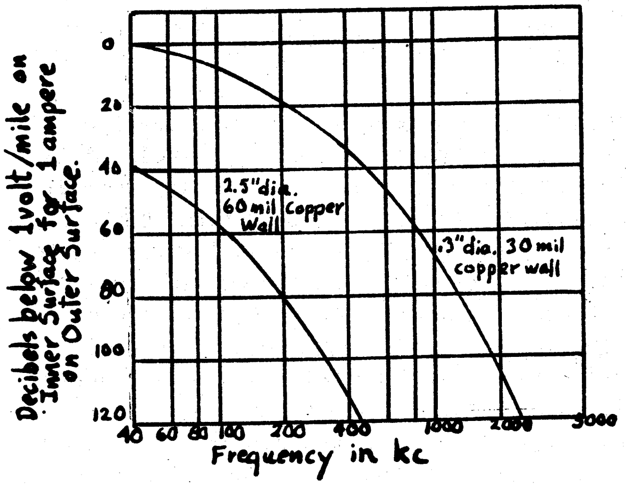


Figure 29 - COAXIAL-CABLE SHIELDING EFFECTIVENESS.

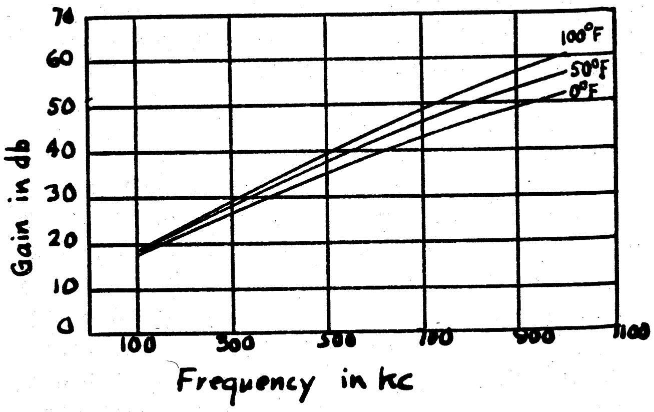


Figure 30 - TEMPERATURE REGULATION OF LINE WITH REPEATER.

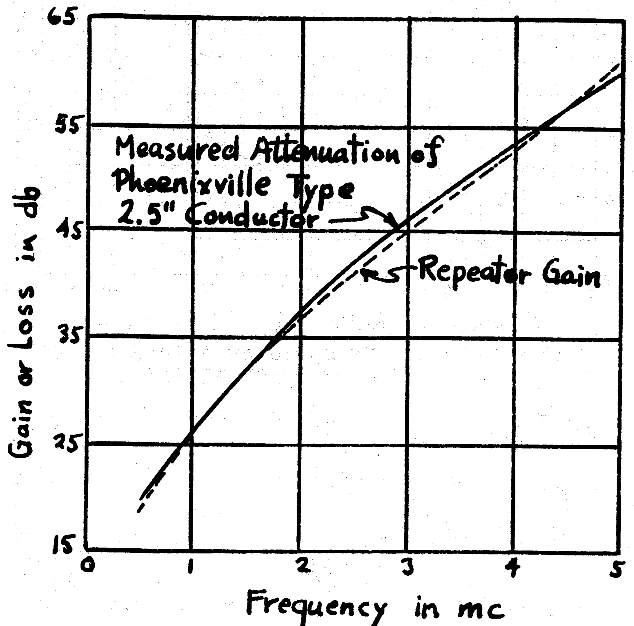


Figure 31 - FREQUENCY CHARACTERISTICS OF LINE WITH REPEATER.

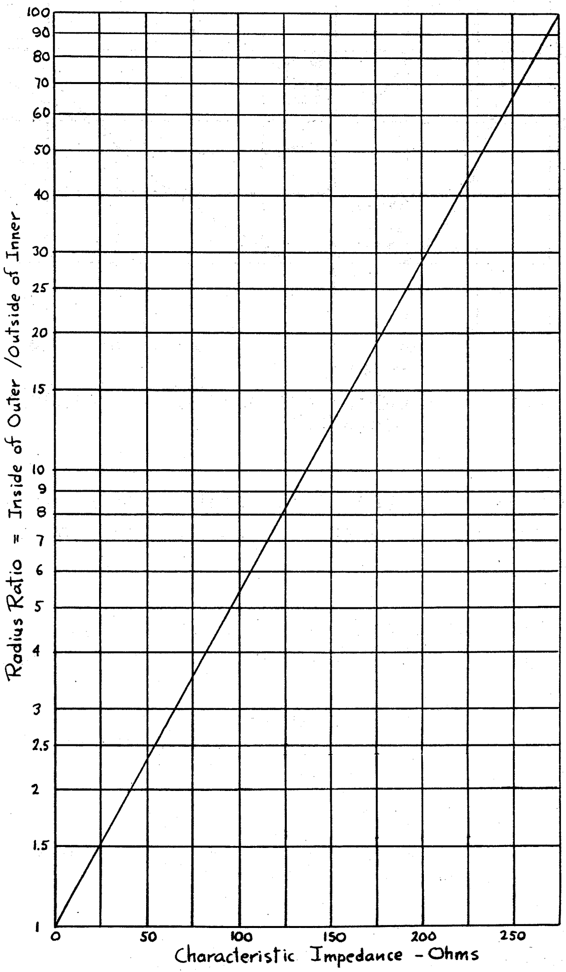


Figure 32 - CONCENTRIC LINE CHARACTERISTIC IMPEDANCE.

Schelkunoff[76] in December published a non-mathematical treatise on the mechanism by which wide-band transmission is accomplished over the cables described above. It is pointed out that if one wire of a parallel pair were replaced by a surrounding coaxial conductor separated from the other conductor by the original spacing, the capacity would double and the inductance halve, so that the surge impedance would have become halved. The optimum thickness of the tube wall should be a quarter-wavelength for that material, which for copper at one megacycle is one fourth of 0.415"; a thicker conductor actually requires heavier current density for the same total current. The mechanics of crosstalk and the degree of isolation of internal and external circuits are explained.

Baldwin[77] in February 1935 gave a complete description with computations and test results of the concentric line at KDYL.

The Engineering Handbook of the National Association of Broadcasters[62] gives a chart of characteristic impedance versus radius ratio in Figure 32, and line loss for broadcast frequencies as a function of outer conductor diameter in Figure 33, for optimum radius ratio.

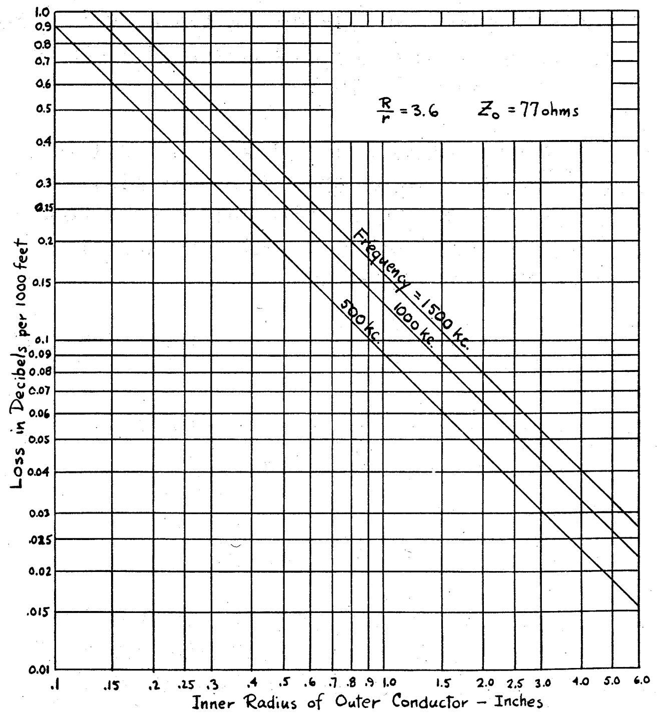


Figure 33 - CONCENTRIC LINE COMPUTED LOSS FOR OPTIMUM RADIUS RATIO.

Gordon and Sperling[65] extend equations (61 and 62) to give the capacity per inch of length as

$$C = \frac{10^{-5}}{7.08 \ ln \frac{r_1}{r_2}} \text{ microfarads,} \qquad (61a)$$

and the inductance per inch of length as

$$L = 5.08 \times 10^{-9} \ ln \frac{r_1}{r_2} \text{ henries,} \qquad (62a)$$

wherein r_1/r_2 is the radius ratio, and derive expressions (63) to get 3.6 for the optimum radius ratio and 76.7 ohms for the corresponding Z_o in equation (66). Their plotted data is in Figure 17. Sperling's[66] second article concerns itself with power losses and attenuation, and he develops the equations

$$C = \left(\frac{tk + q - t}{q}\right)\left(\frac{10^{-11}}{7.08 \ ln \frac{r_1}{r_2}}\right) \text{ farads/inch} \qquad (68)$$

$$Z_o = 60 \left[\frac{q}{tk + q - t}\right]^{1/2} ln \frac{r_1}{r_2} \text{ ohms} \qquad (69)$$

$$\text{Loss} = \frac{23.16 \text{ fkt}}{10^{10} \sqrt{kqt + q^2 + qt}} \text{ db/inch,} \qquad (70)$$

in which t is the thickness of the insulators and q is their spacing within the coaxial tube. His Figure 18 applies to concentric lines as well as open-wire lines.

Greene, Leibe, and Curtis[78] in April 1936 took up the general problem of maximizing and minimizing electrical shielded-circuit components for desired electrical behavior. The expression for attenuation,

$$\alpha = \frac{1}{2c}\sqrt{\frac{f}{\lambda_1}}\left(\frac{c}{b} + \sqrt{n}\right)\frac{\sqrt{e}}{2 \ ln \frac{c}{b}} \text{ nepers/cm,} \qquad (71)$$

where c and b are given in Figure 35,

n = conductivity ratio λ_1/λ_2,

f = frequency in cps,

and e = dielectric constant.

Introducing ρ to represent the radius ratio $\frac{c}{b}$, the attenuation is minimized with respect to this ratio and the optimum ratio found to be given by

$$ln\rho = \frac{\rho + \sqrt{n}}{\rho}. \qquad (72)$$

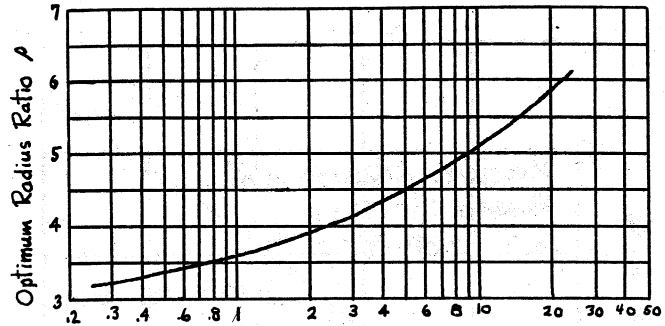


Figure 34 - RADIUS RATIO AS A FUNCTION OF RATIO OF CONDUCTIVITIES.

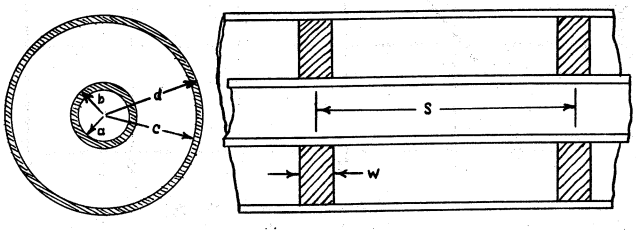


Figure 35 - COAXIAL LINE DETAILS.

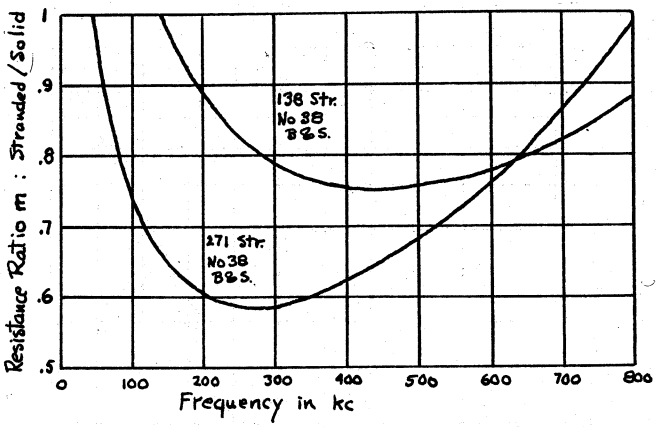


Figure 36 - RESISTANCE RATIOS FOR CONDUCTOR STRANDING.

The value of ρ from this relation is plotted in Figure 34 as a function of n, the conductivity ratio.

When the spacer construction of Figure 35 is taken into account, the attenuation becomes

$$\alpha = \frac{1}{2c}\sqrt{\frac{f}{\lambda_1}}\sqrt{\frac{e_1 s + (e_2 - e_1)w}{s}}\left(\frac{\rho + \sqrt{n}}{2 \ ln \rho}\right) + \frac{p\omega e_2 w}{2\sqrt{s}} \frac{1}{\sqrt{e_1 s + (e_2 - e_1)w}} \text{ nepers/cm,} \qquad (73)$$

where e_1 is the dielectric constant of the gaseous dielectric, e_2 the dielectric constant of the spacers, and p is dielectric power factor. The characteristic impedance becomes

$$Z_0 = \frac{2 \ln \rho}{\sqrt{e_1 + \frac{(e_2 - e_1)w}{s}}} \text{ ohms.} \qquad (74)$$

For solid spacers with flat sides, such as Figure 35, the curve of Figure 34 still applies.

The effect of frequency upon the optimum ratio is discussed. For thin-walled tubes (72) becomes

$$\ln\rho = \frac{\rho + nt}{\rho} \qquad (75)$$

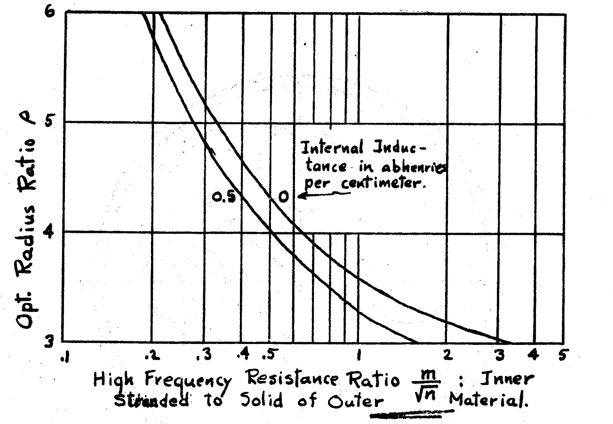


Figure 37 - OPTIMUM RADIUS RATIO FOR STRANDED INNER CONDUCTOR.

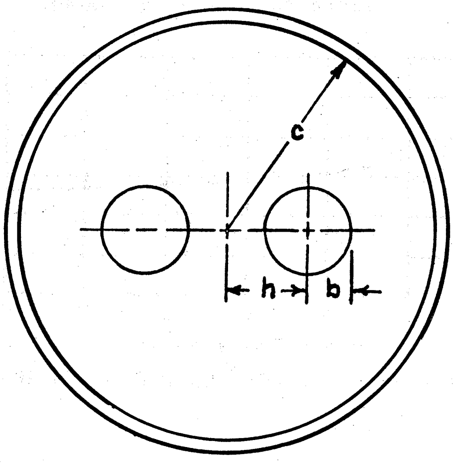


Figure 38 - SHIELDED PAIR.

In comparing stranded and solid conductors, the ratio of resistances of stranded and solid of the same outside diameter is called m, which is plotted in Figure 36. Formulas are given for the case of the stranded conductor surrounded by a solid conductor of any material. The effect upon optimum radius ratio of the high-frequency resistance ratio of a stranded conductor to a solid one of the same size but of outer conductor material, namely $\frac{m}{\sqrt{n}}$, is shown in Figure 37. The formula for the attenuation of the construction shown in Figure 38 is given, and the effect of conductivity ratio upon optimum radius ratio and upon $\frac{h}{c}$ is given by Figure 39. Expressions for characteristic impedance and its maximum value are given, and the affects of dielectric and frequency are considered. The expressions are shown to apply to an infinite-radius shield as a special case, which is of course the open-wire line.

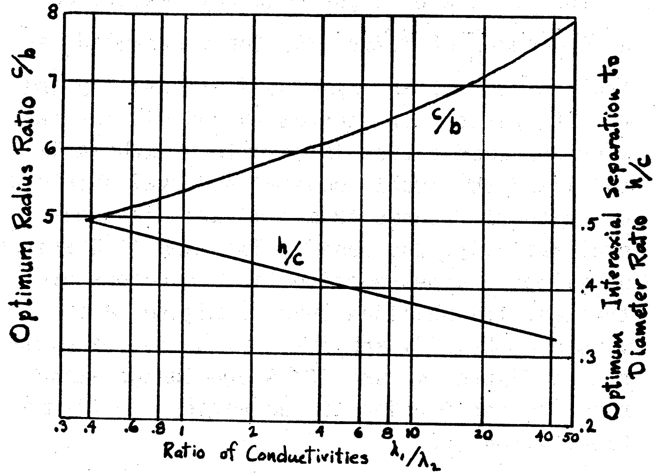


Figure 39 - EFFECT OF CONDUCTIVITY.

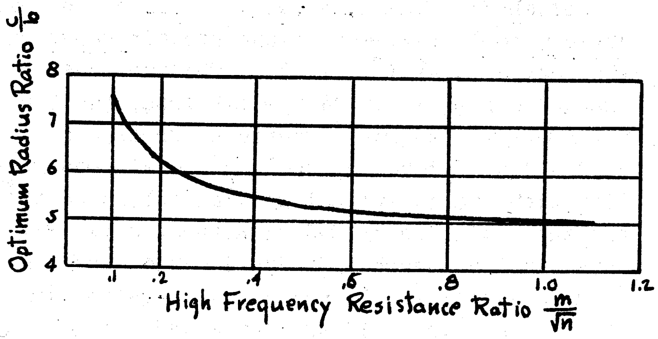


Figure 40 - OPTIMUM RADIUS RATIO OF SHIELDED STRANDED PAIR.

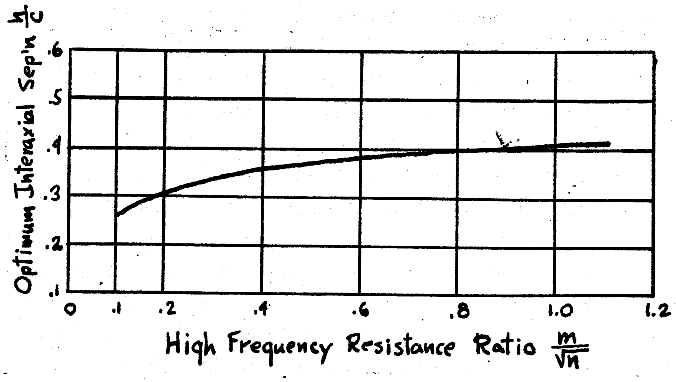


Figure 41 - OPTIMUM SPACING RATIO OF SHIELDED STRANDED PAIR.

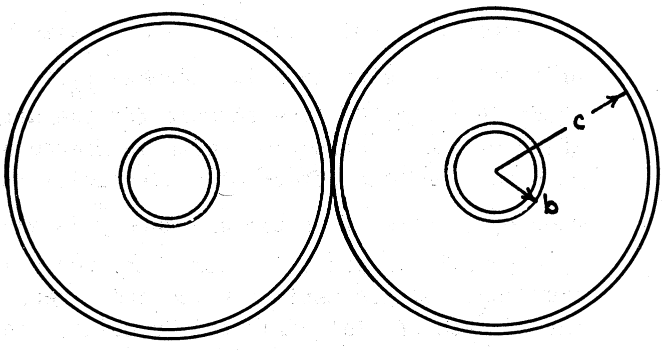


Figure 42 - DOUBLE COAXIAL CIRCUIT.

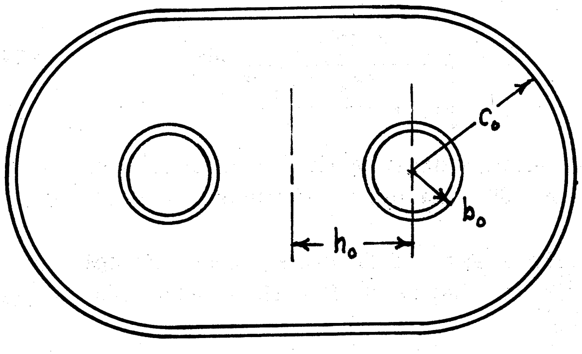


Figure 43 - OVAL SHIELDED PAIR.

The effects of high-frequency resistance upon $\frac{h}{c}$ and $\frac{c}{b}$ of Figure 38, for a stranded central conductor, are discussed extensively and results plotted in Figures 40 and 41. The pair with shielded return is described mathematically and the double coaxial and oval shielded circuits of Figures 42 and 43 are discussed qualitatively. The desirability and limitations of elliptic cross-section for conductors are treated. The shielded quad of Figure 44 is discussed briefly and its optimum ratios plotted in Figures 45 and 46. A conclusion summarizes the properties of the various configurations taken up.

Strieby[79] in January 1937 published the last article to be reviewed in the present paper on the very interesting New York-Philadelphia coaxial cable. It is 94.5 miles long, has ten repeaters at ten-mile intervals, handles a band width of 1000 kc, and has a dual construction with two individual coaxial cables so that the whole will carry 240 simultaneous two-way conversations, each individual cable carrying unidirectional intelligence. The construction is shown in Figure 47 and the as yet incomplete terminal apparatus in Figure 48. Each coaxial shield of the single cable is 0.265" inside diameter insulated from a number 13 B and S copper wire by rubber spacers every three-quarters of an inch. Two quads are also carried within the cable, whose outside diameter is seven-eighths of an inch. The outer coaxial conductors are made up "of nine overlapping copper tapes which form a tube 0.02" thick, held together with a wrapping of iron tape". The use of the cable for carrier telephony is shown in Figure 49. The cable attenuation is shown in Figure 50 and the crosstalk in Figure 51. Figure 52 shows the frequency characteristic of a typical telephone channel.

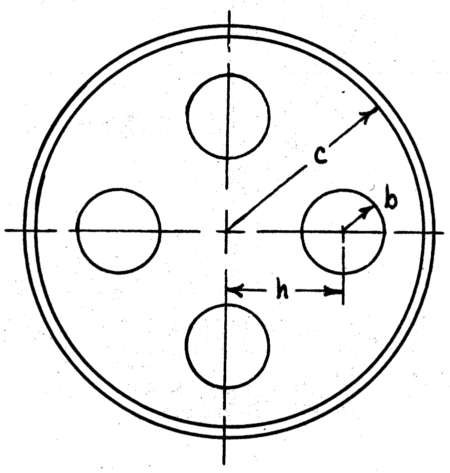


Figure 44 - SHIELDED QUAD.

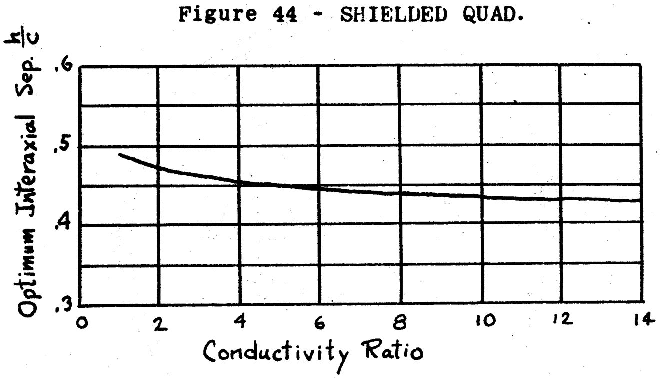


Figure 45 - SHIELDED QUAD OPTIMUM SPACING.

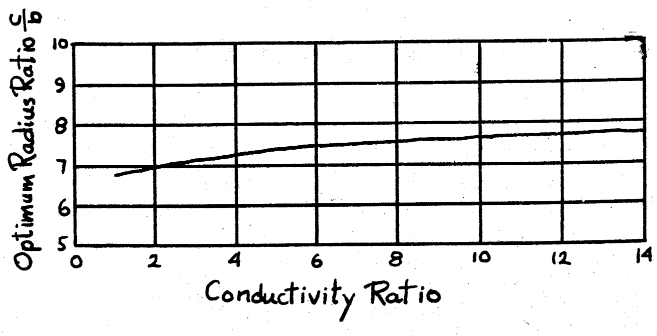


Figure 46 - SHIELDED QUAD RADIUS RATIO.

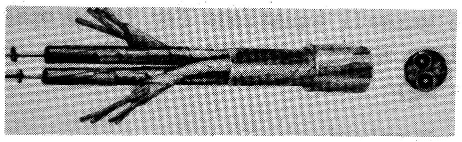


Figure 47
VIEW SHOWING STRUCTURE OF COAXIAL CABLE.

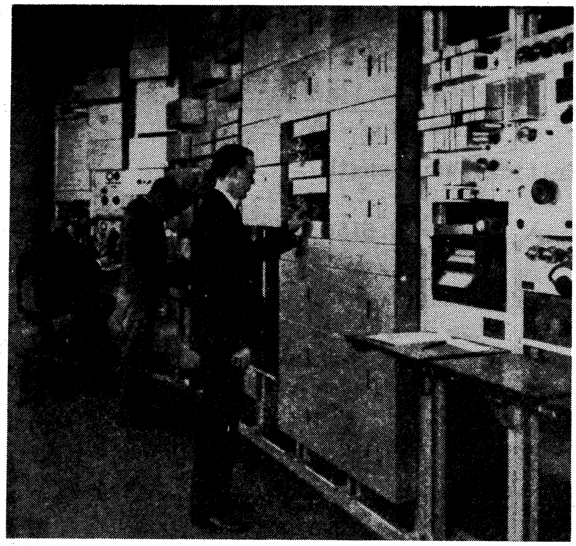


Figure 48
THE NEW YORK TERMINAL OF THE COAXIAL SYSTEM.

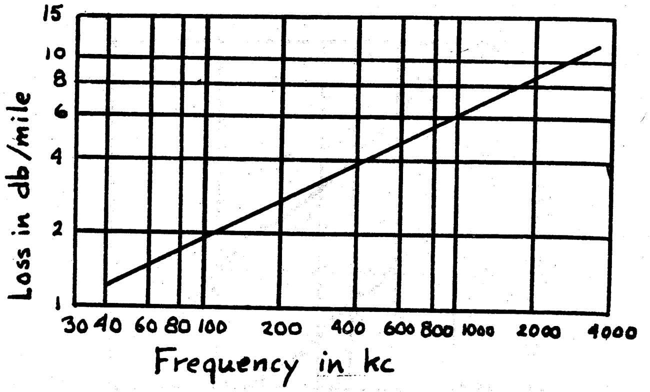


Figure 50 - ATTENUATION OF NEW YORK-PHILADELPHIA CABLE.

(Figure 51 on page 24)

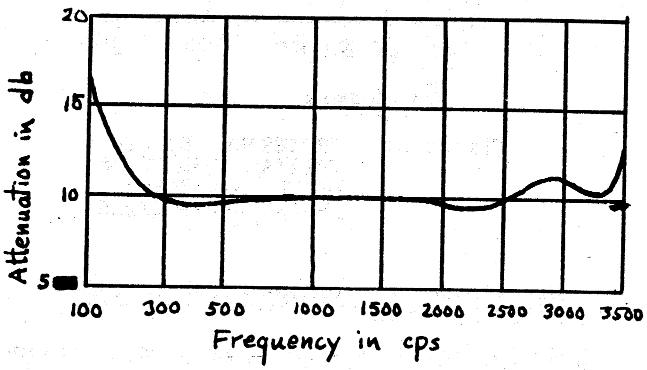


Figure 52 - FREQUENCY CHARACTERISTIC OF A TYPICAL CHANNEL.

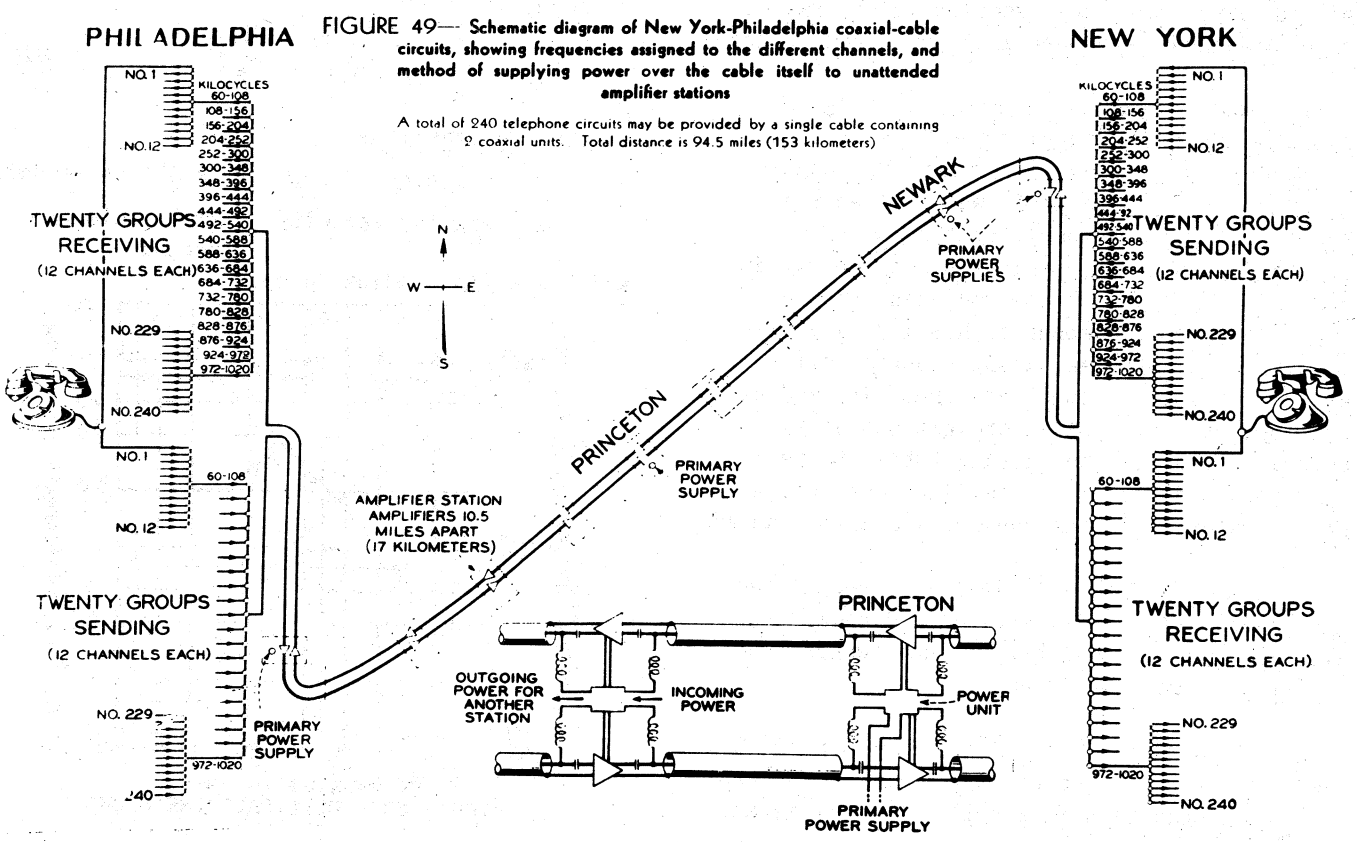


FIGURE 49— **Schematic diagram of New York-Philadelphia coaxial-cable circuits, showing frequencies assigned to the different channels, and method of supplying power over the cable itself to unattended amplifier stations**

A total of 240 telephone circuits may be provided by a single cable containing 2 coaxial units. Total distance is 94.5 miles (153 kilometers)

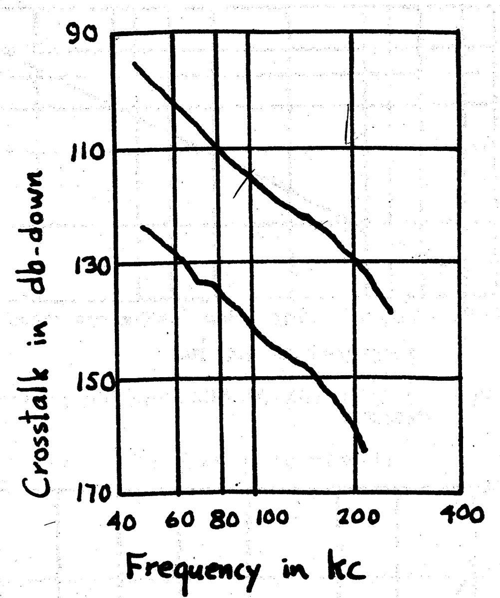


Figure 51 - CROSSTALK BETWEEN COAXIAL CONDUCTORS OF THE NEW YORK-PHILADELPHIA CABLE.

Wave-Guide Development, 1893 to 1936

This new development of intelligence transmission involves the use of either a simple dielectric wire or a hollow conducting tube in which a special dielectric may or may not be placed. Both types are essentially the same problem and so have been included together here.

Initial literature on the subject is confined to be mathematical conjecture of J. J. Thompson[80], Heaviside[81], and Lord Rayleigh[82]. In 1933 a British Patent[83] was awarded the Western Electric Company for the use of a dielectric guide of combinations of "paraffin wax, mico, rubber, phenol condensation products, terpineol, camphor diphenyl derivatives, or similar substances having substantially greater dielectric constants than air, and of low conductivity". Line terminations using several types of transmitting tubes and receivers are depicted in its illustrations. A French patent[84] on the subject was also granted.

Bergman and Kruegel[85] in 1934 investigated the fields within hollow pipes and radiation from their open ends, but did not suggest their use as transmission lines.

Dr. W. L. Barrow[86] in October 1936 published the results of one of the first investigations of wave propagation in a hollow tube of metal.

The following expressions are derived directly from the Maxwell equations for the propagation in a hollow metal tube filled with air ($e_1 = e_o$ and $\mu_1 = \mu_o$):

Phase Constant

$$\beta = \sqrt{\left(\frac{\omega}{c}\right)^2 - \left(\frac{y_o}{a}\right)^2} \qquad (76)$$

Critical Frequency

$$f_o = \frac{cy_o}{2\pi a} \simeq \frac{1.148 \times 10^{10}}{a} \qquad (77)$$

Critical Wavelength

$$\lambda_o = \frac{c}{f_o} \simeq 2.615a \qquad (78)$$

Internal Wavelength

$$\lambda = \frac{2\pi}{\sqrt{\left(\frac{\omega}{c}\right)^2 - \left(\frac{\gamma_o}{a}\right)^2}} \qquad (79)$$

Phase Velocity

$$v_p = \frac{\omega}{\sqrt{\left(\frac{\omega}{c}\right)^2 - \left(\frac{Y_o}{a}\right)^2}} \qquad (80)$$

Group Velocity

$$v_g = \frac{c^2}{v_p} = \frac{c^2}{\omega}\sqrt{\left(\frac{\omega}{c}\right)^2 - \left(\frac{Y_o}{a}\right)^2} \qquad (81)$$

In these expressions

c = velocity of light = 3×10^{10} cm/second,
ω = angular velocity = $2\pi f$, radians/second,
and a = tube radius in cm.

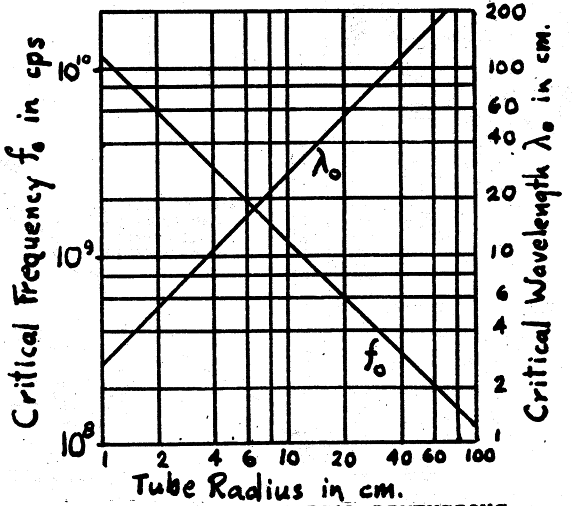


Figure 53 - CRITICAL DIMENSIONS.

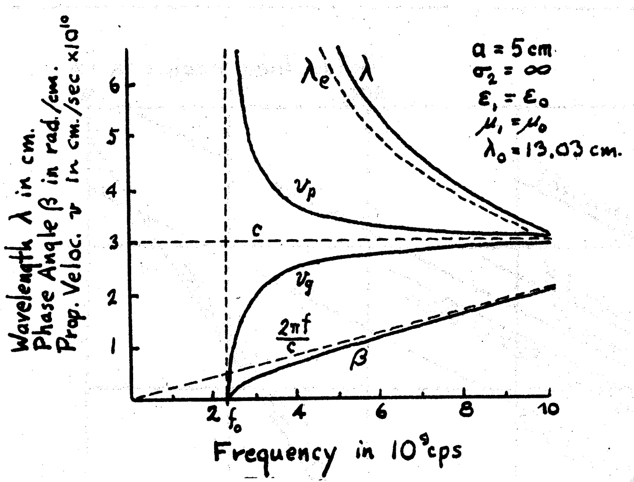
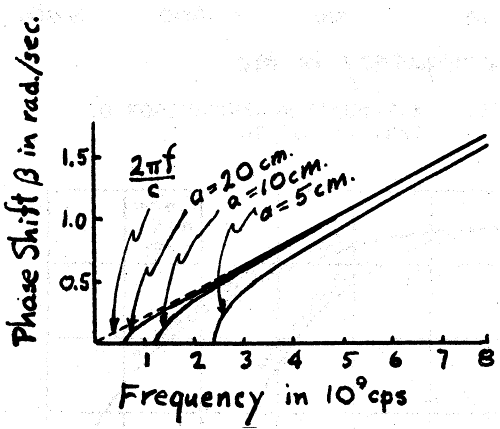


Figure 54 - TRANSMISSION CHARACTERISTICS.



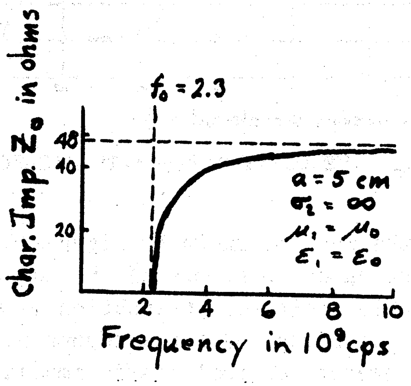


Figure 56 - SURGE IMPEDANCE.

Equation (77) is plotted in Figure 53 together with the corresponding free space wavelength $\lambda_o = c/f_o$. Practical application limits the tube diameters to 4 to 100 cm., or frequencies of 2.3×10^8 to 6×10^9 cps. Equations (76, 78, 79, 80 and 81) are plotted in Figure 54 for a tube of five-centimeter radius. The value of β is shown for several radii in Figure 55 from equation (76). Electric and magnetic fields are computed analytically and plotted.

A quantity Z_o is defined as the ratio of the transverse emf between tube axis and wall to the strength of longitudinal current and is found to be

$$Z_o = 1.44 \times 10^{12} \frac{\beta}{\omega} = \frac{1.44 \times 10^{12}}{v_p} \text{ ohms,} \quad (82)$$

and is plotted in Figure 56. The quantity approaches 48 ohms as an asymptote.

For imperfectly conducting tubes the attenuation constant is found to be

$$\alpha = \frac{\mu_2}{a^{3/2}\,\mu_o} \sqrt{\frac{\mu}{c\mu_2\sigma_2}} \left[\frac{\lambda_e}{a} - \left(\frac{Y_o}{2\pi}\right)^2 \left(\frac{\lambda_e}{a}\right)^3\right]^{-\frac{1}{2}} \quad (83)$$

in which new constants are tabulated in the article; the equation does not hold in the vicinity λ_e near λ_o. Minimization of this expression as shown in Figure 57 yields

$$\frac{\lambda_e}{a} = 1.5 \text{ for minimum attenuation.} \quad (84)$$

The resultant plot of minimum attenuation appears in Figure 58 together with other high-efficiency lines taken from the work of Sterba and Feldman[46]. The application of the tube to high-pass filtering is shown in Figure 59 using equations (76) and (83), and the pass band can be seen to be very good in attenuation and phase for a band 10,000 megacycles wide extending from 2×10^9 cycles to 12×10^9 cycles.

Terminal devices are discussed at length and experimental investigation of the standing waves at the end of the tube is shown graphically. Multiplex transmission using the four types of waves, and field intensity of radiation from the open end of the hollow tube transmission line are discussed.

In April 1936 Carson, Mead, and Schelkunoff[87] published the mathematical theory of their "hyper-frequency wave guides". After outlining previous work done on the subject, they proceed to show that actually but one type of wave may be transmitted by the system, and that its attenuation decreases with increasing frequency. Non-dissipative and dissipative hollow conducting guides are then taken up in detail. The attenuation for the fundamental and first harmonic of both electric and magnetic waves, considering only conductor loss, is plotted in Figure 60. Lastly, dielectric cylindrical guides are taken up hypothetically.

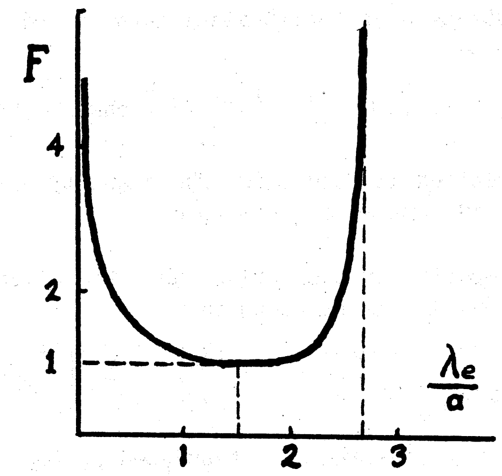


Figure 57 - MINIMIZATION OF ATTENUATION WITH λe/a.

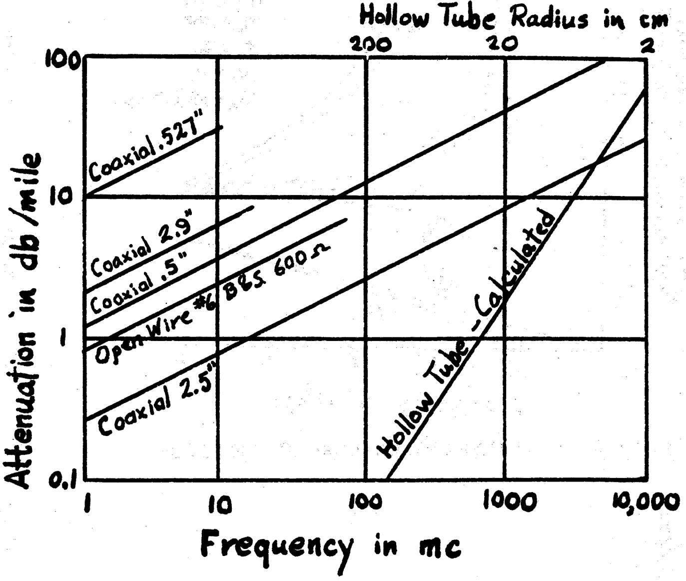


Figure 58 - ATTENUATION COMPARISON OF VARIOUS LINES.

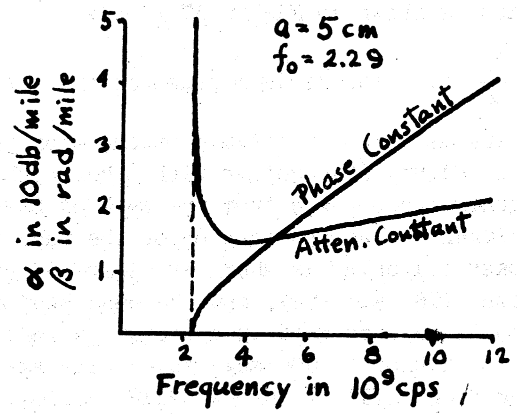


Figure 59 - TRANSMISSION CHARACTERISTICS.

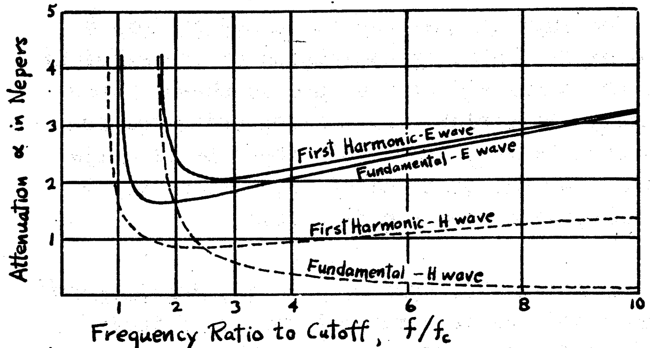


Figure 60 - ATTENUATION OF HOLLOW CONDUCTING CYLINDER.

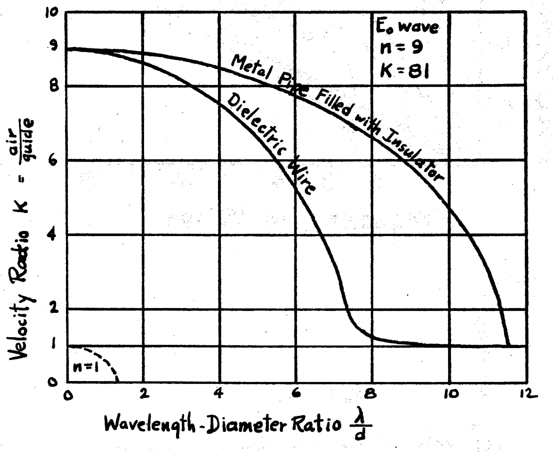


Figure 61 - VELOCITY RATIO WITH DIELECTRIC.

Southworth[88] the same month published a companion paper to the previous one giving general considerations and experimental results. It is pointed out that the H_0 wave (fundamental H wave of Figure 60) has progressively less attenuation as frequency is increased, but that it also requires the highest frequency for its use, putting it in the range where the high-frequency art is least developed. The effect of wavelength-diameter ratio upon the velocity of the wave is shown by Figures 61 and 62 for three cases; portions of the curves were experimentally verified. Attenuation is shown in decibels in Figure 63 for a copper pipe of 5 inches diameter. Characteristic impedance is defined as the cross-sectional area of the integrated Poynting vector divided by the square of the effective current, and is shown in Figure 64 for a copper pipe of 4 inches diameter. Frequency selectivity radiation, and several equipment items are also discussed. In Figure 65 the E_0 and E_1 waves have been so designated because there is an electric force component in the direction of propagation. The similarity of the E_0 and E_1 waves in the case of the shielded conductor and pair of conductors is discussed pictorially. Figures 66 thru 69 show drawings of some of the equipment.

In May 1936 Southworth[89] published still another description of the equipment covering the same ground in very brief and non-technical manner, and including the interesting pictures of his equipment shown in Figures 70, 71 and 72.

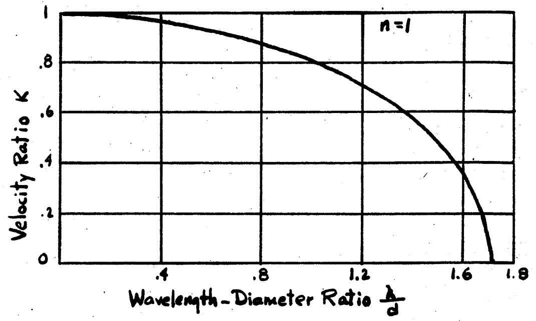


Figure 62 - VELOCITY RATIO FOR HOLLOW TUBE.

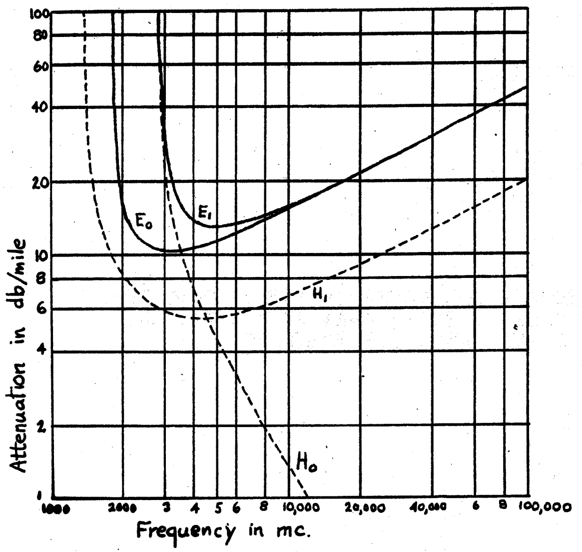


Figure 63 - ATTENUATION OF THE FOUR BASIC TYPES OF WAVES.

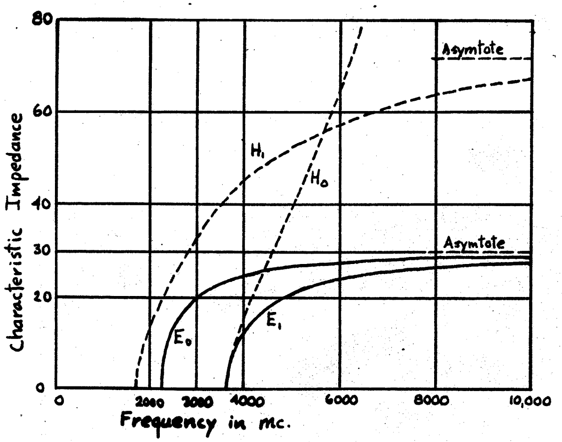


Figure 64 - CALCULATED CHARACTERISTIC IMPEDANCE OF 4" COPPER PIPE.

(Figure 65 on page 28)

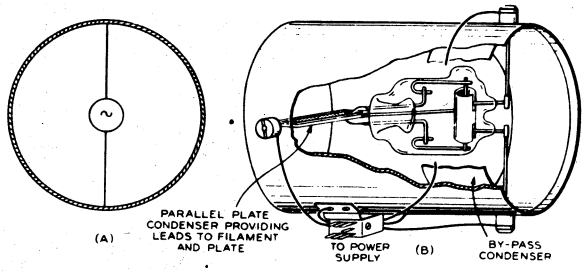
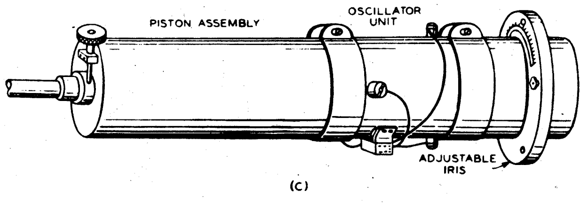


Figure 66 - OSCILLATOR OF WAVE-GUIDE GENERATOR.



Various component parts of a wave guide generator. (*A*) Schematic representation. (*B*) The oscillator unit. (*C*) Complete generator including oscillator piston and iris.

Figure 67 - COMPLETE WAVE-GUIDE GENERATOR.

(Figure 68 on page 28)

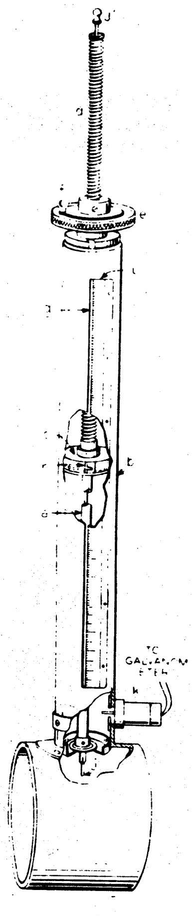


Figure 69
A FORM OF COAXIAL CONDUCTOR WAVE METER PARITCULARLY ADAPTABLE TO WAVE-GUIDE WORK.

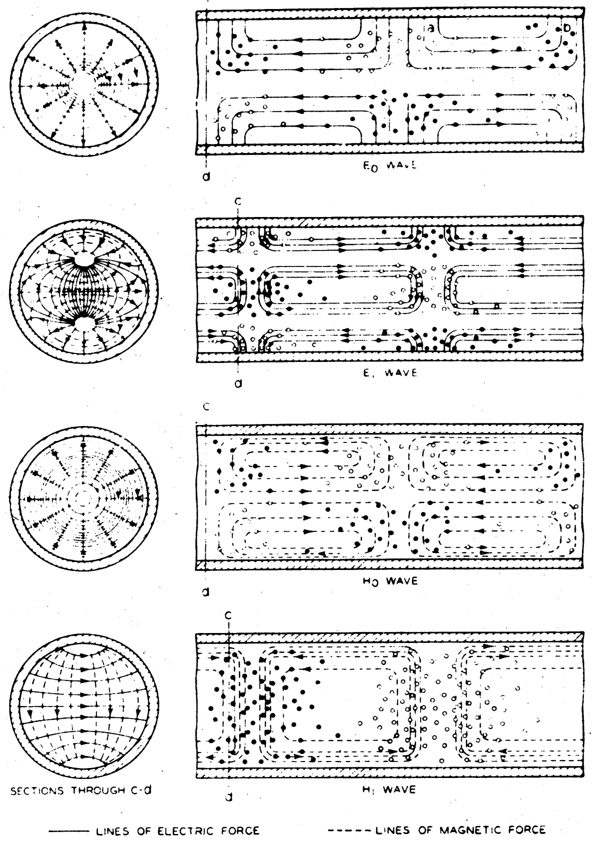


Figure 65 - APPROXIMATE CONFIGURATION OF LINES OF ELECTRIC AND MAGNETIC FORCE IN A TYPICAL WAVE-GUIDE. SMALL SOLID CIRCLES REPRESENT LINES OF FORCE DIRECTED AWAY FROM OBSERVER. PROPAGATION IS ASSUMED TO BE DIRECTED TO THE RIGHT AND AWAY FROM THE OBSERVER.

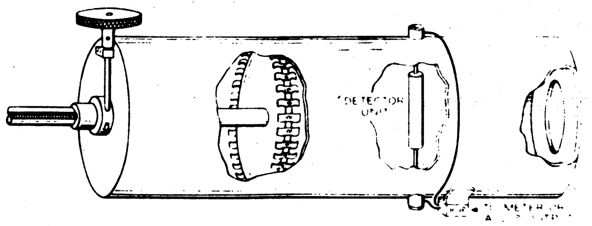


Figure 68 - A TUNED RECEIVER BASED ON THE RESONANT CAVITY PRINCIPLE.

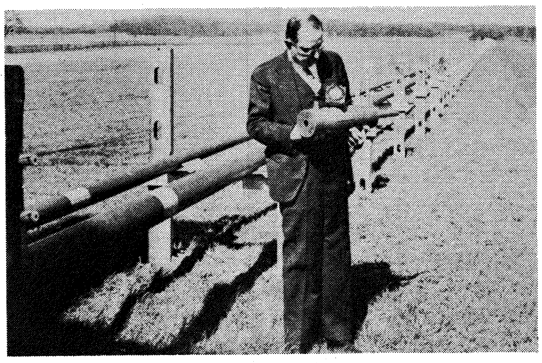


Figure 71 - G.G. SOUTHWORTH HOLDING ONE OF THE RESONANT CHAMBERS USED FOR TESTS OF WAVE-GUIDE TRANSMISSION. BEHIND HIM ARE THE TWO EXPERIMENTAL LINES.

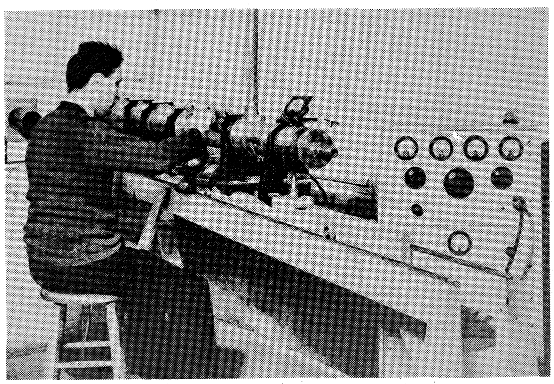


Figure 72 - A.P. KING AT THE SENDING END OF THE EXPERIMENTAL WAVE-GUIDE AT HOLMDEL.

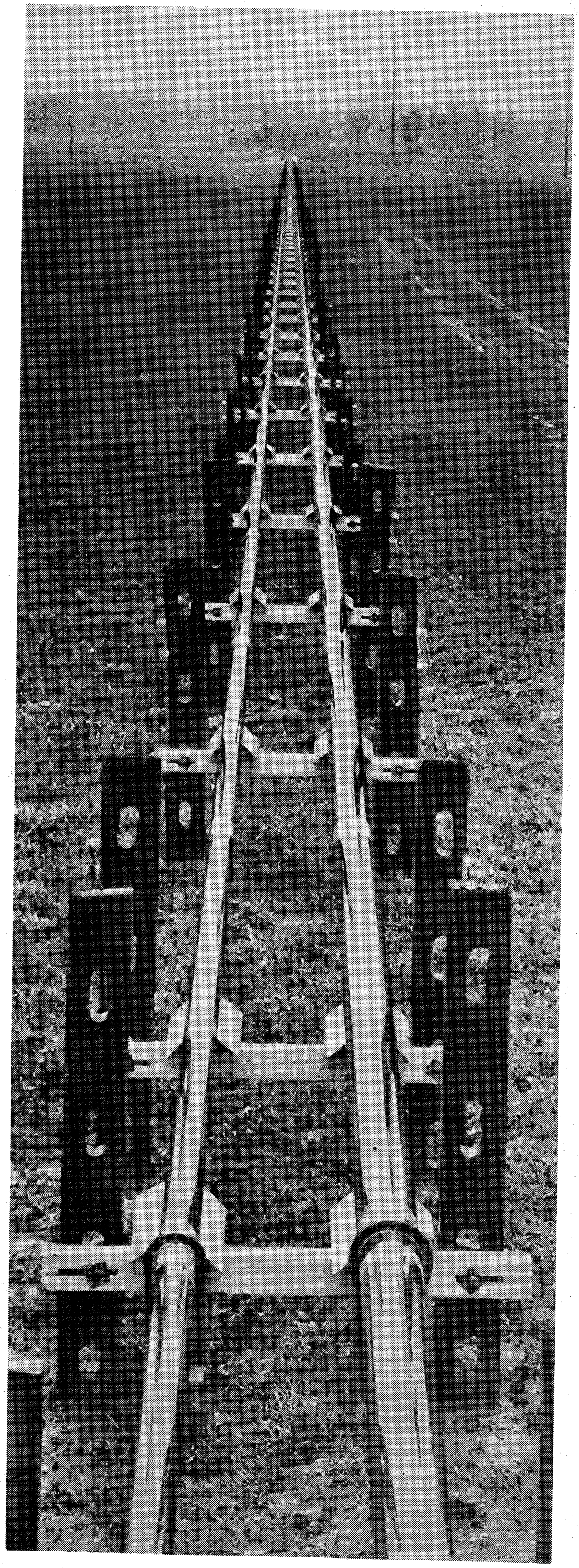


Figure 70 - EXPERIMENTAL WAVE-GUIDES AT THE HOLMDEL LABORATORY.

Allied Researches Affecting the General Theory

In all theoretical discussions of transmission lines of any of the previously mentioned types many physical constants enter the calculation. Electrical and heat conductivity, magnetic permeability, coefficient of expansion, hardness, chemical inertness, and many other properties each have their particular effect upon the substances used in transmission-line construction. And of course these substances are legion. Most important may well be the conductor itself, but for accurate theoretical prediction, the properties of the surrounding media, the internal medium (for dielectric guides), and the insulating supports must be taken into account.

Skin and proximity effects no doubt play one of the largest parts in transmission-line design; of the prodigious amount of material that has been published on the subject only a very small portion having direct bearing upon the high-frequency problem has been included. Skin effect is the tendency of current to flow on the outside of its conductor as frequency is increased, while proximity effect is the redistribution of current within the conductor due to the presence of another nearby conductor; both are redistribution effects.

Kennelly, Laws, and Pierce[90] in 1915 did some of the very early work with frequencies up to 5000 cycles and H. B. Dwight[91] extended this to 70,000 cycles for tubular and flat conductors, giving a checked method, obtained from an asymtotic interpretation of the formula, for calculation of the "resistance ratio". Again in 1922 Dwight[92] studied proximity effect and spirality effect and suggested a tubular form to reduce the skin effect; calculations are shown for thin tubes also. In 1923 he published still other papers, especially for thin tubes[93] and for isolated tubes[94].

Carson[13] in 1921 developed formulas for a-c resistance and proximity effects from elementary considerations, and C. Snow[95] in 1925 published a quite comprehensive mathematical discussion and derivation of the a-c distribution in cylindrical conductors, obtaining coefficients for inductance, capacitance, resistance, conductance, attenuation, phase velocity, and energy.

Lee and Miyauchi[96] prepared in 1927 papers concerned with the mathematical methods used, results obtained, problems not yet analyzed, and the extent of experimental checks in the history and treatment of skin and proximity effects.

Dwight[97] in 1931 discussed the characteristics of lead sheaths for underground cables of one, two, and three conductors, with sheaths open and short-circuited, and Webb[98] provided an experimental check on the formulas of Carson and Butterworth obtaining the curves shown in Figures 73 and 74. The quantity $2\sqrt{\omega/R_0}$, in which ω is angular frequency and R_0 is d-c resistance per centimeter, is used as abscissa and the experimental verification was carried only as high as 6000 cps. While not radio frequency, these articles are important steps in the r-f evaluation.

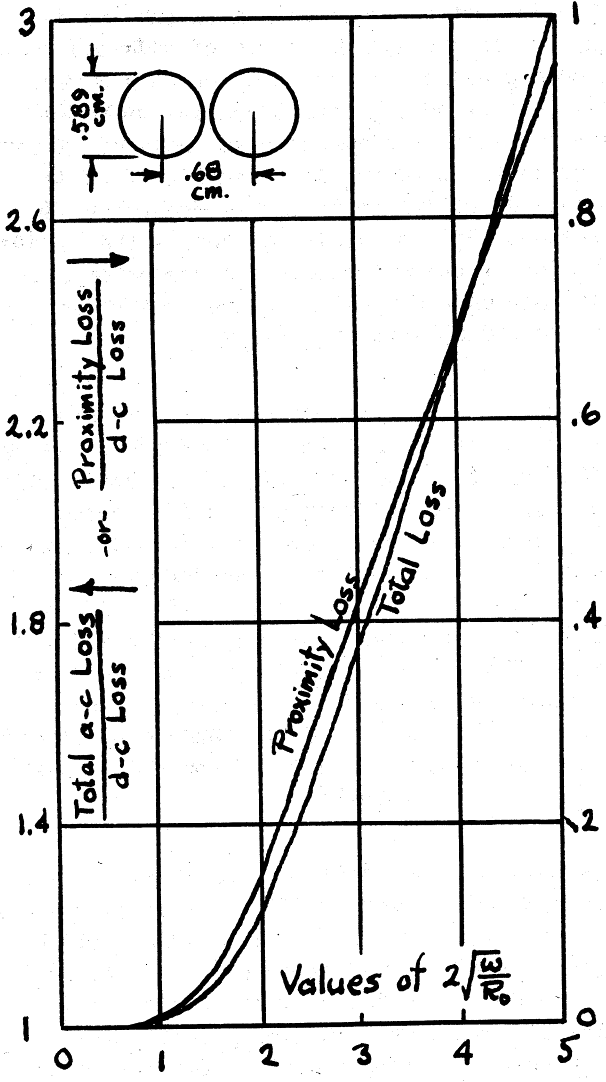


Figure 73 - TOTAL AND PROXIMITY LOSSES.

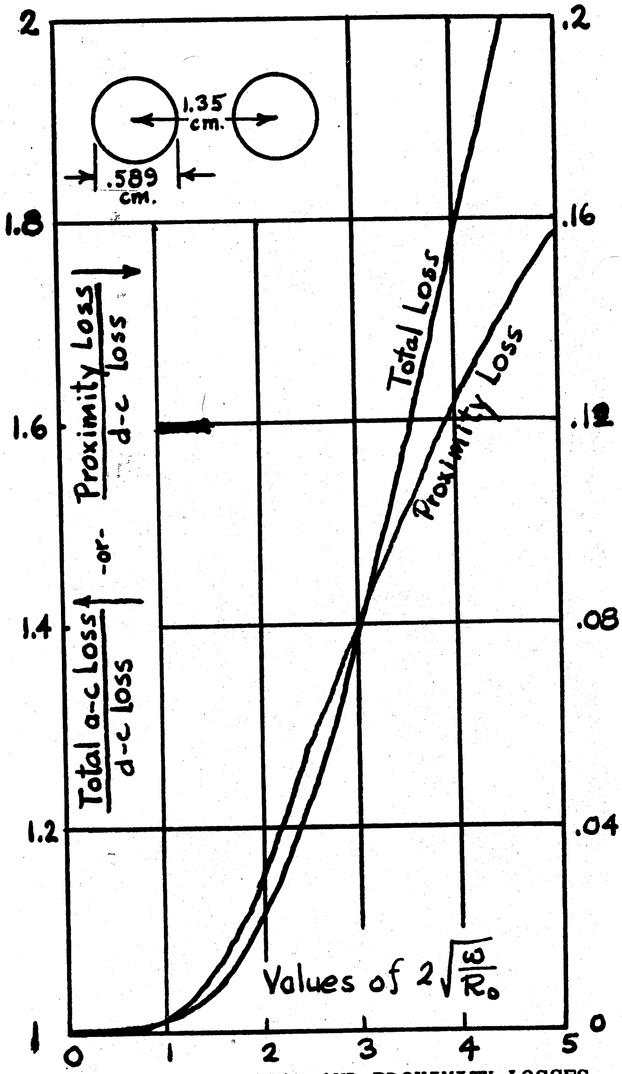


Figure 74 - TOTAL AND PROXIMITY LOSSES.

Sterba and Feldman[46] in July 1932 gave the high-frequency resistance of a concentric-tube line in equation 64 of page 15 and Figure 19 of page 15; the remarks in that section are pertinent to the question of high-frequency resistance. Schelkunoff[99] in December 1932 wrote a very understandable account of the behavior of r-f resistance. By considering the inner conductor of a coaxial pair to be composed of elementary cylinders he shows that the current in each one, beginning with the outside, shifts in phase and magnitude according to points 1, 2, 3, 4, 5, and 6 in Figure 75. At radio frequencies the amplitude diminishes much more rapidly than can be shown on a small scale such as Figure 75. The net effect is a variation of resistance with conductor (shell) thickness as shown in Figure 76. Another Arnold[100] article in 1936 concerned itself with the a-c resistance of tubular conductors in various configurations but confined to audio frequencies.

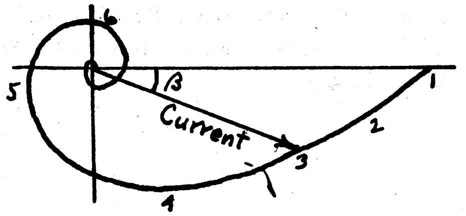


Figure 75 - PHASE SHIFT AND DECLINE OF CURRENT INWARD FROM THE SURFACE.

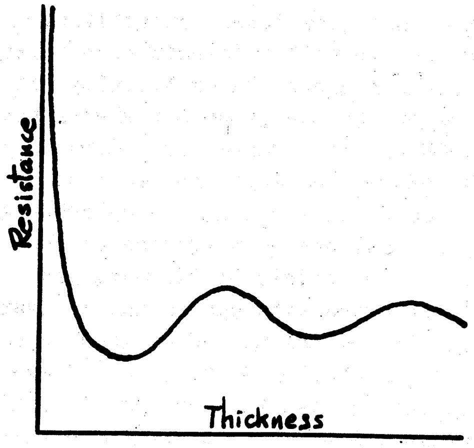


Figure 76 - RADIO FREQUENCY RESISTANCE.

Mechanical Design, Construction, and Maintenance

Notwithstanding the scarcity of material in the literature concerned with mechanical design considerations, the purely physical requirements of a transmission line probably rank equally with electrical requirements. For instance the open-wire line requires that special thought be given to insulation and the reliability of that insulation during adverse weather; the insulators must be supported by cross-arms and poles -- involving labor and material costs far exceeding the line itself -- and suitable terrain must be found over which it is possible to erect the line. Exhaustive weathering tests should precede the erection of every type of wooden pole. The change of electrical properties of the line as oxides form on the surface must be predetermined. H. Ilgen[101] has written on this subject; his tests between 1000 and 15,000 kc show that oxidation has little effect upon inductance, but that electrolytic coverings have a great effect upon capacitance. The catenary itself should be investigated for sag with time and for strength in the face of heavy winds and sleet. H. B. Dwight[102] introduced a non-trial-and-error method for sag calculations with variable temperature, support heights, tension, and span length.

Carr found that the normal component of force from oblique winds on telephone wires is proportional to the square of the normal velocity component, using a wind tunnel operating at 30 to 90 miles per hour.

A very important problem is the ease of handling wires or cable of great length; it is nearly always imperative that the line be wound upon spools to facilitate "paying out" during installation. While this is only too logical for open-wire lines it becomes an important limitation on cables of all types. Standardization of components is essential to both low manufacturing cost and ease and reliability of construction. The initial cost, invariably a large sum, must be economically justified by the best of engineering foresight in terms of the probable use of the facility after installation; especially is this speculative with the introduction of new types of service such as was the case with trans-oceanic telephony, television, and picture transmission.

The cable or coaxial line differs very much mechanically from open-wire construction in that the insulation problem is entirely within the sheath and not generally subject to weather wear. Coaxial cables, which have come into wide use for radio-frequency purposes, may be kept filled with an inert gas such as nitrogen to minimize chemical disintegration of the metals and insulation. At radio frequencies the insulation problem becomes one of obtaining the very best in dielectric properties. Mechanical requirements are at cross purposes with electrical in the matter of insulator spacing within the tube, the electrical ideal of course being no support at all. The sheath itself, however, is probably the greatest worry. To allow winding on spools for installation convenience the diameter must be small, yet the band width (hence the number of available carrier channels or amount of television definition) increases with the square of its radius. Optimum wall thickness for copper (depending upon the diameter of the tube) may be too thin for handling and flexibility, while the lower conductivity of lead may cause its wall thickness to far exceed mechanical requirements; the difficulty has been circumvented by the use of copper sheath covered with lead. Chemical corrosion of the outside of the lead sheath is the same problem that has always worried audio-frequency telephone engineers. A good deal has been written on the subject, notably the article by Blomberg and Douglas[104] concerned with causes and mitigation of corrosion, using the method of localized currents, and another by R. M. Burns[105] explaining the mitigation for lead and lead-alloy cable sheathing. Aerial and underground plants are discussed and it is shown that embrittlement and not corrosion is the primary factor in cable life. Buried cable corrosion is influenced more by the soil than by the metal constituents. Charts are given for the various alloys.

Interference with previously existing services has been one of the primary factors governing the design of lines. Open-wire lines have admittedly better attenuation characteristics but cause no end of trouble with their unfortunate ability to pick up energy from the fields of other telephone lines nearby, or from a-c power mains not so nearby. Transposition has been used for years at audio frequencies but is at best an attack upon the effect and not the cause. At radio frequencies, however, the situation for two two-wire lines on the same poles is extremely severe; the electric and magnetic fields, no longer completely collapsing on the wires, radiate very effectively and the unwanted energy is promptly picked up by the other circuit.

The coaxial line by reason of its virtually complete shielding at all but very low frequencies (at which it is not used) is not bothered by and does not bother other circuits. In addition to this its ease of handling - for the designs now in use - and the absence of the exterior insulating requirement make it nearly ideal for radio-frequency line work. The cable manufacturing cost naturally is very much greater than for the open-wire type, but the installation cost is correspondingly low.

The Bureau of Standards has issued two handbooks governing the installation of transmission lines and obligations to other services. One[106] is concerned with the safety rules for installation and maintenance of supply and communication lines, both overhead and underground, especially where crossing other utilities such as railroads, pipes, etc., giving clearances and strengths of construction. The other[107] gives rules for the selection and installation of wood poles for overhead electric lines.

Transmission of Energy

Direct-current and audio-frequency line characteristics are important to the designer of radio-frequency transmission lines, first because the equations now used to describe the behavior of most radio-frequency transmission lines were developed out of the expressions already in use at audio frequencies, and second, any service intending to use an r-f line may immediately be confronted with the possibility of using an existing line which has been designed for low-frequency use. For example, telegraph lines, when sufficiently short to transmit a sharp-cornered Morse dot are often found satisfactory for the lower radio frequencies, and the same may apply to very short telephone lines. However, the inductively loaded line is out of the question on account of the low-pass-filter characteristic caused by the series inductances.

Next higher in the frequency spectrum lies carrier telephony in which it is only in recent years that the frequency limit has gone above the neighborhood of 50 kilocycles - in the lowest region of radio frequencies. No extensive account of the development of carrier-current telephony will be included in this paper, as another is in preparation by Robinett and Whitcomb[108] covering the entire history. Conception of the "new high frequency multiplex telephony" was first published in 1912 by Stone[109] but was not perfected and put into use until about 1918. A very extensive and intensive paper on carrier telephony and telegraphy was that of Blackwell and Colpitts[110] in 1921. Practical problems showing the desirability of avoiding the use of the same frequency in both directions were discussed by Rose[111] in 1923, and Slaughter[112] the following year described a complete system, giving both pictures and schematic diagrams.

Ilgenfritz[113] in October 1934 discussed the regulation of open-wire lines at carrier frequencies, which has a distinct bearing upon their use at higher frequencies. The attenuation and a-c resistance functions of frequency, the weather effects upon resistance and conductivity, insulator leakage, temperature effects, ice, and salt fogs are taken up, and it is pointed out that many small variations add up to a very important total fluctuation, necessitating the use of a pilot channel for compensation. The article was followed in the same issue by Terry's[114] description of pilot-channel-controlled automatic gain regulation. The apparatus functions every fifteen seconds, and selectors hold the attenuation constant within two decibels over the entire operating frequency range.

The use of carrier circuits for transmission of broadcast programs over open-wire lines with a carrier frequency of 42.5 kc and an audio channel of 35 to 7500 cps was taken up by Hodgson, Ralph, and Jacobson[115] in January 1935.

Frequency stability at ultra-high frequency is often obtained by the use of transmission lines for high-Q tanks.

Wide-band transmission is a method of recent years; most previous services have required a line to carry a high frequency but only a relatively narrow band at that frequency. In general it may be said that if a line is good for one frequency, it is good for all below it -- with a few exceptions of course. The cable extensively described on page 22 and shown in Figure 47 has extended the number of channels from the previous twelve at 60 kc to 240 at 1020 kc according to Strieby[116] in a July 1935 article. Twelve voice channels, each 4000 cps wide are "moved up" in frequency by using them to modulate twelve carrier frequencies ranging from 60 to 108 kc. Then by combining these modulated carriers for use in the modulation of a "regular" carrier, whose frequency may be anywhere between 108 and 1020 kc, twelve voice channels have been impressed upon a single carrier. Twenty such arrangements are used, each "regular" carrier in succession being sufficiently above the last in frequency that overlapping does not occur. Wide-band transmission is of course essential to television definition, and the cable en toto has been used for this purpose. A public "talking test" of the cable was described in Telephony[117] for December 1936.

Isolation of Equipment

The compactness with which it has become necessary to build both transmitting and receiving equipment has for some years required thorough shielding of all components having electric or magnetic fields. Yet it is necessary to transfer energy from one unit to the next by means of a conductor; it must be insulated from bodies at other potentials and yet not allow the field of one to get into the other's territory. In cases of extreme differences of power level between components - which is not at all unusual - the associated conductors must be separated electrically from each other to reduce undesirable stray fields. In all such cases transmission lines are required and used to transfer the energy between components. Being very short compared to the wavelength and size of conductor, the attenuation is negligible, but shielding or balance is not. Coaxial lines have generally replaced the twisted-pair link of a few years ago for this purpose.

Probably the most obvious use of a line for isolation purposes is the installation between transmitter and antenna. Twisted-pair feeders were apparently used first by J. L. Reinartz at a time when 100 meters was the shortest useful wavelength, but an unbelievably small amount of literature has been published on the subject.

Roosenstein[118] in 1930 described the use of quarter-wavelength lines for matching feeders to the antenna; he also discussed "Rudenberg's transformation lines" offering a gradual impedance change. The ARRL handbook[119] described various feeding systems for amateurs. Feeder construction is discussed emphasizing the use of light feeder spacers and transposition blocks.

Sterba[120] in 1931 pointed out the tremendous effect of transmission lines upon an array fed by them. Currents induced by the antenna flow in the open-wire line as if it were a single conductor. Half-wave or quarter-wave lines may be added to one side of the line at the proper point to cancel out such pickup. A half-wave two-wire line shorted at a quarter-wave distance from one end is an effective short to in-phase currents and a high impedance to the desired line currents. Many other interesting tricks are pointed out in the article.

Graham[121] in January 1935 gave a detailed account of the factors influencing the choice of twisted pairs for radio frequencies, and showed proper methods for connecting to the transmitter and antenna.

Several articles appeared on the shifting of antenna directivity by means of the manner in which power is fed into a multiple-wire feeder. The first of these, by Reinartz[122] in February 1935 gave patterns and test results . for a 64-foot flat-top with three forty-foot feeders. Griffin[123] in October described a system connecting two vertical antennas by a two-wire line, with a remote switch to control directivity. Another short note by Sanders[124] in November gave a variation of Reinartz's methods, and an article by Pool[125] in May 1936 showed a novel combination of two antennas and a three-wire feeder achieving choice of three directions. Keen[126] in March suggested the use of transmission lines to compensate for limited antenna length.

In an article primarily intended for transmitter development above 300 mc, Lindenblad[127] told of his experiences with lines at those frequencies. He used shorted quarter-wave lines as "metallic insulators" at corners and bends in the line. Stranded wire was found conducive to high loss and twisting made for high inductance. A half-wave filament line was used to effectively ground the filaments of the tubes - which on reflection is a novel and certainly a good way to do it. Using a line of number four B and S gauge wires one inch apart, he found the attenuation to be 0.037 nepers/100 feet. Efficiencies were 91.4% for 100 feet and 40.7% for 1000 feet. Larger conductors were found better for longer line and cleanliness had little effect.

Figure 77 shows a recent use by Trevor and Dow[128] of coaxial lines for transferring power at 177 Mc between a balanced power amplifier and its balanced antenna by means of a 100-foot concentric line without destroying the final balance. The 83-ohm IKLM is made of one-inch and quarter-inch sizes of copper tubing with insulators every quarter-wavelength, and has an efficiency of 95%. The plate circuit of the transmitter A has the two parallel tank circuits B and C which are coupled to D at a current loop of one of the tanks. Conductors RG and QF are the same length to provide voltages 180° out of phase at F and G. The HJI portion of concentric line is exactly a half wavelength and the two lengths FH and GI are equal but of any convenient length, which makes H and I of opposite phase. The wave from H, however, in transversing HJI becomes in-phase with that from GI and they add and then travel along the line IKLM. Element IK is an impedance transformer matching the 75-ohm loop to the 83-ohm line. The same scheme is used at the antenna as shown. The article shows the use of other resonant circuits in the oscillator and amplifier.

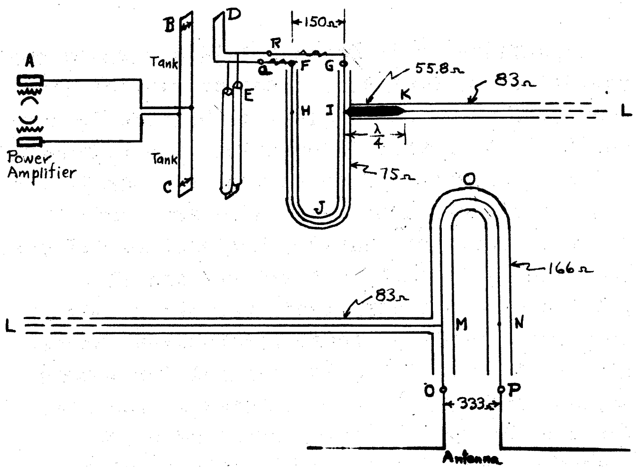


Figure 77 - USE OF CONCENTRIC LINES AT ULTRA HIGH FREQUENCIES FOR UNBALANCED TRANSMISSION OF POWER BETWEEN BALANCED TERMINAL IMPEDANCES.

The increasing use of directional arrays covering acres of real estate and the necessity of their location on relatively flat terrain free of surrounding hills has made the employment of transmission lines a necessity. Not only do feeders having some radiation of their own distort the directional pattern but the comparatively strong field of the antenna is picked up by such a line, causing severe unbalance and pattern-distorting reflections. The use of buried coaxial lines on the other hand has entirely removed these limitations, substituting for them the very few of its own which have been previously discussed.

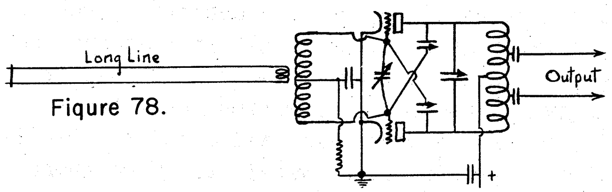
Substantially the same problems appear for receiving antennas. Lines may be smaller since they carry practically no power, but they must be designed to avoid affecting the directional pattern if one is intended. Most important, the line must be substantially incapable of picking up extraneous voltages along its length; the balance or shielding properties of a feeder are directly responsible for the amount of man-made interference reaching the receiver. Several noise reducing systems have been described in the articles of Johnson[129], Hatry[130], and Crossland[131].

Rettenmeyer[132] described equipment for operation, in the Waldorf-Astoria Hotel and other large establishments, of thousands of radio receivers from a single antenna. The units following the 600-foot coaxial cable from the antenna may be either passive or amplifying, and several 50 to 100-ohm lines are described. The line installed in the hotel is a #18 untinned copper wire covered in turn by cotton wrapping, eighth-inch rubber wall, cotton wrap, copper braid, and lead sheath; its attenuation is 0.3 db per 100 feet at 1000 kc and its impedance is 65 ohms.

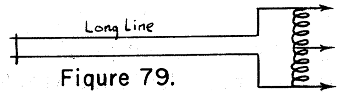
The use of twisted pair not primarily intended for operation at radio frequencies is covered very well by Harris[133] in an article appearing in March 1936. Wheeler and Whitman[134] in October include a page of transmission-line discussion in their article on doublet antenna design. A special high-grade rubber-insulated 115-ohm line is mentioned having 2 db attenuation per hundred feet at 18 Mc. Fabric coverings are also discussed.

Frequency Discrimination

Since its accidental trial in 1929 by RCA engineers, the quarter-wave line has been used very extensively in place of the tuned circuit. When used in grid or plate circuits of oscillators to minimize undesirable frequency variation, the circuits have variously been called "long-line control", "resonant-line control", and "distributed-constant-network frequency control", all meaning that any line may be used as a high-Q tank circuit. The term *frequency discrimination* is chosen to include their use in both oscillating circuits and in passive or pure amplifier circuits -- as anti-resonant circuits.



Fiqure 78.



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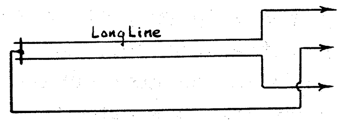
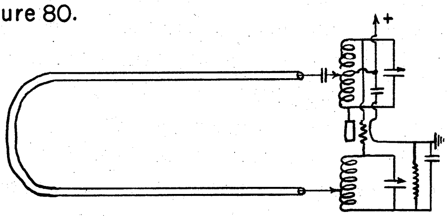


Fiqure 80.



Fiqure 81.

LONG LINE FREQUENCY CONTROL CIRCUITS.

Karplus[135] in 1931 showed oscillators tuned in the grid or grid-plate circuit by a transmission line at 600 megacycles (50 cm. wavelength). Conklin, Finch, and Hansell[136] published a paper in the same year giving the circuits they had found successful at the RCA transmitter laboratories. It is pointed out that there are in general two methods of using such lines. One, characterized by Figures 78, and 79, and 80 uses the line as a resonant circuit. The other Figure, 81, employs an electrically very long line for plate-to-grid coupling, in which small changes of frequency in the plate circuit produce a magnified phase-angle change at the far end of the line (in the grid circuit), which brings the oscillator almost back to the original frequency. The lines of Figures 78, 79 and 80 are all-power-factor (high-Q) circuits of large power-dissipating capacity and are used directly in place of the crystal which would be used at lower powers or lower frequencies. A circuit of this type was demonstrated at 13,390 kc at station WIK with two 20-kw water-cooled tubes and a line of #6 wire 10.25 wavelengths long. Line temperature and frequency characteristics are given. The "aperiodic long-line control" of Figure 81 may of course employ an open-wire

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* Reference given in the article.

line instead of the coaxial depicted. Such an oscillator actually holds better frequency control for heavier loads. A line is "least effective at the point of maximum oscillator efficiency". A given line will hold the oscillator to within approximately the same number of cycles regardless of frequency.

Schaible[137] in August 1931 conducted a mathematical investigation into the limitations and relations governing the line control of frequency. He concluded that although the line may be placed either in the grid or plate circuits, the latter gives better control, and that a combination of both is the most desirable. The use of line-shunting tank capacitances was found unsatisfactory, and the mathematical equations of lines used as inductances in both circuits did not lend themselves to concrete analysis. The degree of control increased with line length. The optimum ratio of separation to wire diameter was found to be forty, beyond which there is excessive radiation.

R. A. Hull in February 1935 described a practical circuit using coaxial lines in the grid circuit of tubes operating at 60 and 120 megacycles. Another interesting article by Hansell[139] appeared in August 1935 but its contents are thoroughly covered in a paper, of which he is a joint author, discussed in the following paragraph.

Hansell and Carter[140] in April 1936 published a very good paper showing the effectiveness of line control at high frequencies. The form of line found best consists of coaxial tubes, one completely enclosing the other. The Q is thus kept high by preventing radiation and extraneous coupling. Copper or aluminum may be used. The use of lines more than a quarter wavelength long does not influence Q but does increase the power-handling capabilities; however it is suggested that larger diameters be employed for this purpose. Equations (61 to 64), already given above from another author, are restated in somewhat different notation, wherein

a = outer radius of inner conduction in cm,

b = inner radius of outer conductor in cm,

f = frequency in cps,

I = current in line at point of maximum current,

and λ = wavelength.

Then $L = 2 \times 10^{-7} \ln \frac{b}{a}$ henrys per meter, (85)

$$C = \frac{10^{-9}}{18 \ln \frac{b}{a}} \text{ farads per meter,} \qquad (86)$$

$$Z_o = 60 \ln \frac{b}{a} \text{ ohms,} \qquad (87)$$

and $R = 41.6 \times 10^{-7} \sqrt{f} \left(\frac{1}{a} + \frac{1}{b}\right)$ ohms per meter (88)

For a line constructed of copper

$$\alpha = \frac{R}{2Z}, \qquad (89)$$

and for a tuned line of copper the power loss is given by

$$W = \frac{I^2 R \lambda}{8} \text{ watts per quarter wavelength.} \qquad (90)$$

The oscillatory energy per quarter wave is

$$VA = \frac{\pi f L I^2 \lambda}{4} = \frac{I^2}{16 \pi f C}. \qquad (91)$$

Also $Q = \frac{VA}{W} = \frac{2\pi f L}{R} = \frac{1}{2\pi f C R}$, (92)

and the maximum Q obtainable from a copper line with a given value of b is

$$Q_{max} = \frac{1}{6.86 \times 10^{-4} \sqrt{\frac{\lambda}{b}}} = \frac{1460b}{\sqrt{\lambda}}. \qquad (93)$$

For the minimum attenuation the optimum radius ratio is

$$\frac{b}{a} = 3.6 \qquad (94)$$

The maximum voltage gradient is

$$\frac{E}{a \ln \frac{b}{a}}. \qquad (95)$$

The radius ratio giving smallest maximum voltage gradient for a given maximum voltage and value of b is

$$\frac{b}{a} = 2.72, \qquad (96)$$

and the radius ratio giving minimum voltage gradient for a given oscillatory energy and value of b is

$$\frac{b}{a} = 1.65. \qquad (97)$$

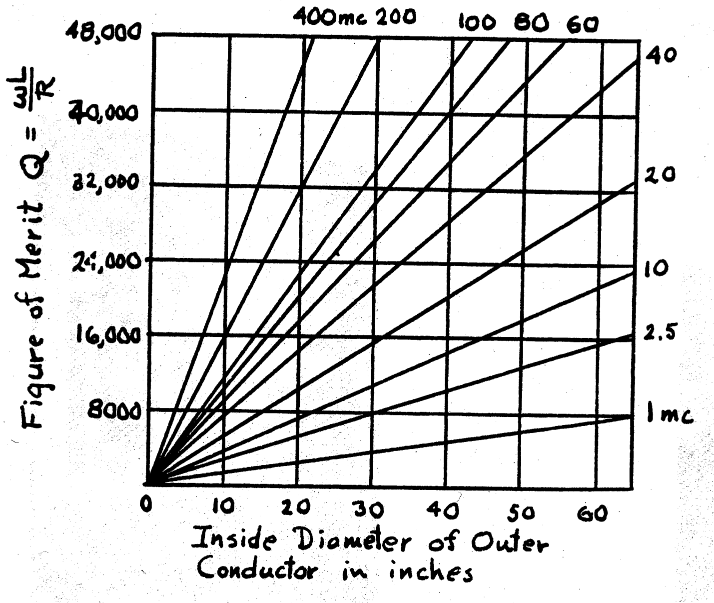

Figure 82 - FIGURE OF MERIT AT OPTIMUM RADIUS RATIO.

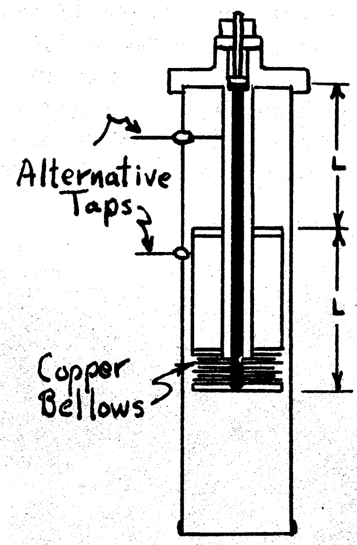

Figure 83 - TEMPERATURE-COMPENSATED ADJUSTABLE FREQUENCY SHORT LINE.

For equal thicknesses of inside and outside conductors the radius ratio for greatest energy storage per unit mass of copper is

$$\frac{b}{a} = 4.68, \qquad (98)$$

and for maximum impedance of a line having a given b (for instance in use as an insulator) the radius ratio becomes

$$\frac{b}{a} = 9.18. \qquad (99)$$

The Q of a line is proportional to the square root of frequency, inversely proportional to the square root of resistivity of the material, inversely proportional to the square root of wavelength, and proportional to the diameter of conductors so long as the radius ratio is held constant. For constant radius ratio the maximum allowable oscillatory energy is proportional to the square of diameters*.

Figure 82 shows Q as a function of frequency and of b, the inside diameter of the outer conductor. A table of temperature coefficients is given and it is found that the frequency-temperature coefficient is very nearly equal to the mechanical expansion coefficient. Figure 83 shows a convenient form of line in which the length of the inner conductor is held constant by means of an invar rod; the general effect of expansion caused by heating is to change other dimensions correspondingly, which results in the same frequency for the complete tank circuit. The use of two radius ratios as shown materially shortens the physical size of the apparatus and it will be noted that the portion of large radius ratio is effectively an inductance and the lower section a capacitance. Two other temperature-compensation methods are described. The simplest circuit is that of Figure 84; the regeneration control (or neutralizing capacitor) is adjusted for minimum feedback for stable operation, allowing the line to have fullest frequency control. Other methods of coupling to the line are described. For extreme frequency stability such

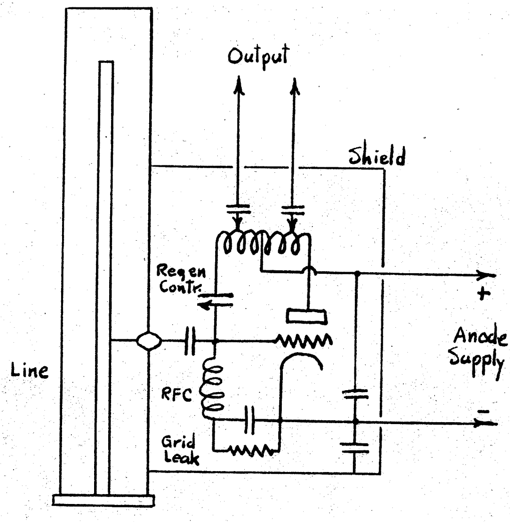

Figure 84 - SINGLE TUBE LINE CONTROLLED OSCILLATOR.

an oscillator should be followed by one or two stages of amplification, using doublers for higher-frequency output and secondary oscillators for very low-frequency output to keep the line size down. Figures 85, 86 and 87 are typical examples of line-controlled installations.

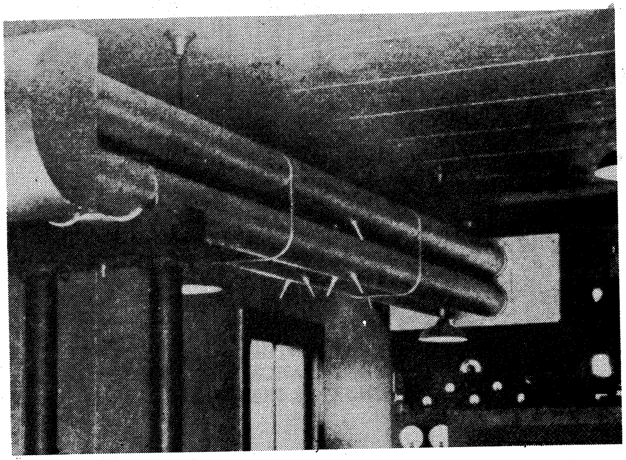

Figure 85 - LINE CONTROLLER MASTER OSCILLATOR FOR COMMERCIAL TRANSMITTER WQO - 6725 KILOCYCLES.

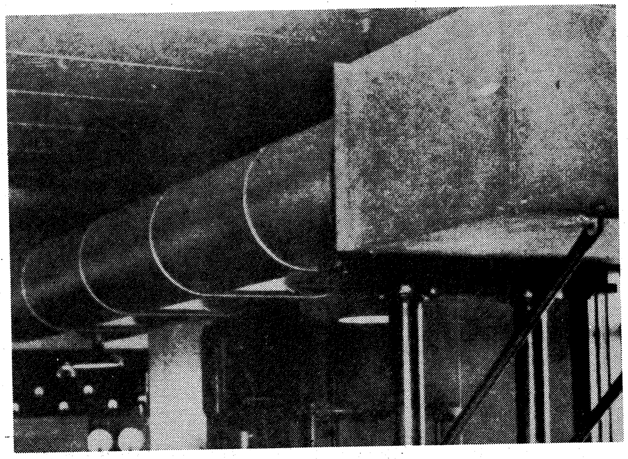

Figure 86 - LINE CONTROLLER MASTER OSCILLATOR FOR COMMERCIAL TRANSMITTER WHR - 13420 KILOCYCLES.

Hull[141] again in September 1936 built a one-meter "trough" transmitter, the trough being a coaxial line employing a round inner conductor and square outer conductor minus one of its four sides. Such a system has fairly good efficiency and a low external field. Although it cannot have extremely high Q's, its convenience makes it very desirable for general line control. A Western Electric Type 316-A tube was used.

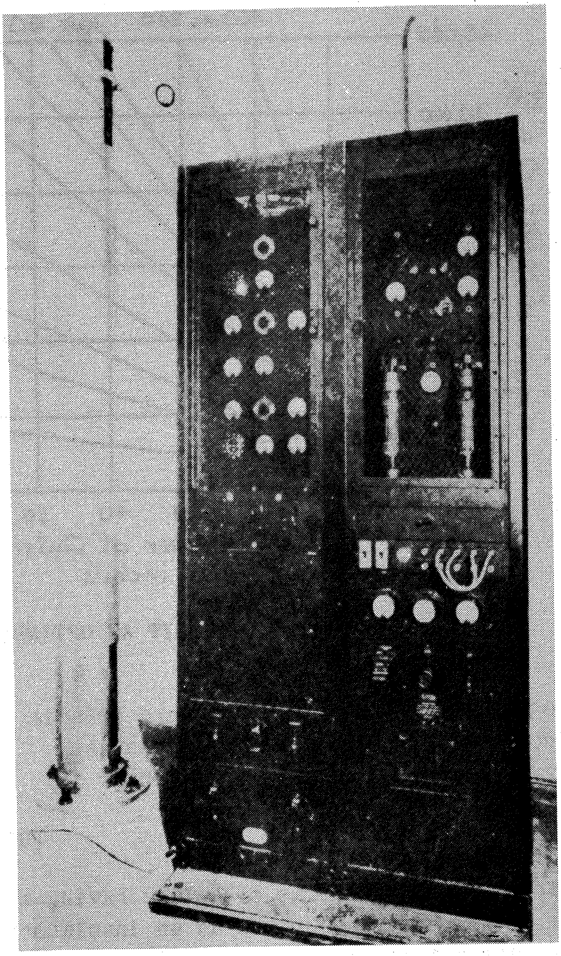

Figure 87
LINE CONTROLLER TRANSMITTER ON ROOF OF RCA BUILDING, 30 ROCKEFELLER PLAZA, NEW YORK CITY, OPERATED AT 25.7 MEGACYCLES OR HIGHER.

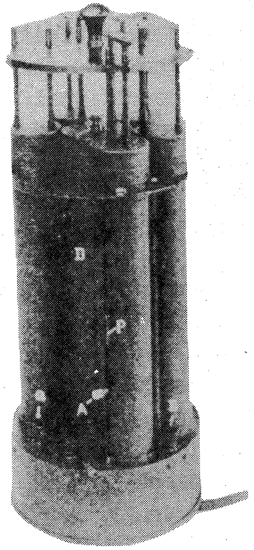

Figure 88 - A FOUR STAGE CONCENTRIC-LINE TUNED RADIO RECEIVER FOR 175 MEGACYCLES TO 300 MEGACYCLES (1.71 TO 1.0 METERS).

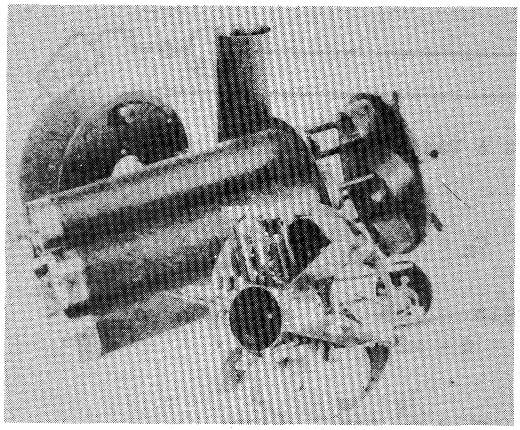

Figure 89 - THE CONCENTRIC LINE RECEIVER SHOWING METHOD OF ASSEMBLY.

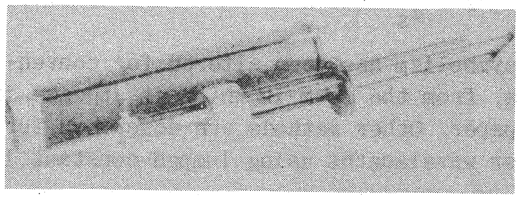

Figure 90
ONE OF THE CONCENTRIC LINES SHOWING TUNING PLUNGER AND SLIDING ANTENNA INPUT TERMINAL.

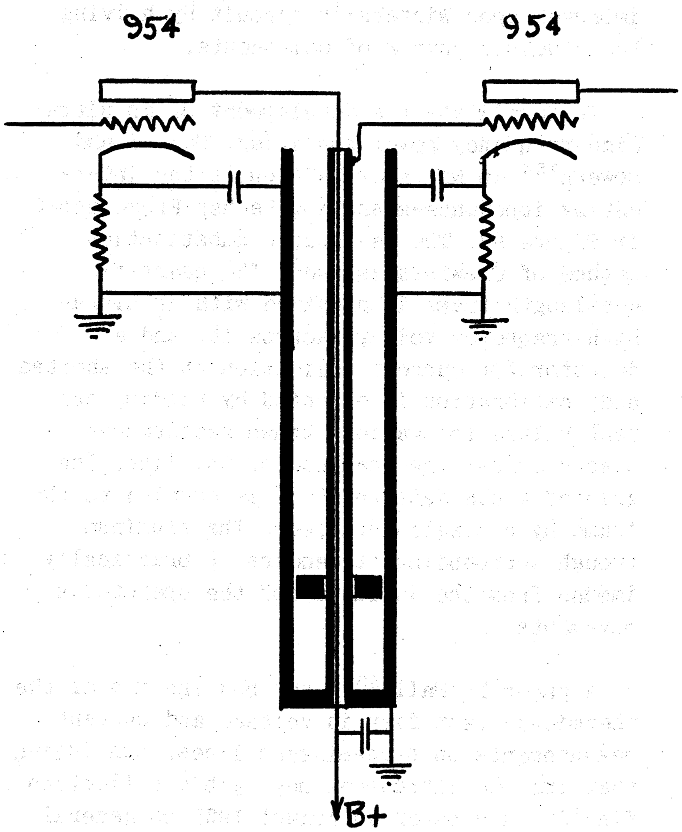

Figure 91 - COAXIAL LINE AS INTERSTAGE COUPLER.

of the line is of course free to swing at radio frequency corresponding to the plunger position, the grid or plate-to-ground impedance being nearly all within the acorn tube itself.

Sanders[144] in August gave a description of a 56-megacycle resonant-line plate-controlled oscillator.

Zottu[142] in October built a much modified concentric line in which the outer conductor is replaced by three equilaterally spaced rods. Convenience, again, is its chief asset.

The very interesting infernal machine of Figures 88, 89 and 90 is the brain child of F. W. Dunmore[143] and is actually a four-stage tuned-radio-frequency receiver for ultra-short waves employing concentric-line interstage coupling and acorn tubes. The receiver covers a useful range of 175 to 300 Mc, and has a gain of 2, 6, and 16 per stage at 300, 200, and 100 Mc respectively. Radius ratio is chosen for maximum impedance rather than selectivity. Figure 91 shows the electrical features. Unity coupling is afforded by the capacitance between the plate lead and the inner line tube which connects to the succeeding grid. The upper end

Measurements

Transmission-line phenomena being one aspect of oscillating energy, it is only natural that calibrated lines should be used for the measurement of wavelength or frequency. C. R. Englund[145] in 1928 explained the accurate use of Lecher wires (another name for the open-wire line used generally in connection with frequency measurements) for frequency measurement. The influences of dielectric dissymmetry, non-linearity, size of conductor, of indicating instruments, and framework upon accuracy were investigated. End correction is discussed. The investigation carried on at 38 to 75 megacycles. Takagishi[146] in 1930 studied the conditions for occurence of the double-hump phenomena quite apart from that due to excessive coupling, and Field[147] the same year studied the use of the neon glow tube with shorting bar for more accurate node determination.

Mohammed and Kantebet[148] in 1931 reviewed briefly the pioneer work of Lecher, Lodge, and

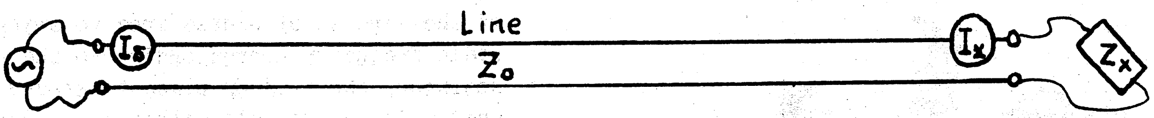

Figure 92 - IMPEDANCE MEASUREMENT USING A TRANSMISSION LINE.

Blondlot, the non-sine-wave correction of Hund, the work of Dunmore and Engel in 1923, and other later work. It is stated that distribution of current is controlled by system dimensions unless the length is an exact multiple of a half wavelength and bridged at the far end. Then it is likely to be complex and humped.

Hoag[149] in 1933 gave the mathematical treatment, experimental circuit, and test results for Lecher-wire measurements of frequency at ultra-high frequency. Using a wavelength of 64 cm, he gives an analysis of the system with matched input and shorted output, leading to an equation for determining ultra-high frequencies to three significant figures and independent of end effects. The equation may also be used to get velocity of propagation in other metals.

A section of Winters'[150] seminar paper is devoted to Lecher-wire measurement of frequency, giving a few of the transmission line equations already written down and explaining the general procedure used in locating nodes and antinodes. Reference is made to Field's[147] paper.

Transmission lines, having a characteristic impedance dependent only upon the geometry of their construction, are naturally suited to use in the measurement of impedance. J. W. Labus[151] in March 1931 used a quarter-wave line in the circuit of Figure 92, at 15 kc for this purpose. The line may be any number of quarter wavelengths long, but is practical only for wavelengths up to about 150 meters. The absolute value of the unknown impedance is

$$|Z_X| = Z_o \frac{I_S}{I_X} \text{ ohms} \qquad (100)$$

but to get resistive and reactive components it is necessary to parallel Z_X with a capacitance C_o and use the expressions,

$$R_X = \frac{2qZ_o}{\sqrt{2q^2(b_1^2 + b_2^2) - q^4 - (b_1^2 - b_2^2)^2}} \qquad (101)$$

and $$C_X = \frac{b_2^2 - b_1^2 - q^2}{2qZ_o}, \qquad (102)$$

wherein

$$q = \omega Z_o C_o,$$

$$b_1 = \frac{I_X}{I_S} \text{ for } Z_X \text{ alone,}$$

and $$b_2 = \frac{I_X}{I_S} \text{ for } Z_X \text{ paralleled with } C_o.$$

The symbolism has been altered for convenience, from the form in which it appears in the paper. Other methods are described for longer wavelengths using lumped-constant lines.

Winters[152] in May 1934 developed a lumped-constant network simulating a transmission line, to be used in the manner described by Labus[151] and Taylor for obtaining impedances at a frequency neighboring 50 kc. Kornetz[153] improved upon Winters[152] circuit by halving the required number of components.

To facilitate the development of an ultra-high-frequency power amplifier, Samuel and Sowers[154] in November 1936 built the interesting impedance-measuring Lecher Frame shown in Figure 93. The resistance-substitution method of Crawford is used. The quarter-wavelength frame is provided with an ultra-high-frequency voltage across it, and a detector for current indication at the shorted end; calibration is effected by reading current values for various known resistances placed across the open end of the line. The grid of a 955 detector tube is coupled to the frame by a single-turn loop. The aluminum trough surrounding it renders it practically immune from the influence of the operator's movements.

A paper by Hall[155] described the use of the thermionic rectifier in voltage and current measurements on transmission lines, concluding that its use introduces negligible reflection. King[156] in a paper in August 1935 on general measurements at ultra-high frequencies de-

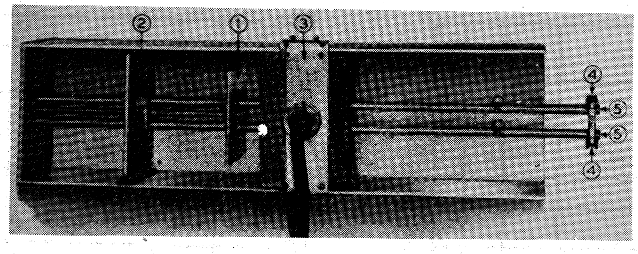

Figure 93 - PHOTOGRAPH OF IMPEDANCE MEASURING LECHER FRAME

1. Short circuiting bridge carrying detector tube and detector coupling coil.
2. Auxiliary bridge for breaking up unbalance currents flowing on the frame.
3. Input circuit. Note electrostatic screen between frame and input coil mounted on end of flexible transmission line leading to driving oscillator.
4. Clips carrying calibrating resistors.
5. Jacks into which plugs on amplifier input circuit fit.

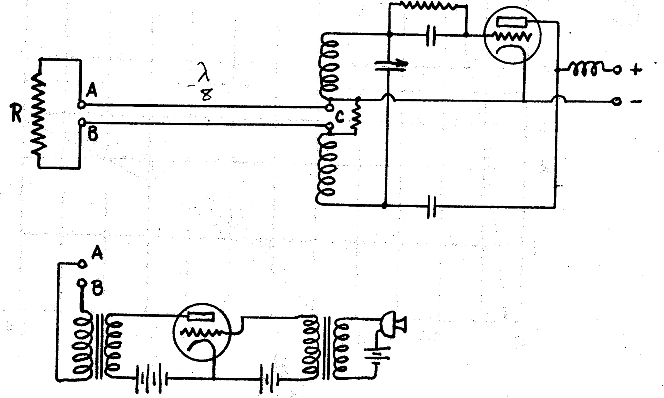

Figure 94 - HARTLEY OSCILLATOR WITH TRANSMISSION LINE MODULATOR, AND SPEECH AMPLIFIER TO REPLACE R.

scribed a convenient parallel-wire system for measuring current, voltage, reactance, resistance, permeability of wires, and the component frequencies in an ultra-high-frequency source. A complete solution is given for parallel wires bridged at each end by a general impedance, and the conditions for maximum current and voltage are stated. At ultra-high frequency, radiation occurs from the terminal impedances, and both radiation and ohmic resistance must be taken into account. The constants of a convenient frequency-measuring apparatus are supplied. Rhode and Bahnemann[157] gave in August 1931 a Lecher-wire method for obtaining potential differences at very high frequencies without upsetting the circuit being measured.

Frequency Modulation

Eastman and Scott[158] in July 1934 published a paper dealing with the use of an eighth-wavelength line for frequency modulation. In Figure 94 a change of resistance at AB, effected by voice modulation, causes fluctuation of line impedance at C according to the curves of Figure 95. These are according to the following equations derived in the paper:

$$R_C = Z_0 \csc 2\beta l \qquad (103)$$

$$X_C = -Z_0 \cot 2\beta l \qquad (104)$$

$$\beta l = \pm \cot^{-1}\frac{R}{Z_0}, \qquad (105)$$

in which the notation has been changed to conform with the previous discussions. Lumped lines were used for the experiments but it is shown that the same results were obtainable with twisted-pair lines which may be conveniently coiled and put into a very small space.

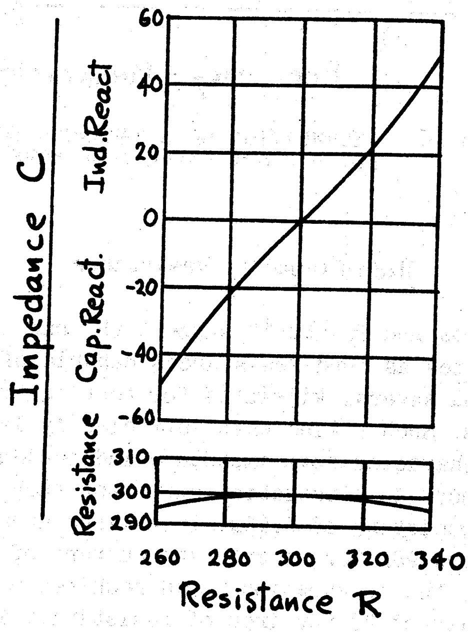

Figure 95 - IMPEDANCE FLUCTUATION AS A FUNCTION OF INPUT RESISTANCE.

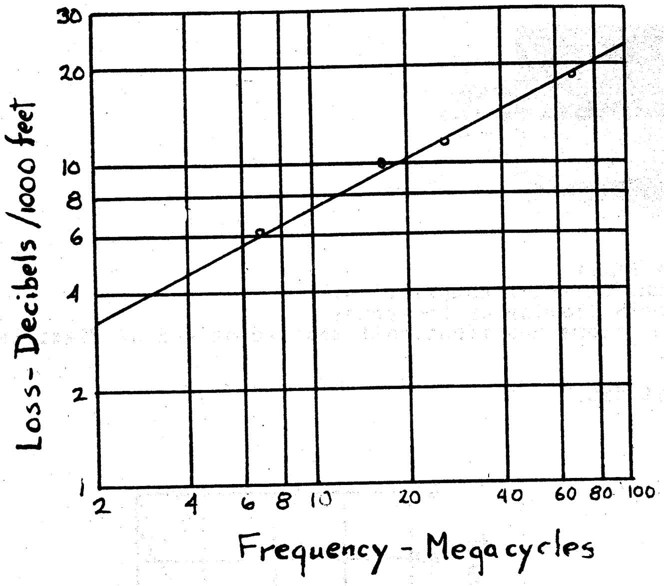

Figure 96 - ATTENUATION OF A BALANCED 600 OHM NO. 6 B & S GAUGE IRON WIRE LINE.

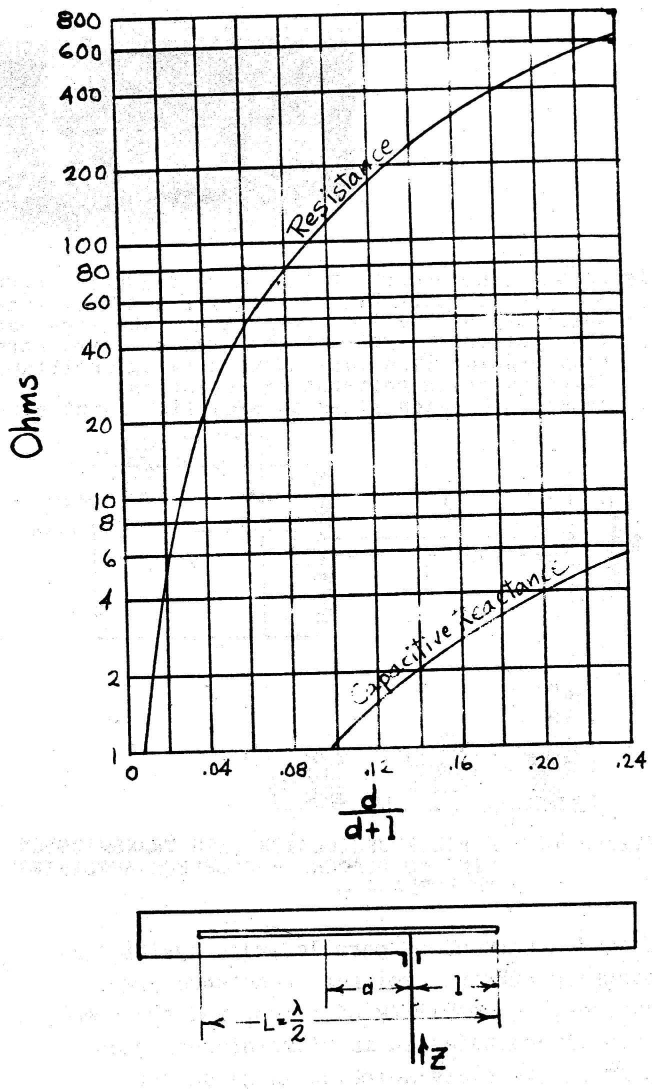

Figure 97 - USE OF THE CONCENTRIC LINE SECTION AS A VARIABLE CALCULABLE RESISTANCE AT HIGH FREQUENCIES.

High-Frequency Resistance

Sterba and Feldman[46] suggest the use of iron lines as load resistances capable of dissipating several kilowatts for testing transmitters. Such a line when sufficiently long has a characteristic impedance independent of frequency. An attenuation curve for such a line consisting of 0.162" iron wires appears in Figure 96. For an entering current of one ampere, the attenuation found requires a permeability of 92 for iron of resistivity 12,300 emu. A 1600-foot line has been used for several years to dissipate 15 kw at 20 mc.

Concentric-tube lines may be advantageously employed as high-frequency standard resistances, according to the same paper. Such a device is shown in Figure 97 together with curves giving its properties. The letter d represents the distance from the tap point to the center of the inner conductor, and *l* is the distance from the tap to the end. This type of r-f resistor can be designed for ruggedness and high power capability.

Impedance Transformers

Sterba and Feldman[46] describe two convenient methods of using short transmission lines as impedance-matching devices. In Figure 98 a quarter-wave line of characteristic impedance Z_0 is inserted in cascade between two lines of different impedances Z_1 and Z_2. Then equation (50) may be used to dimension Z_0 so that

$$Z_0^2 = Z_1 Z_2, \qquad (106)$$

in which case perfect matching occurs. Alternatively if a line already exists and operates mismatched, the standing waves may be observed and used with the charts of Figures 99 and 100. One chart is for application of the shunt correcting line to a portion of the regular line having lagging power factor, and the other is for leading power factor. Location of the current ratio on the charts gives both the length of the shunt and its position along the line.

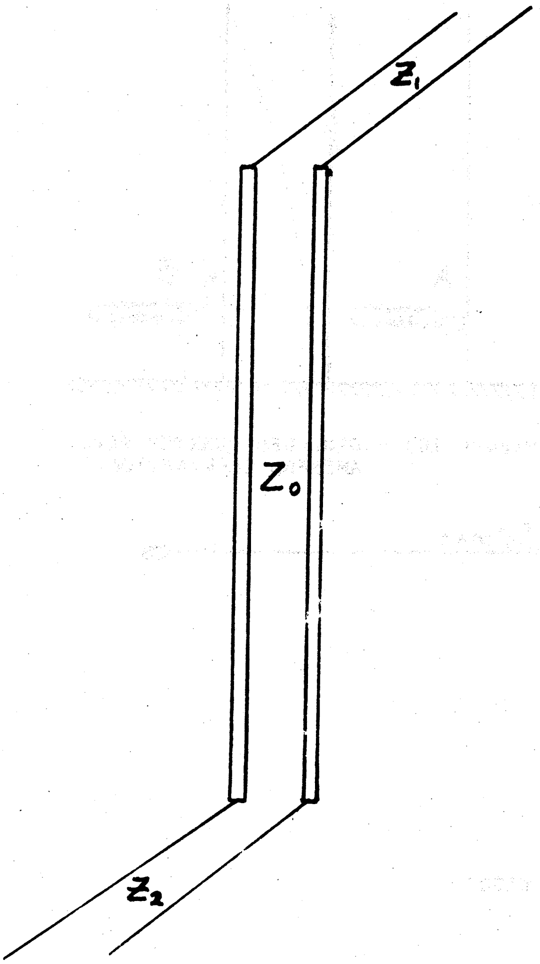

Figure 98 - USE OF QUARTER WAVELENGTH LINE AS IMPEDANCE TRANSFORMER.

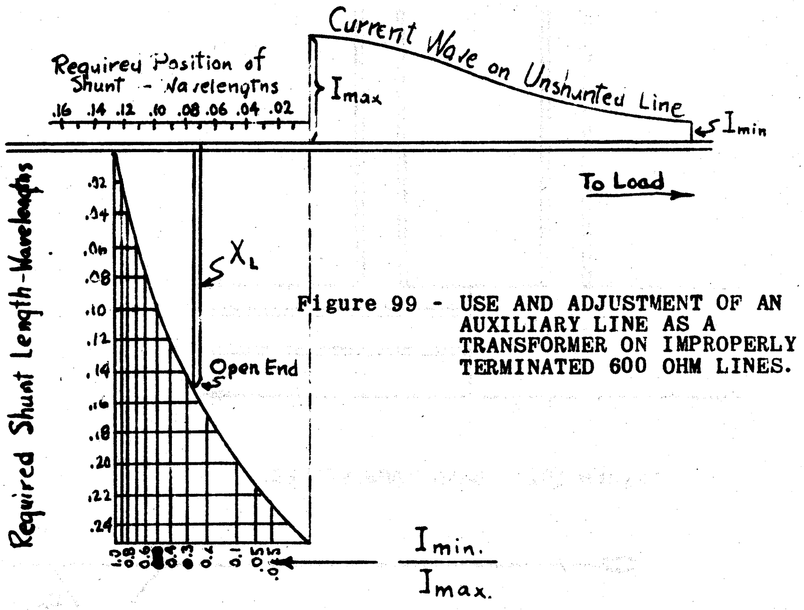

Figure 99 - USE AND ADJUSTMENT OF AN AUXILIARY LINE AS A TRANSFORMER ON IMPROPERLY TERMINATED 600 OHM LINES.

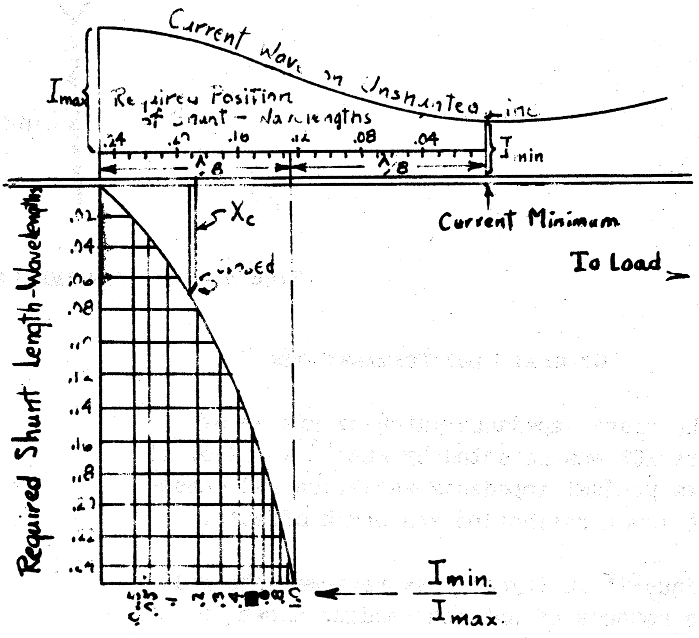

Figure 100 - USE AND ADJUSTMENT OF AN AUXILIARY LINE AS A TRANSFORMER ON IMPROPERLY TERMINATED 600 OHM LINES.

Filters

F. E. Terman[159] in July 1934 published an important paper giving an extensive summary of line equations and other material including the band-pass filter circuits of Figure 101 and 102. The first of these consists of two resonant lines loosely coupled together, each having a length of a quarter wave. Figure 102 has in the high side of the circuit two open-end quarter-wave lines, each acting as a series resonant circuit. The line drawn horizontally in the figure is not a quarter wave in length, and is proportioned to give either an inductive or capacitive reactance of the desired amount.

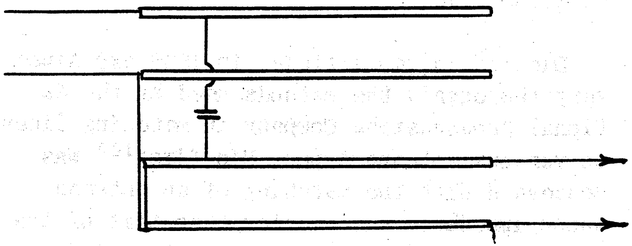

Figure 101 - BAND PASS FILTER.

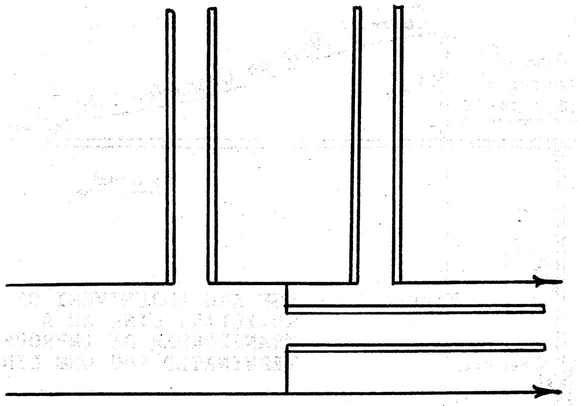

Figure 102 - BAND PASS FILTER.

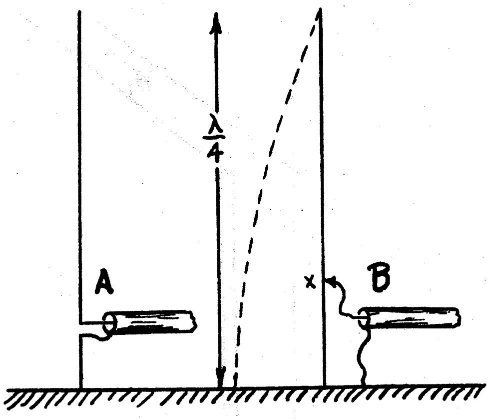

Figure 104 - GROUNDED QUARTER WAVE ANTENNA TERMINATION.

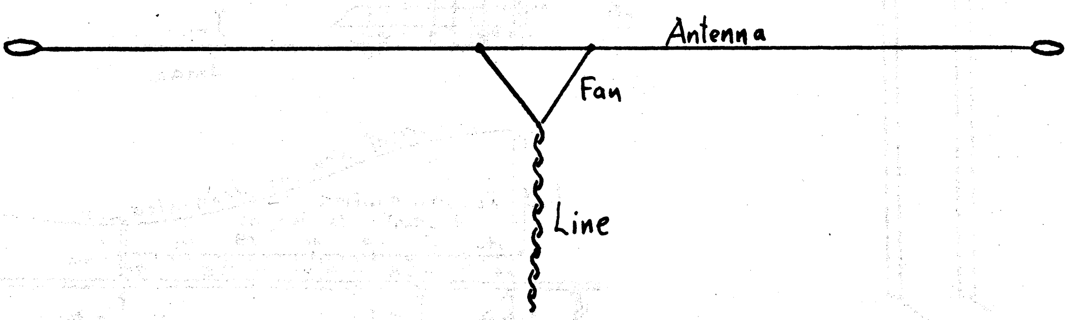

Figure 103 - FAN IMPEDANCE MATCH.

General Line Terminations

The "fan" impedance-matching system of Figure 103 was patented by RCA[160] in 1930. It allows gradual impedance variation and consequent lower reflection and brush effects.

Tinus[161] in August 1935 reviewed the well-known methods of antenna feeding shown in Figures 104, 105 and 106. In Figures 106-A and 106-B it should be noted that there is no impedance match and the line is behaving as an impedance transformer which must be properly terminated at the transmitter end according to equation (106).

Dietsch in two articles in 1936 explained very thoroughly the methods used by the National Broadcasting Company in matching lines to various antenna types. His first[162] was concerned with the matching of an antenna whose impedance was smaller than that of the line. Recent use of the coaxial line led to the very complete discussion in his second[163] article for the converse case. Although the

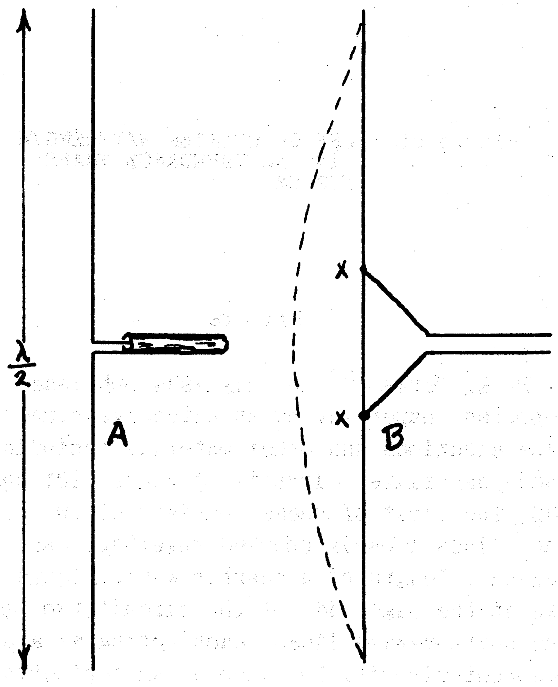

Figure 105 - HALF WAVE ANTENNA TERMINATIONS.

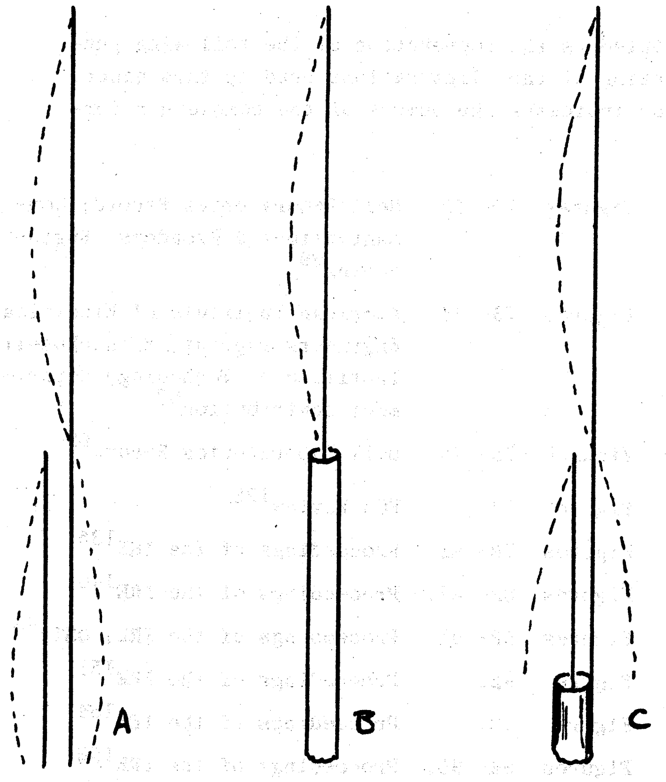

Figure 106 - HALF WAVE VOLTAGE FEEDS.

article is very complete and very good, it is concerned with proper matching networks to be used between the line and antenna and is not considered sufficiently pertinent to line design to be reviewed here in detail.

General Remarks

It is a bit difficult to set down any few particular problems regarding transmission-line theory or technique because the art is advancing so rapidly. It would seem the problems have been solved as fast as they have presented themselves. For a line of considerable length, nevertheless, there are a few general drawbacks which have been reduced to a great extent but are still much in evidence.

Attenuation has of course never been eliminated. For long lines this imposes a requirement of frequent repeater stations, requiring additional labor and maintenance costs. Engineers are of course continually reducing it by improved methods of design. Frequency discrimination imposes the additional requirement of equalizer circuits. Non-linear phase characteristics likewise demand phase compensation networks which never perfectly compensate. Stability of line constants for variable temperatures and other ambient conditions is an important factor in maintaining reliable service. Broad-band transmission has its own peculiar problems in keeping down extraneous noise and the design of filters which do not require too much territory on the frequency spectrum for broad cutoff.

ACKNOWLEDGMENT

The Radio Club of America gratefully acknowledges the cooperation of the following publications in granting permission for the reprinting of the illustrations used in this paper. The number following the name of the publication indicates the number of the complete reference in the Bibliography beginning on page 47.

Figures 1- 5.	Experimental Wireless & Wireless Engineer[26]
Figures 6-11.	Proceedings of the IRE; Bell System Technical Journal; Bell Mono[46]
Figures 12-15.	Proceedings of the IRE[50]
Figure 16.	Engineering Handbook of the National Association of Broadcasters[62]
Figures 17-18.	Communications & Broadcast Engineering[65]
Figures 19-25.	Proceedings of the IRE; Bell System Technical Journal; Bell Mono[46]
Figures 26-31.	Electrical Engineering; Bell System Technical Journal; Electrical Communications; Bell Mono[75]
Figures 32-33.	Engineering Handbook of the National Association of Broadcasters[62]
Figures 34-46.	Bell System Technical Journal; Bell Mono[78]
Figures 47-52.	Electrical Engineering[79]
Figures 53-59.	Proceedings of the IRE[86]
Figures 60-69.	Bell System Technical Journal; Bell Mono[88]
Figures 70- 72.	Bell Laboratories Record; Communications & Broadcast Engineering[89]
Figures 73- 74.	American Institute of Electrical Engineers reprint, Massachusetts Institute of Technology department contribution[97]
Figures 75- 76.	Bell Laboratories Record[99]
Figure 77.	RCA Review[128]
Figures 78- 81.	Proceedings of the IRE[135]
Figures 82- 87.	Proceedings of the IRE[140]
Figures 88- 91.	Proceedings of the IRE; QST[143]
Figure 92.	Proceedings of the IRE[151]
Figure 93.	Proceedings of the IRE[154]
Figures 94- 95.	Proceedings of the IRE[158]
Figures 96-100.	Proceedings of the IRE; Bell System Technical Journal; Bell Mono[46]
Figures 101-102.	Electrical Engineering[159]
Figure 103.	RCA, German Patent[160]
Figures 104-106.	Electronics[161]

Acknowledgment is made to Charles E. Dean and C.J. Le Bel for the many hours they spent in the compilation of the material for this survey.

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EDITOR'S REMARKS

The original manuscript by Mr. Winlund was submitted to the Radio Club for publication prior to World War II. The Publication Committee was hampered during the War, but when the pressure was relieved, work was resumed. In the meantime many additional references became available and were included in the bibliography.

Your present editor received the manuscript approximately one year ago. Because of the delay in publication, it seems advisable to add a few additional remarks summarizing the general trends in the use and development of transmission lines for the period from 1937 to 1951.

This summary touches only the high lights. A complete treatment would even tax the ability of a full-time editor.

The wartime standardization of coaxial cables, waveguides, and fittings facilitates rapid interconnection of all types of electronic equipment. Much of this work is summarized in Reference #680.

A new propagation mode of waves along single wires, referred to as "G-Strings", has been described by Rust[644], Goubau[646], and Coleman[657]. This mode is subject to increase in loss by adverse weather conditions; so it has many problems in common with open wire lines.

The demand for microwave equipment initiated by the wartime use of radar (see Vol. I of #680) has produced many interesting wave guide devices such as:
directional couplers[373, 435, 470, 565, 653, 680]; magic tee[557, 680]; hybrid rings[546, 558].

The coaxial equivalents of these devices have been combined with coaxial filters to produce diplexers[677], triplexers[677]; all used in feeding television sight and sound signals into a common radiating antenna.

Slotted lines are still the standby for accurate measurements above 50 megacycles[400, 454, 474, 601].

Tapered lines are widely used for broad band impedance matching[188, 255, 258, 274, 321, 399, 419, 439, 484, 500, 634, 671].

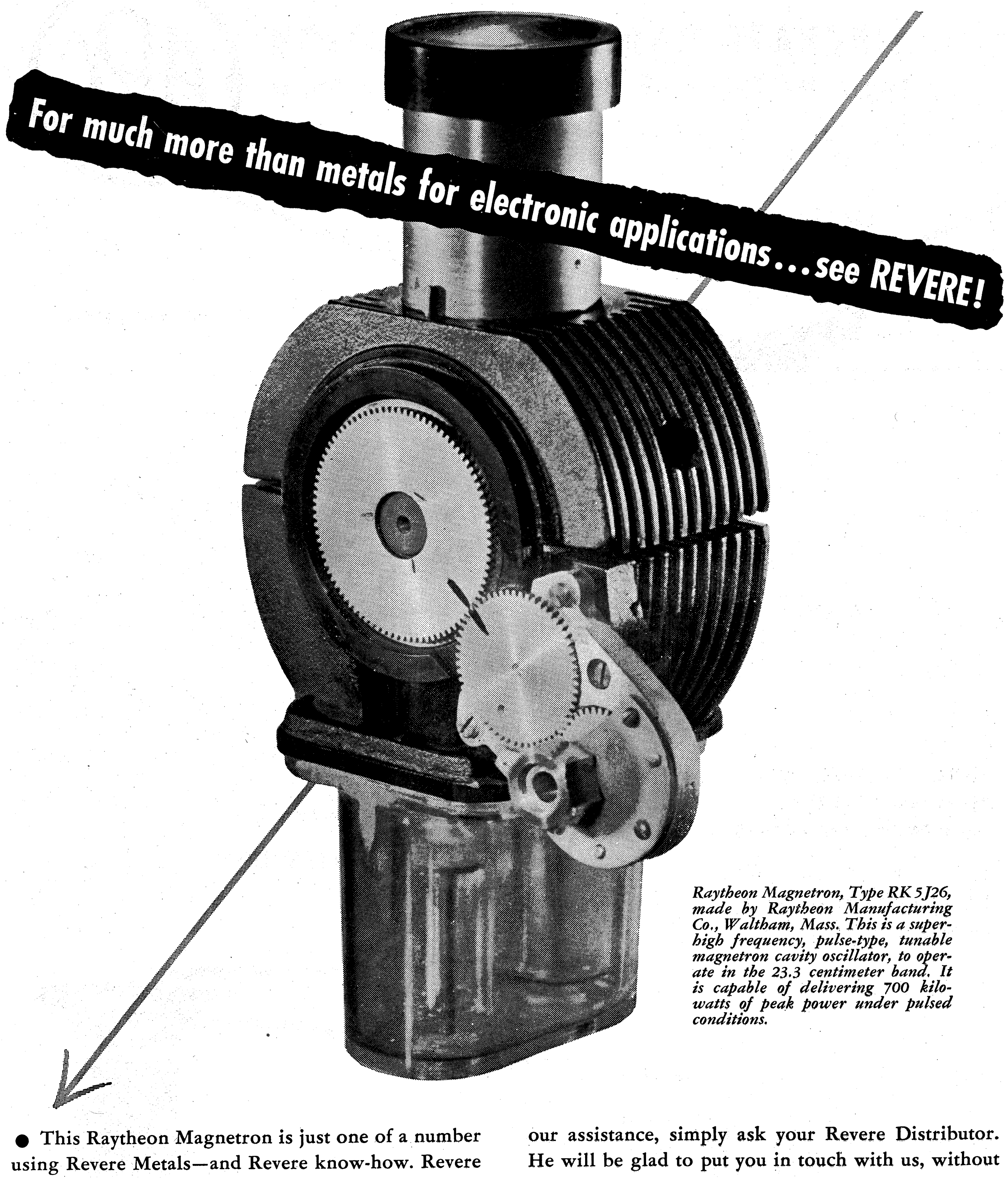
Delay lines[269, 356, 405, 411, 510, 663] are applied to precise time measurements not only in radar but also in loran equipment[680]. They are being widely used in airborne distance measuring equipment[681]. They are also used for temporary information storage in various calculating and computing machines. Precise timing circuits in television synchronizing signal generators utilize delay lines[683].

The wide use of wave guide below cutoff attenuators in modern test equipment was described originally in 1935 by Harnett and Case[582] and more recently by Briot[651].

The increase in the use of slotted lines has resulted in the development of a specialized set of coordinates and a slide rule by Smith[187, 254, 616]. Smith Chart plots of impedance and admittance are widely used today.

Every television-equipped home makes use of transmission lines today[406, 486, 487]. Television coaxial cable links are rapidly being supplemented by special microwave relays. The economics for coast-to-coast television seem to favor the use of microwave repeater chains[682]. In September 1951, the first coast-to-coast television broadcast carried the proceedings of the San Francisco Japanese Peace Conference with the new microwave relay installed by the A.T.& T. The World Series baseball games for 1951 were viewed by the largest audience in history through the use of this new relay system.

In a general sense one can consider the use of these narrow angle microwave relay systems to be a type of transmission line through space without need for conducting guides. Maintenance is zero for the transmission medium, and the maintenance of remote repeaters has been greatly reduced by the extensive use of telemetering and standby facilities.



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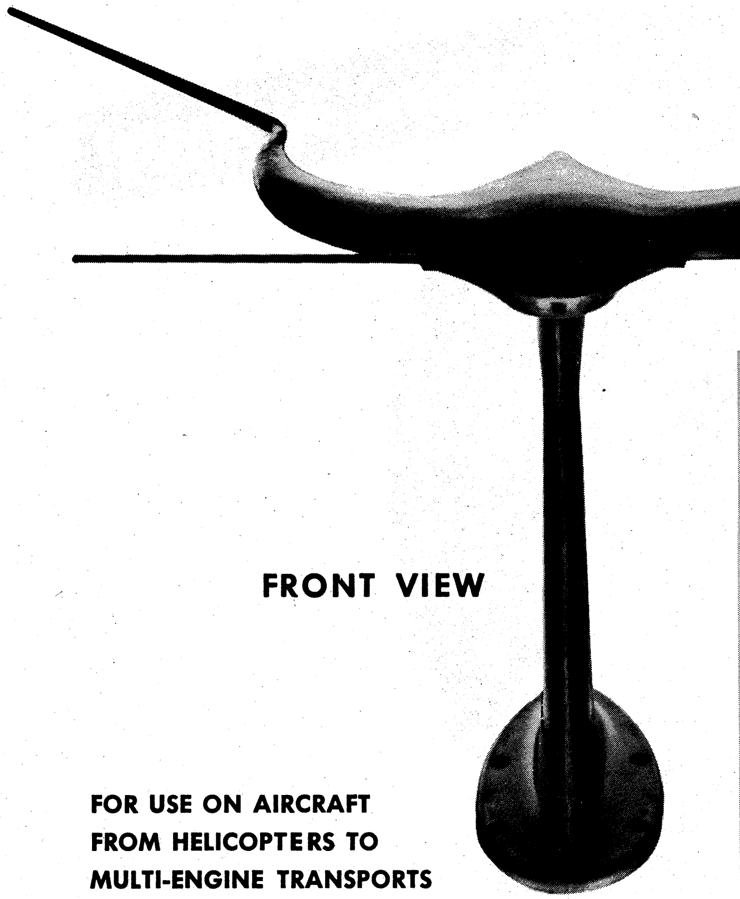

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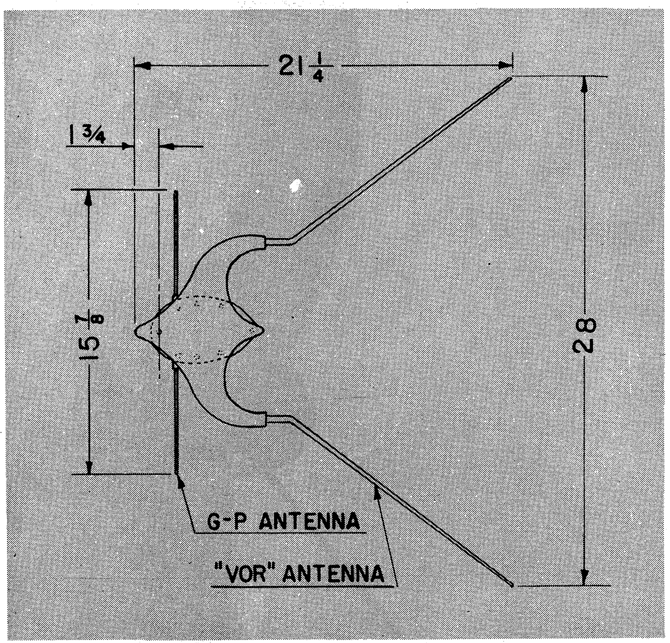

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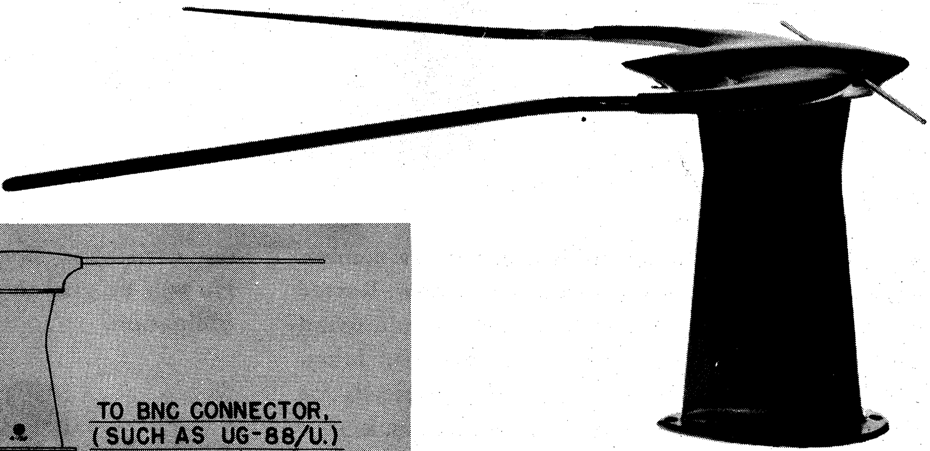

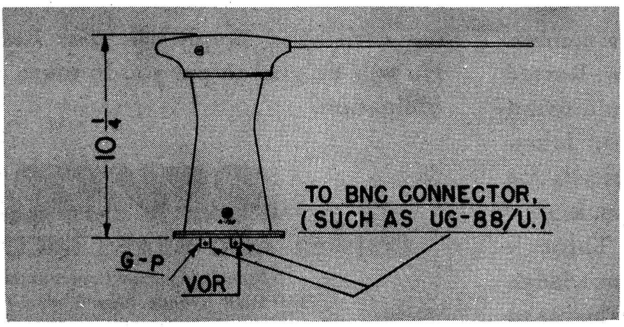

SIDE VIEW

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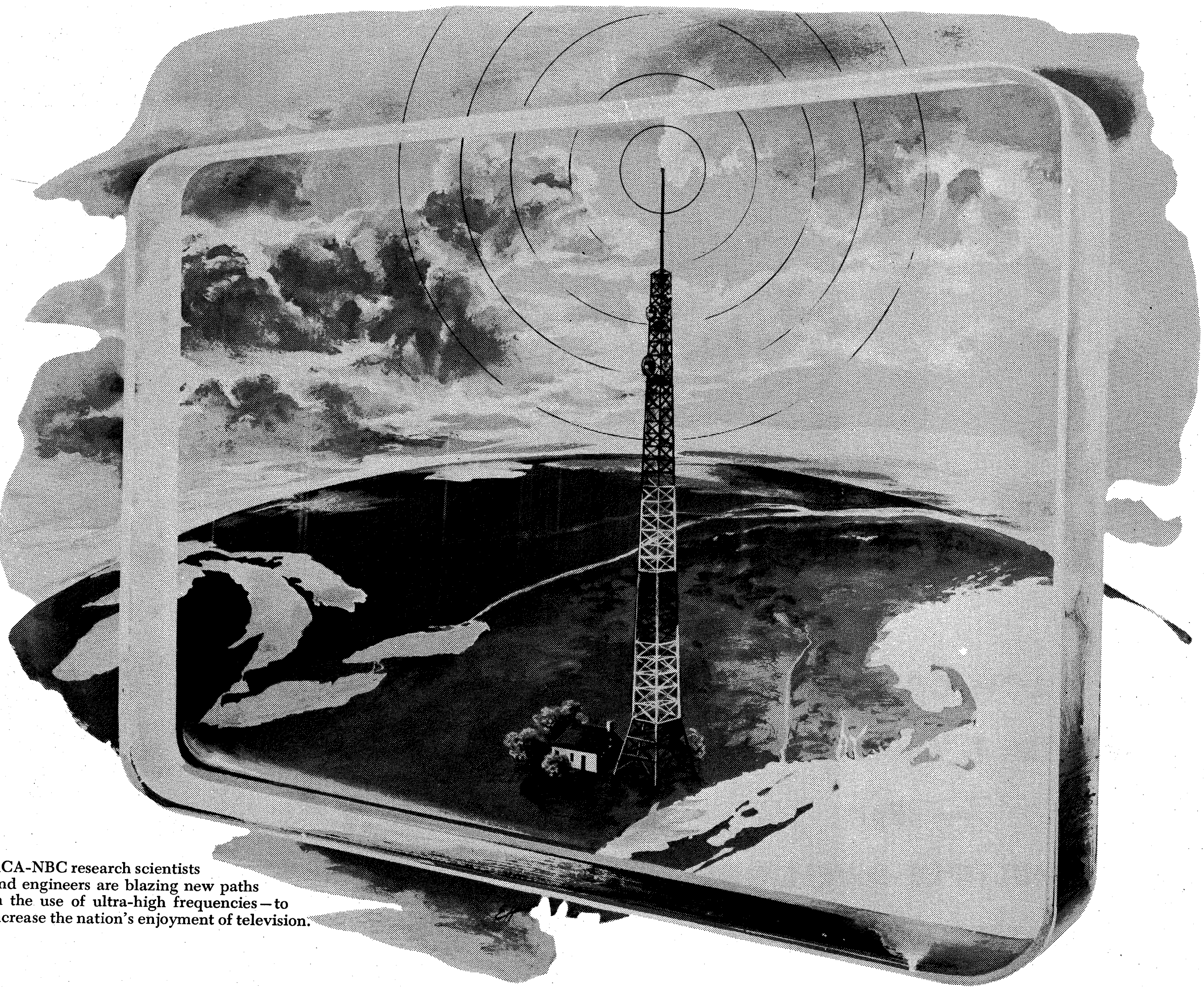

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Although television now reaches 45 million people in more than 12 million homes, thousands of communities are still too far from existing stations to be reached by *any* programs. Moreover, under present conditions, many cities with limited program service want, but can't have, additional TV stations.

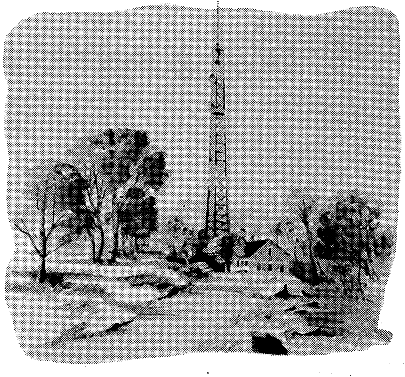
In preparation for the establishment of a country-wide television service, RCA has pioneered for many years in ultra-high-frequency (UHF) research.

Today—an experimental station built by RCA at Bridgeport, Conn., is supplying the practical experience and engineering facts needed to design the best UHF equipment—including transmitters, receivers, and converters. NBC programs on the air during the full broadcast day are used by RCA—and other manufacturers, too—for large-scale field tests.

From results of this pioneering, RCA engineers have determined that practical UHF equipment can be built to serve the public, and that present RCA Victor television sets can be readily adapted to give equally fine performance on both UHF and VHF.

* * *

See the latest in radio, television, and electronics at RCA Exhibition Hall, 36 W. 49th St., N. Y. Admission is free. Radio Corp. of America, RCA Building, Radio City, N. Y. 20, N. Y.

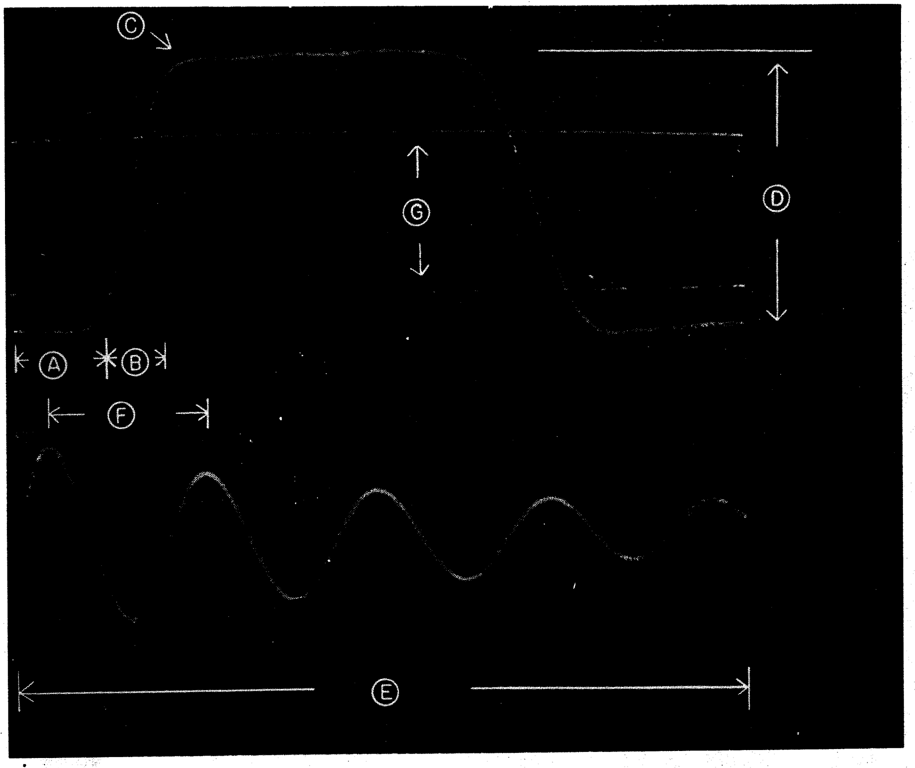


Built by RCA at Bridgeport, Conn.,—first UHF transmitter to operate on a regular schedule.

RCA

RADIO CORPORATION of AMERICA

World Leader in Radio — First in Television



a SINGLE OSCILLOGRAM demonstrates...

- **A. Signal Delay**
- **B. Transient Response**
- **C. No High-Frequency Overshoot**
- **D. Undistorted Deflection**
- **E. Sweep Linearity and Speed**
- **F. Time Calibration**
- **G. Amplitude Calibration**

Here, woven around the quantitative investigation of a 0.25 microsecond pulse, is a graphic account of the performance features which make the Type 303 an exceptionally fine, high-frequency cathode-ray oscillograph.

...IN THE NEW DU MONT TYPE 303

A. SIGNAL DELAY built into the Y-axis amplifier insures complete display of the steep pulse rise. As illustrated by a portion "A", the 10% point of rise does not occur until sometime after the sweep starts. Y-axis frequency response, of the instrument, includes the performance of the signal-delay line.

B. EXCELLENT TRANSIENT RESPONSE—wholly essential to the proper study of high-speed phenomena—is depicted by the rise time which is reproduced without appreciable degradation. A rise time of 0.01 microsecond, or greater, will be reproduced as a rise time not exceeding 0.033 microsecond.

C. NO OVERSHOOT is observed even on extremely steep wavefronts. The low-frequency response limit is a 3% slope on a 30-cycle squarewave. As shown on the frequency-response curve, there is no positive slope above the mid-frequency range. Since the response tapers off so slowly, the Type 303 is usable at frequencies beyond 10 megacycles. The synchronizing circuits will lock in sine-wave signals as high as 20 megacycles.

D. UNDISTORTED DEFLECTION provided by the Y-axis amplifier is 2 inches for unidirectional pulses. An equivalent undistorted deflection of 4 inches is available for symmetrical signals and may be positioned over the useful area of the cathode-ray tube. Even at the highest attenuation ratios, the Y-axis input is not frequency sensitive, as shown by the illustrated pulse which has been attenuated 4000 times. The direct-coupled X-axis amplifier of the Type 303 will provide over 5″ of undistorted deflection.

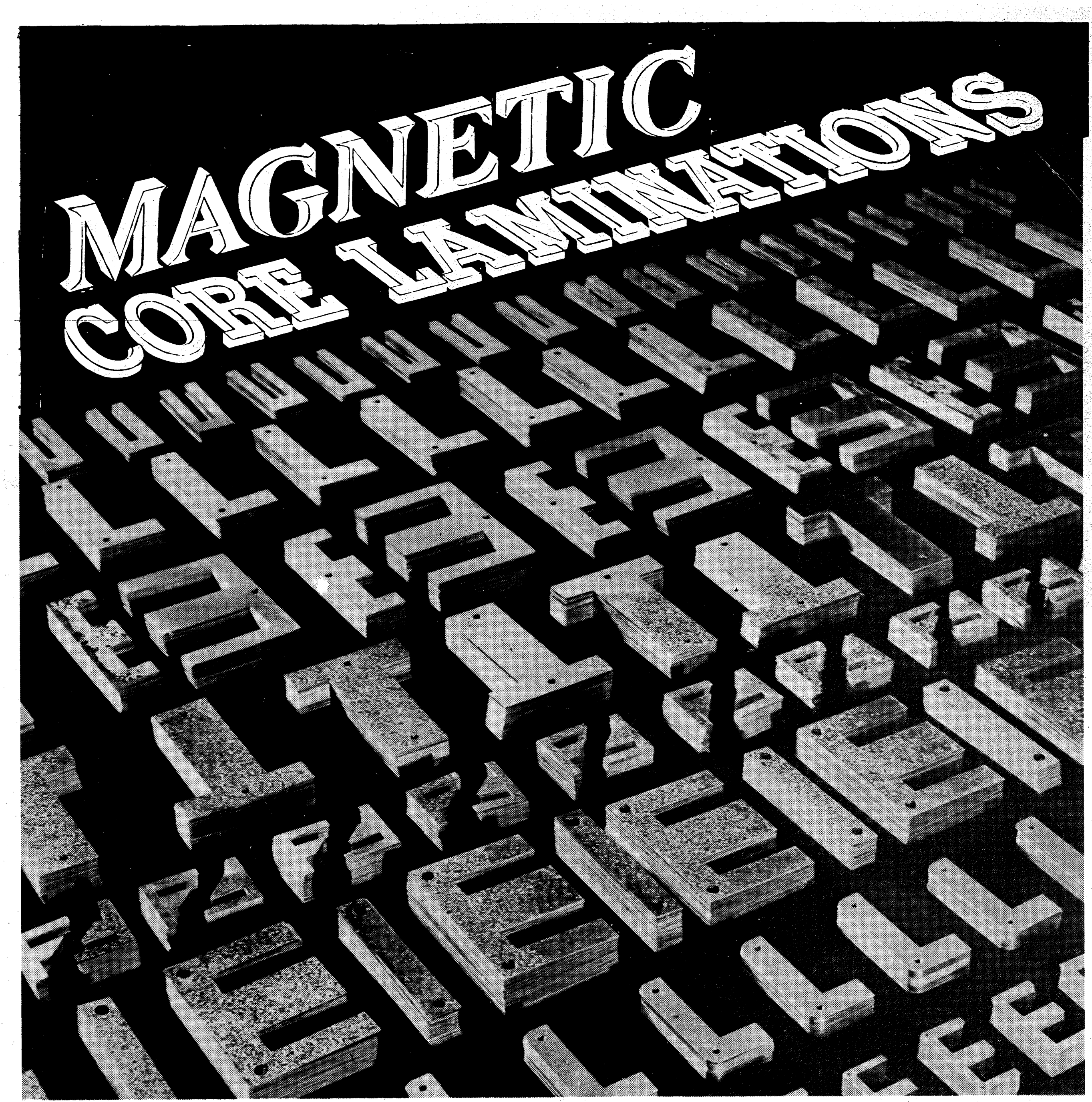
E. SWEEP SPEEDS available in the Type 303 make possible a presentation which is practical for qualitative and quantitative analysis of a pulse as short as 0.25 microsecond. Both driven and recurrent sweeps are continuously variable from 0.1 second to 5 microseconds. Through sweep expansion, sweep length is variable from a fraction of an inch to an effective 30 inches, any portion of which may be positioned on the screen. As shown above, even at the fastest sweep range, the sweep is extremely stable and linear. Notice the absence of jitter.

F. TIME CALIBRATION in the Type 303 is accomplished by substituting a damped sinewave for the signal. Double exposure by photographic recording of calibrating sinewave and signal provides a permanent quantitative analysis of the signal. In addition to the 10-megacycle signal shown above, calibrating frequencies of 10 KC, 100 KC, and 1 MC are also available. Accuracy of time calibration is within 3%.

G. AMPLITUDE CALIBRATION completes the precise, quantitative analysis of the signal. A built-in, regulated, voltage-calibrator provides peak-to-peak signals of 0.1, 1.0, 10, and 100 volts. Similar to time calibration, the amplitude calibrating square wave is substituted for the signal. Amplitude calibration is accurate within 5%.

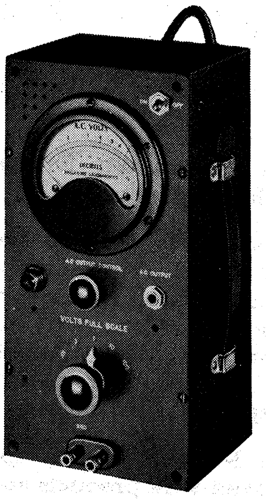
price $820.00 FOR COMPLETE DETAILS WRITE for bulletin TYPE 303

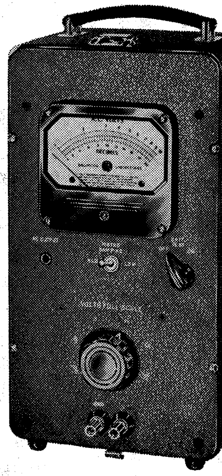
ALLEN B. DU MONT LABORATORIES, INC. Instrument Division 1000 Main Avenue, Clifton, N. J.

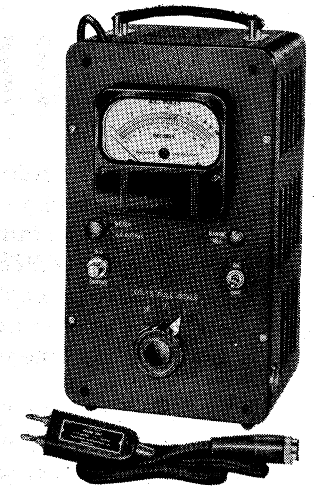

MAGNETIC
CORE LAMINATIONS

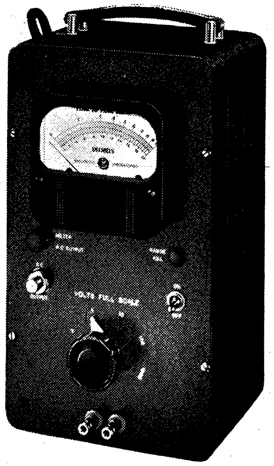
BALLANTINE

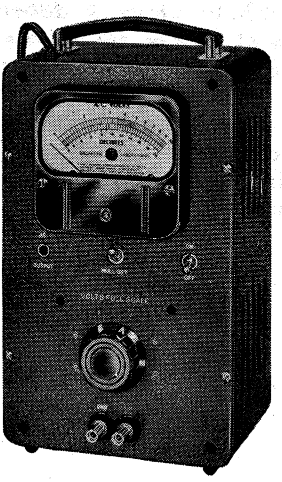
STILL THE FINEST IN ELECTRONIC VOLTMETERS


MODEL 300


MODEL 302B

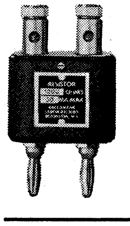

MODEL 304


MODEL 305


MODEL 310A

MODEL	FREQUENCY RANGE	VOLTAGE RANGE	INPUT IMPEDANCE	ACCURACY
300	10 to 150,000 cycles	1 millivolt to 100 volts	1/2 meg. shunted by 30 mmfds.	2% up to 100 KC 3% above 100 KC
302B Battery Operated	2 to 150,000 cycles	100 microvolts to 100 volts	2 megs. shunted by 8 mmfds. on high ranges and 15 mmfds. on low ranges	3% from 5 to 100,000 cycles; 5% elsewhere
304	30 cycles to 5.5 megacycles	1 millivolt to 100 volts except below 5 K C where max. range is 1 volt	1 meg. shunted by 9 mmfds. on low ranges. 4 mmfds. on highest range	3% except 5% for frequencies under 100 cycles and over 3 megacycles and for voltages over 1 volt
305	Measures peak values of pulses as short as 3 microseconds with a repetition rate as low as 20 per sec. Also measures peak values for sine waves from 10 to 150,000 cps.	1 millivolt to 1000 volts Peak to Peak	Same as Model 302B	3% on sine waves 5% on pulses
310A	10 cycles to 2 megocycles	100 microvolts to 100 volts	Same as Model 302B	3% below 1 MC 5% above 1 MC


MODEL 220
BATTERY OPERATED DECADE AMPLIFIER gives exact voltage gains of 10 or 100, permitting a corresponding increase in voltmeter sensitivity from 10 to 100,000 cycles.

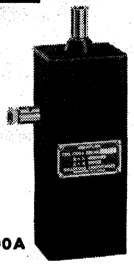


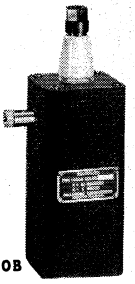
PRECISION SHUNT RESISTORS

Four different types of Precision Shunt Resistors, varying from 1 to 1000 ohms, permit the Voltmeters to be used to measure currents from 1 ampere to one-tenth of a microampere.

MULTIPLIERS

Five different types of Multipliers, whose input resistance varies from 5 to 40 megohms, permit the voltage range of the Voltmeters to be increased 10 or 100 times.


1300A


1300B

BALLANTINE pioneered circuitry and manufacturing integrity assure the maximum in

SENSITIVITY • ACCURACY • STABILITY

- All models have a single easy-to-read logarithmic voltage scale and a uniform DB scale.
- The logarithmic scale assures the same accuracy at <u>all</u> points on the scale.
- Multipliers, decade amplifiers and shunts shown above extend range and usefulness of voltmeters.
- Each model may also be used as a wide-band amplifier.

For further information, write for catalog

BALLANTINE LABORATORIES, INC.
BOONTON, NEW JERSEY,

Production Facilities
AT YOUR COMMAND

Precisely so. For Hudson Wire offers you the production capacities of FOUR wire plants that can draw and insulate FINE WIRES to your most exacting requirements, plus a wire cloth plant if your needs are in the WOVEN-WIRE category.

The main plant in Ossining, N. Y., draws Hudco fine wires in copper, brass, bronze, zinc, silver-plated, tin, phosphor-bronze, nickel-silver, cadmium and other metals and alloys. The plant at Winsted, Conn., specializes in Winco insulated wires. It handles the widest range of insulations such as enamel, cement-coating, silk-covered, celanese, glass-fibre, E-Z Sol, Formvar and other types. Also multiple-wire conductors such as Litz.

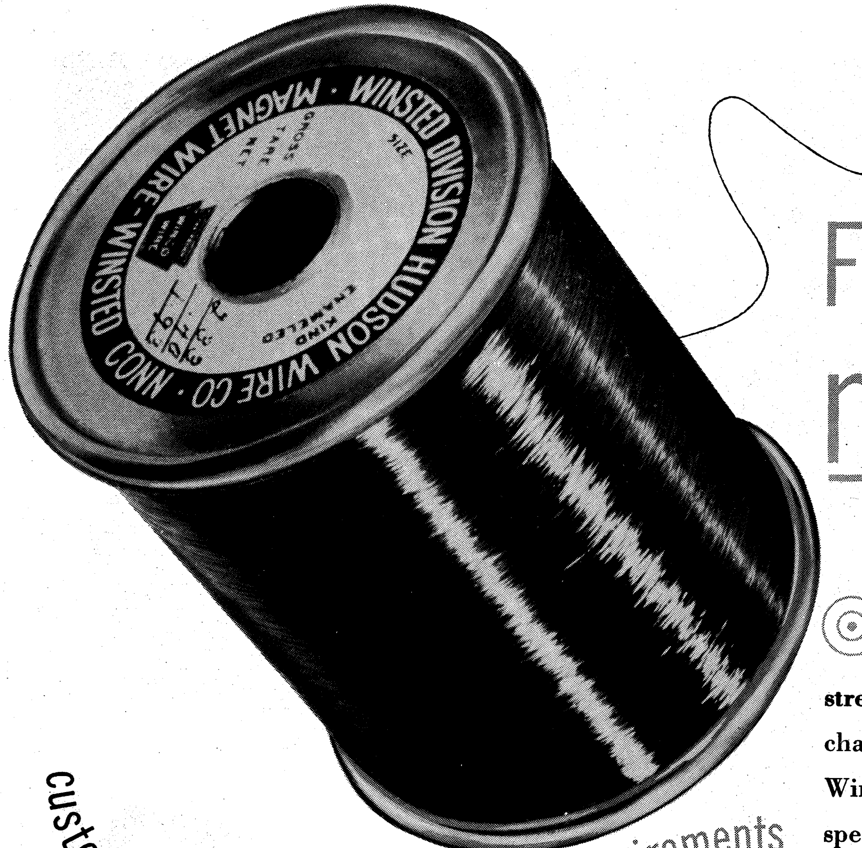
The plant at Pownal, Vt., handles certain drawing operations.

For the convenience of Mid-West wire users, a new plant is being set up at Cassapolis, Mich., for drawing and insulating fine wires. This plant insures virtual overnight shipments over a wide area, as well as substantial transportation economies.

The plant at Norwalk, Conn., specializes in woven-wire products such as wire mesh.

Large enough to prove a dependable source of supply, regardless of quantity requirements; small enough to provide individual attention and interest in all orders no matter how small or how large — such are the Hudson Wire production facilities at your command.

custom drawn
custom insulated
custom spooled
to your most exacting
requirements



FINE WIRE made finer

custom drawn ◎ custom insulated ◎ custom spooled

to your most exacting requirements

◎ Custom-made fine wire! Just specify the electrical properties, flexibility, tensile strength, laying speed, uniformity and other characteristics you must have. Our Hudson and Winsted Divisions will meet and maintain your specifications.

Uniformity of product is guaranteed by our critical supervision which guards against variations in size, structure and electrical values. Yes, "Fine Wire Made Finer!" That is why Hudson-Winsted fine wires are the first choice of electrical, radio, television and electronic manufacturers whose products are noted for reliability and long life.

◎ Tell us your wire problems and requirements. Our research, engineering and production facilities are at your disposal. Let us quote!

BARE WIRES
(HUDSON WIRE DIVISION)

Copper
Brass
Zinc
Tinsel
Tin
Nickel-Silver
Cadmium
Oxygen-free Copper
Silver-plated
Bronze
Phosphor-Bronze
Silver
Monel
Lead Wire
Fuse Wire
Specialty Wires

TEXTILE-COVERED WIRES
(WINSTED DIVISION)

Nylon
Celanese
Silk
Cotton
Rayon
Fiberglas

All available on bare or enameled wire, single or double covered.

INSULATED WIRES (WINSTED DIVISION)

MATERIALS
Copper
Aluminum
Iron
Copper-clad Steel

TYPES
Instrument Tubing
Litz
Multiplied and Twisted

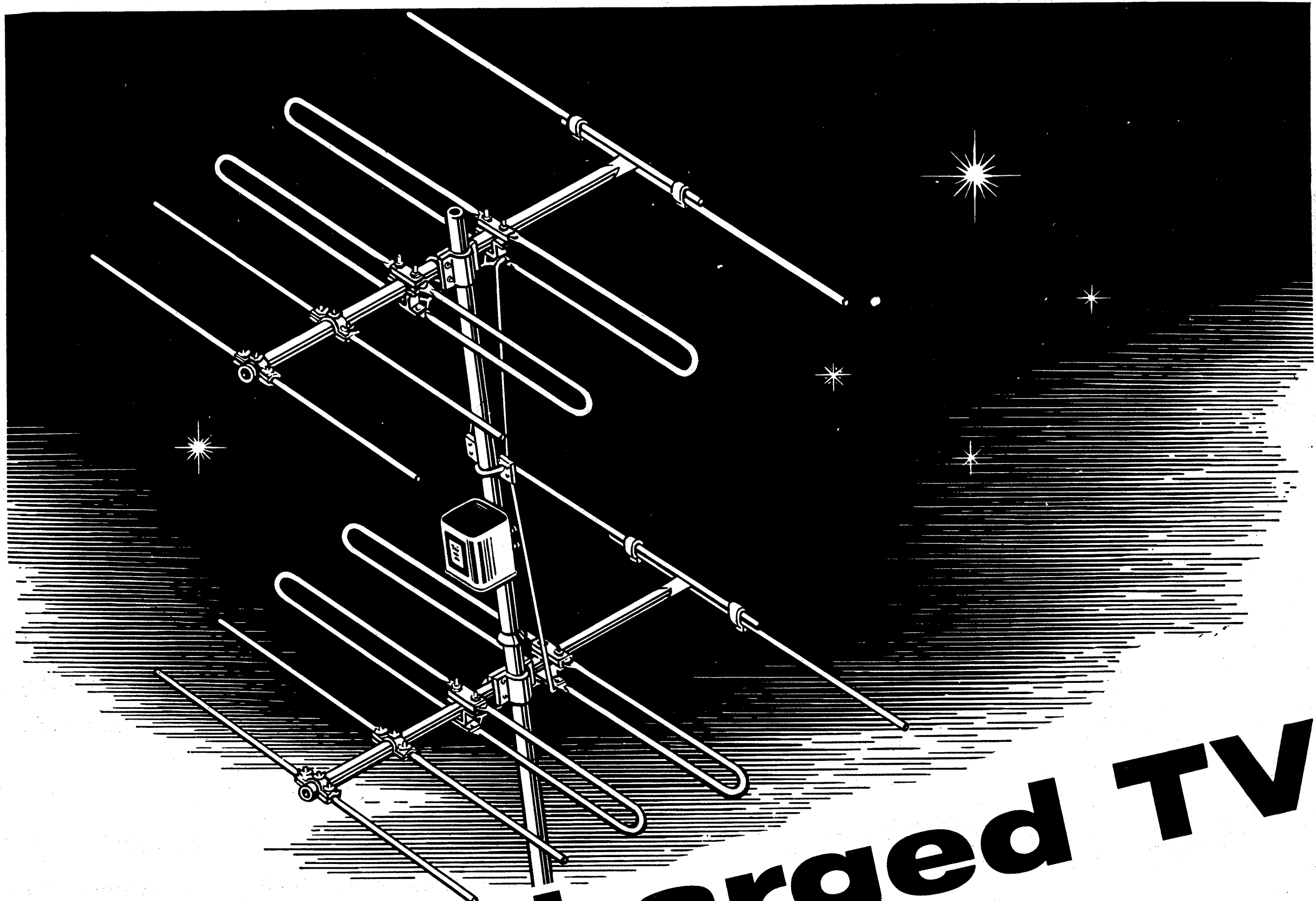
COVERINGS
Plain and Heavy Enamel
Formvar
EZsol (Liquid Nylon)
Cement-coated Enamel

FINE WIRES (HUDSON & WINSTED DIVISIONS)

Specializing in fine wires, custom-drawn and insulated, to critical needs—size, material, insulation. Your consideration is called particularly to the finest wire sizes —Nos. 44 to 50.

hudson wire company

general offices: ossining, n. y.   winsted division: winsted, conn.

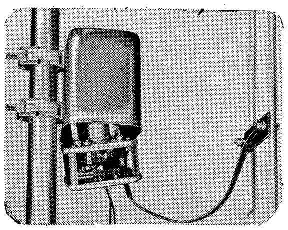


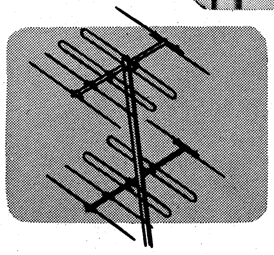
supercharged TV

The Finest-- When Results Count!

Here is the unbeatable Taco team. Made up of the Super 980 Yagi and the Antenna Supercharger this combination offers you 25 db gain *plus* a clarity of signal unparalleled due to the sharp directivity of the antenna and the sharp tuning of the Supercharger.

If it's results you're after -- try this combination!





Taco Antenna Supercharger
A power amplifier mounted up at the antenna. Provides powerful gain with maximum signal-to-noise ratio. Can be used with any good yagi-type antenna, but performs better with the . . .

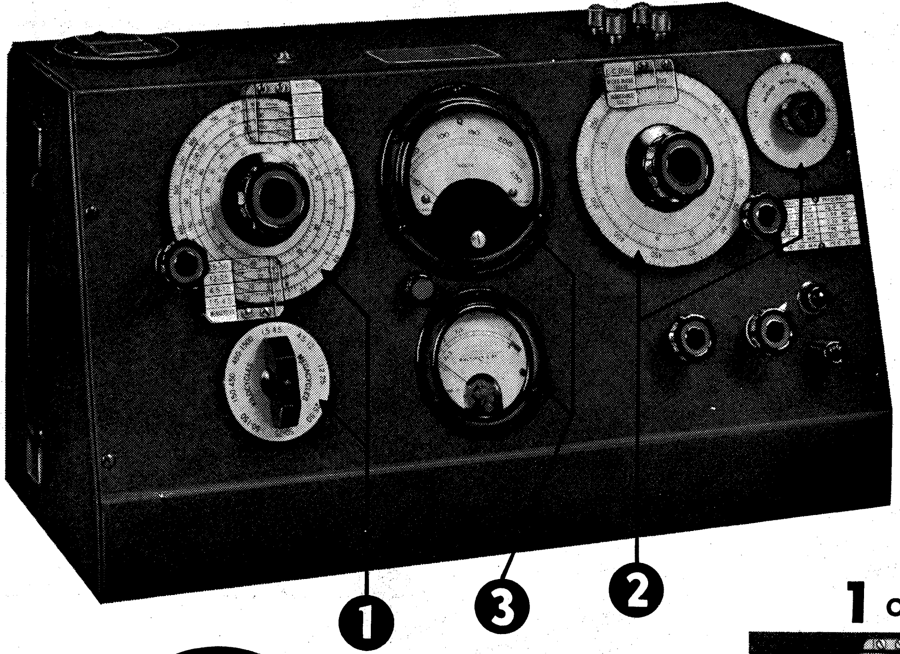
Taco Super 980 Yagi
Combines all the advantages of the 5-element design with the best features of the Twin-driven design. Tops in gain, directivity and mechanical dependability. (For high-band use Taco 5-element Yagi)

Tacoplex
Distribution systems for apartments, hotels, schools, and community installations.

TACO

TECHNICAL APPLIANCE CORPORATION
ONE TACO AVE., SHERBURNE, N.Y.
Tacoplex Master Antenna Systems, Taco Antennas, Taco Antenna Superchargers Taco Antenna Accessories.
In Canada: Stromberg-Carlson Co., Toronto 4, Ont.

*AVAILABLE AT YOUR TACO DISTRIBUTOR



EXAMINE THESE
Direct Reading Features
WHICH SIMPLIFY ACCURATE MEASUREMENTS

The Q METER
TYPE 160-A

50 kc. to 75 mc.

Radio frequency circuit design often requires the accurate measurement of Q, inductance and capacitance values. For this application the Type 160-A Q-Meter has become the uncompromising choice of radio and electronics engineers in this country and abroad.

Each component part and assembly used in the manufacture of this instrument is designed with the utmost care and exactness. Circuit tolerances are held to values attainable only in custom built instruments.

With the 160-A Q-Meter, as with other Boonton Radio Corporation instruments, the keynote in design is to embody accurate *direct reading* features which save time and simplify operation.

SPECIFICATIONS

Oscillator Frequency Range: 50 kc. to 75 mc. in 8 ranges.

Oscillator Frequency Accuracy: ±1%, 50 kc.—50 mc.
±3%, 50 mc.—75 mc.

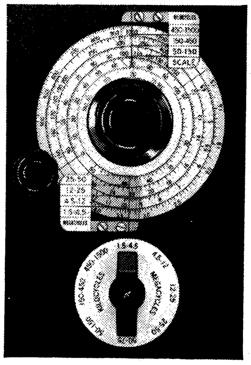
Q Measurement Range: Directly calibrated in Q, 20-250. "Multiply—Q—By" Meter calibrated at intervals from x1 to x2, and also at x2.5, extending Q range to 625.

Q Measurement Accuracy: Approximately 5% for direct reading measurement, for frequencies up to 30 mc. Accuracy less at higher frequencies.

Capacitance Calibration Range: Main capacitor section 30-450 mmf, accuracy 1% or 1 mmf whichever is greater. Vernier capacitor section +3 mmf, zero,—3 mmf, calibrated in 0.1 mmf steps. Accuracy ±0.1 mmf.

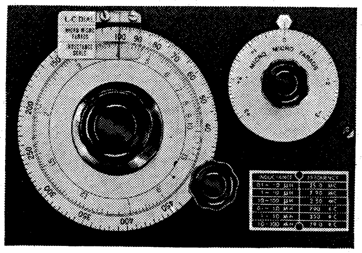
Catalog "H" containing further information available upon request.
(In Canada, direct inquiries to RCA Victor Co., Ltd., Montreal.)

1 OSCILLATOR FREQUENCY DIAL.



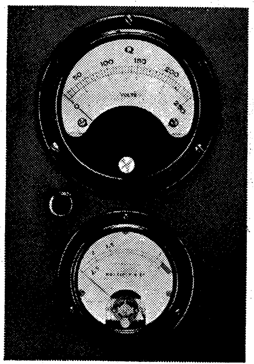
This large 4½" open faced dial has eight overlapping frequency ranges, each calibrated *directly* in kilocycles or megacycles, with scales conveniently divided for maximum readability. A vernier dial drive enables fine settings to be made with ease. All frequency ranges are accurate to within ±1% except the 50-75 megacycle range which is accurate to ±3%. The clearly marked range change switch located directly beneath the frequency dial facilitates rapid and positive selection of the desired frequency band.

2 Q-TUNING CAPACITANCE DIALS.



L-C dial serves twofold purpose of (1) conveniently and accurately indicating tuning capacitance *directly* in MMF, and (2) providing an effective inductance scale which also becomes *direct* reading at certain defined frequencies shown on frequency reference plate. Incremental capacitance dial at right calibrated from +3 MMF through zero to —3 MMF, accurate to ±0.1 MMF.

3 Q-VOLTMETER AND MULTIPLIER METER.



For the indication of Q values the 160-A Q-Meter employs a Weston Model 643 Meter calibrated *directly* in terms of Q over the range from 20-250. The damping of the meter movement is ideal for the rapid determination of exact resonance without sluggishness or overshoot. The lance type pointer enables Q readings to be obtained to the nearest unit. Located directly beneath the Q voltmeter is the "Multiply-Q-By" meter which provides Q multiplier factors of X1 to X1.5 in 0.1 steps, X2, and X2.5 thereby extending the useful range of Q indication to 625. This meter is carefully matched to a particular thermocouple element for maximum accuracy.

DESIGNERS AND MANUFACTURERS OF THE Q METER · QX CHECKER
FREQUENCY MODULATED SIGNAL GENERATOR · BEAT FREQUENCY
GENERATOR AND OTHER DIRECT READING INSTRUMENTS

BOONTON RADIO *Corporation*
BOONTON · N · J · U · S · A ·

Slotted Line Equipment

for measurements of **POWER ☆ VOLTAGE**

IMPEDANCE ☆ ATTENUATION ☆ STANDING-WAVE RATIO

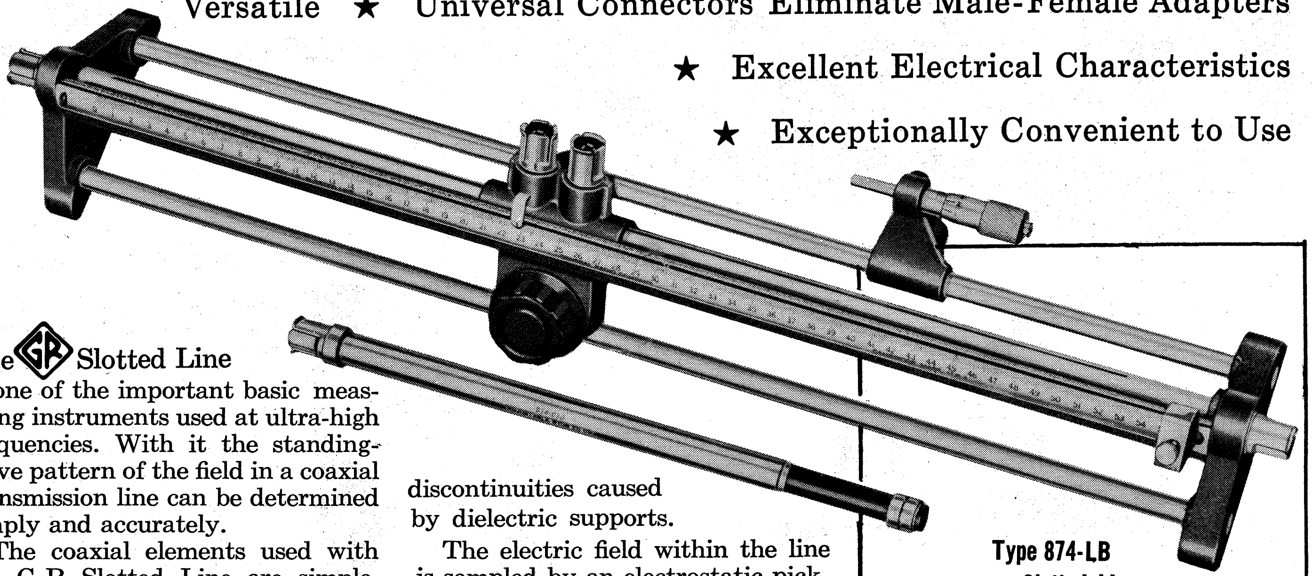
between 300 and 5,000 Mc

Ideal for the UHF T-V Band ★ Interchangeable Units ★ Inexpensive

Versatile ★ Universal Connectors Eliminate Male-Female Adapters

★ Excellent Electrical Characteristics

★ Exceptionally Convenient to Use



The GR Slotted Line is one of the important basic measuring instruments used at ultra-high frequencies. With it the standing-wave pattern of the field in a coaxial transmission line can be determined simply and accurately.

The coaxial elements used with the G-R Slotted Line are simple, inexpensive and readily interconnected for many different measurement setups. Wherever possible throughout the complete line, the characteristic impedance is 50 ohms.

A relatively small investment in these versatile components provide-equipment for most u-h-f measures ment problems.

The Type 874-LB Slotted Line is a 50-ohm, air dielectric, coaxial transmission line with a longitudinal slot in the outer conductor. The inner conductor is supported at its ends only, by two Type 874 Connectors, minimizing reflections and discontinuities caused by dielectric supports.

The electric field within the line is sampled by an electrostatic pick-up probe, mounted on a sliding carriage. Coupling between the line and the probe is adjusted by changing the probe penetration. The maximum longitudinal travel of the probe is 50 cm.

A built-in crystal rectifier is used as a detector of the r-f voltage induced in the probe. It is tuned to the operating frequency by a Type 874-D20 Adjustable Stub which plugs into a connector on the probe carriage. A receiver may be used, also, as a detector. Type 874 terminals are provided for connecting the receiver.

Type 874-LB Slotted Line

Frequency Range: *300 to 5,000 Mc*

Detector: *silicon crystal supplied . . . connectors for receiver built in . . . Stub (illustrated) available, as an accessory, for tuning the crystal*

Characteristic Impedance: *50 ohms ± 1%*

VSWR of Terminal Connectors: *less than 1.02 at 1,000 Mc; less than 1.07 at 4,000 Mc*

Constancy of Probe Coupling: *variations along line less than ± 2½%*

Completely Adjustable Depth of Probe Penetration

Adjustable Centimeter Scale: *simplifies calculations; calibrated in millimeters*

Intermittent Slow-Motion Drive: *disengaged by upward pressure of knob for free sliding; engaged by downward pressure for fine adjustment*

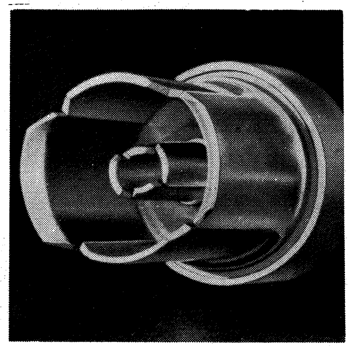
Accessory Micrometer Vernier: *for measurement of high standing-wave ratios by width-of-minimum method*

Convenient for Field Use: *sturdy, light weight (8 pounds) and portable*

Type 874-LB Slotted line	**$220.00**
Type 874-D20 Adjustable Stub	**10.50**
Type 874-LV Micrometer Vernier	**20.00**

GENERAL RADIO COMPANY Cambridge 39, Massachusetts

The versatility of the entire line of G-R u-h-f measuring equipment is based on the Type 874 Connector with which all of the coaxial elements are equipped. These connectors are truly 'universal' not only for v-h-f and u-h-f devices, but also for lower-frequency measurements where shielding and small ground inductance are essential.



1. Each will plug into any other connector
2. Electrical characteristics are excellent
3. Adaptable to panel mounting, rigid air lines, or flexible cables, without basic change
4. Accept not only other connectors of same type, but also Type 274 banana plugs
5. Designed for simple, quick plug-in connect and disconnect
6. Characteristic impedance is 50 ohms, generally used throughout the field

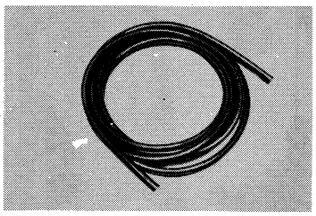
Basic Kit **Type 874 EK BASIC COAXIAL KIT** $342.25

To assist in the selection of elementary slotted line measuring equipment, with which a considerable number of impedance and standing-wave measurements can be made, the G-R Type 874-EK Basic Coaxial Kit is available.

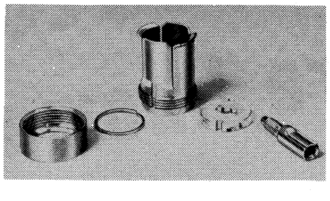
This kit includes all of the coaxial elements shown on this page, as well as the Type 874-LB Slotted Line and the Type 874-D20 Adjustable Stub illustrated on the opposite page. It does not include an oscillator for driving the line or a tuned amplifier for use with the built-in crystal detector. Suitable and inexpensive 'unit' oscillators and amplifiers are available. Write for full information.

A large number of associated elements and inexpensive auxiliary units are available: Unit Oscillators, Unit Power Supplies, Amplifiers and Detectors, Mixer Rectifiers, Voltmeter Rectifiers, Bolometer Bridge, Voltmeter Indicator, Attenuators, Line Elements, Filters, Adaptors, Patch Cords, etc.

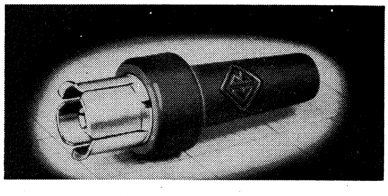
Send coupon for Complete Information on these and other G-R U-H-F Measuring Instruments and Components, fill-in this coupon for a 16-page booklet.



25 ft. 874-A2 Polyethylene Cable *for permanent or semi-permanent use; for long patch cords. Characteristic impedance 50 ohms ± 5%. Double-shielded with good mechanical flexibility. Attenuation at 100 Mc about 2.6 db/100 ft.; at 1,000 Mc about 10.5 db/100* ft. **$6.75**

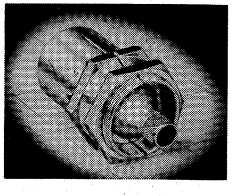


Two 874-B Basic Connectors *for rigid, 50-ohm, air-dielectric coaxial lines. Two @ $1.25 =* **$2.50**

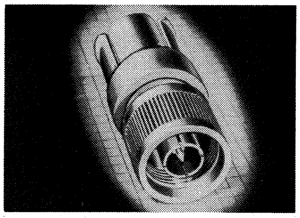


Two 874-C Cable Connectors *for G-R Type 874-A2 Polyethylene Cable. Transition pieces tapered to maintain 50-ohm characteristic impedance. Two @ $2.00 =* **$4.00**

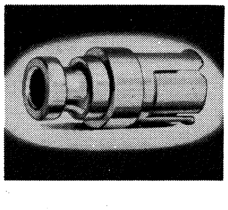
Two 874-C8 Cable Connectors (*not illustrated, similar to 874-C*) *for Army-Navy Type RG-8/U Cable. Two @ $2.00 =* **$4.00**



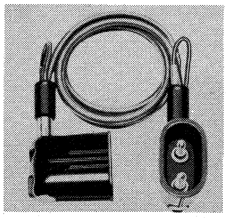
Two 874-P Panel Connectors *fit Type 874-A2 Cable. Similar to the cable connector, except panel adaptor and nut supplied in place of rubber guard. Two @ $2.50 =* **$5.00**



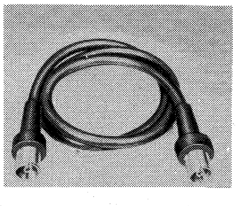
One 874-Q1 Adaptor to Type N *plugs into Army-Navy Type UG-22/U and similar jack-type connectors.* **$4.50**



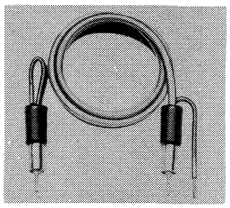
One 874-Q6 Adaptor *fits G-R Type 274-NF Shielded Cable.* **$2.00**



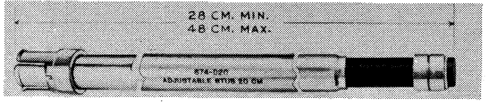
One 274-NE Shielded Connector *shielded lead assembly with shielded terminals.* **$5.50**



Two 874-R20 Patch Cords *Three ft. of Type 874-A2 Cable (double shielded) with Type 874-C Connectors. Two @ $6.00 =* **$12.00**

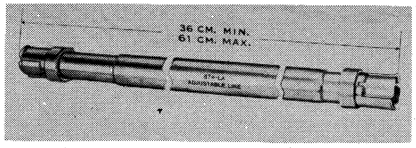


One 274-NF Patch Cord *Three ft. shielded cable with Type 274-NE Shielded Connectors.* **$2.50**

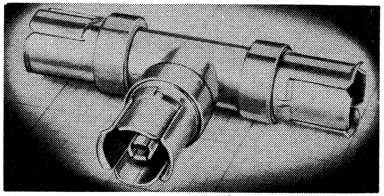


One 874-D20 Adjustable Stub *For matching or tuning, and use as a reactive element. Coaxial lines with sliding short-circuit moved by bakelite tube. Reference marker for use as wave meter with scale. Maximum travel of short-circuit is 20 cm.* **$10.50**

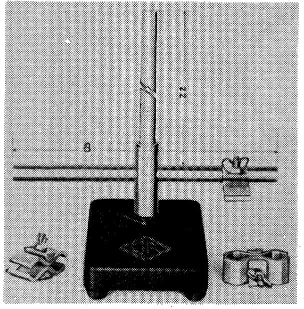
One 874-D50 Adjustable Stub *identical with Type 874-D20 except maximum travel is 50 cm.* **$12.00**



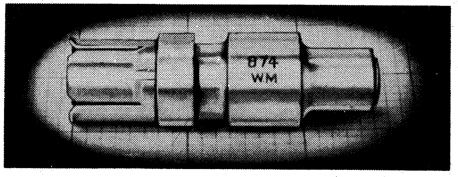
One 874-LA Adjustable Line (*Line Stretcher*) *An air-dielectric, coaxial line that can be telescoped to change its length. Used in matching networks. Length change is 25 cm.* **$15.00**



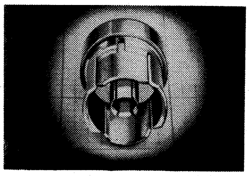
One 874-T Tee *For connecting stubs and other elements in shunt with a coaxial line.* **$7.50**



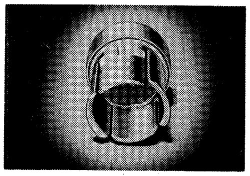
One 874-Z Stand *Supports parts of a wide variety of coaxial systems; heavy bronze base; 22-inch vertical and 8-inch horizontal brass rods; three universal clamps; will not rust or corrode.* **$12.50**



One 874-WM Matched (50 Ohm) Termination *Provides good impedance match for 50-ohm coaxial systems from dc to several thousand Mc. Consists of a 50-ohm cylindrical resistor mounted in a tapered coaxial holder.* **$10.50**



One 874-WN Short-Circuit Termination *A fixed short-circuiting strap mounted in a connector.* **$3.50**



One 874-WO Open-Circuit Termination *A shielding cap for open-circuited lines.* **$2.00**

For measurements of standing-wave ratios higher than 10, the following additional coaxial elements are required:

One 874-LV Micrometer Vernier Attachment (*illustrated on opposite page*). **$20.00**

One 874-F1000 Low-Pass Filter *for reduction of harmonics from an u-h-f generator. Cutoff Frequency is 1000 Mc.* **$22.50**

One 874-F500 Low-Pass Filter *identical with above except cutoff frequency is 500 Mc.* **$22.50**

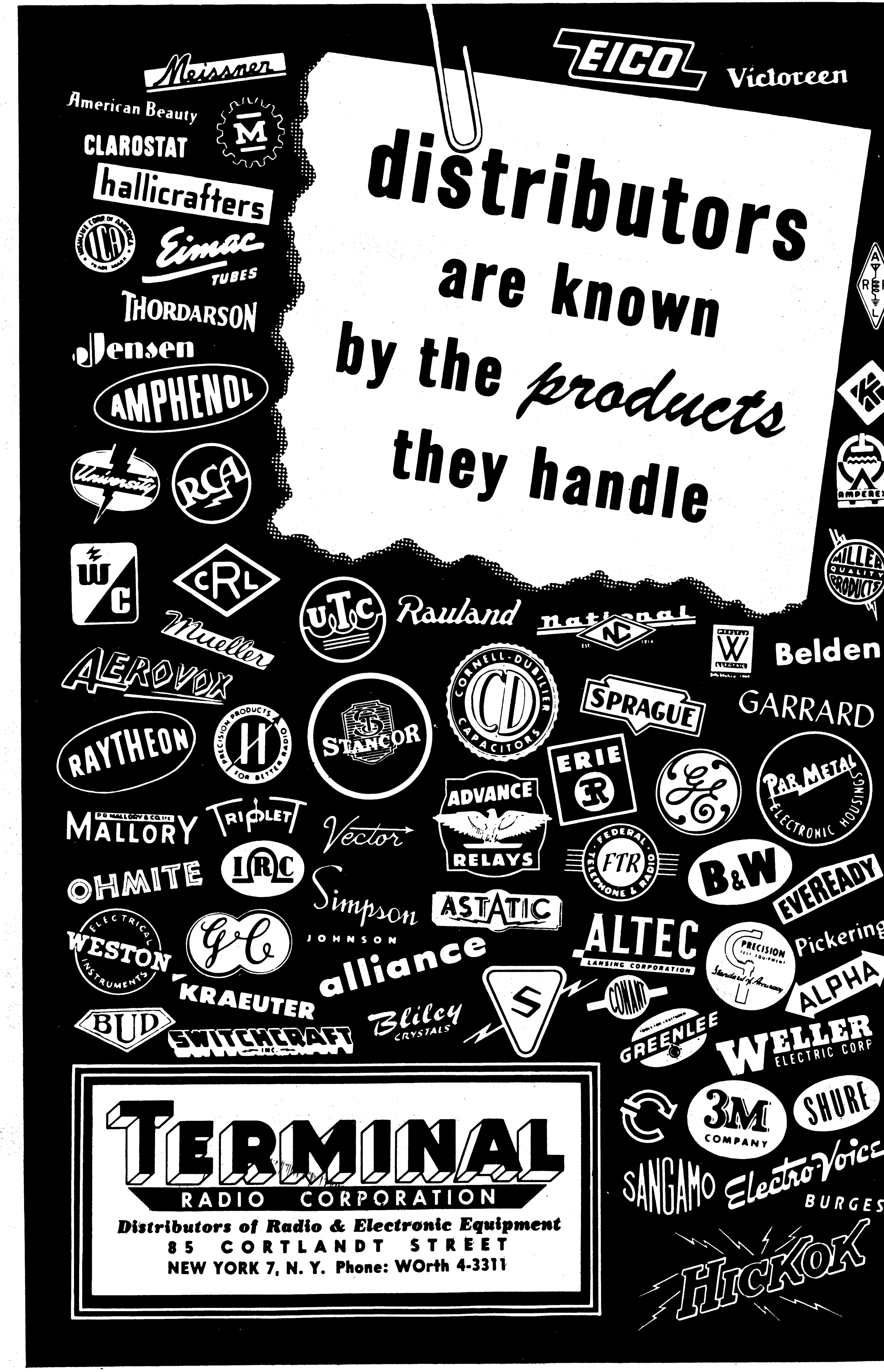
GENERAL RADIO Company 275 Massachusetts Ave.
Cambridge 39, Mass.

Please send me a copy of the 16-Page booklet describing the complete line of G-R Type 874 Coaxial Elements

Name ....

Address ....

....



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RADIO CORPORATION

Distributors of Radio & Electronic Equipment

85 CORTLANDT STREET

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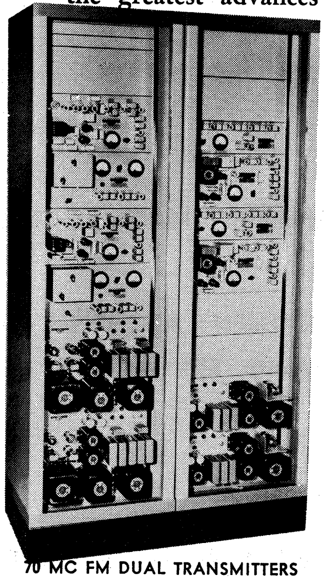
REL . . . *Engineering Leadership*

in FREQUENCY MODULATION MULTIPLEXING EQUIPMENT

All employing the ***SERRASOID**** Modulator, noted for its simple and reliable operation. Has no tuned circuits. Uses only standard receiver type tubes.

Because of high standards of research and design, coupled with rigid standards of quality, REL equipment has always represented the greatest advances in FM transmitters and receivers.

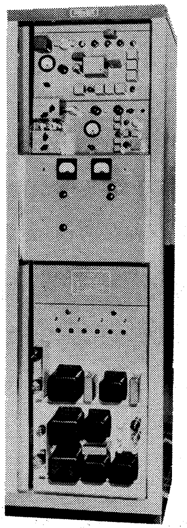


70 MC FM DUAL TRANSMITTERS & RECEIVERS Model 759

Preferred by

BROADCAST STATIONS • AIRLINES • TELEPHONE & TELEGRAPH COMPANIES • INDUSTRIAL SYSTEMS • UNIVERSITIES FOREIGN & DOMESTIC MUNICIPAL DEPARTMENTS.

Radio equipment as illustrated can be supplied to meet rigid operational requirements in carrier frequency range from 70 to 2,000 MC and in modulation band widths up to 200 KC in multiplex operations, suitable for voice, program, telegraph, teletype, remote control, telemetering and facsimile up to the equivalent of 50 voice circuits.



900 MC FM TRANSMITTER & RECEIVER for MULTIPLEX RELAY. Model 707-757

Some Other Products of REL

FM BROADCAST TRANSMITTERS • MOBILE REMOTE PICKUP • ***SERRASOID**** FM MODULATORS • SPECIAL PURPOSE FM RECEIVERS

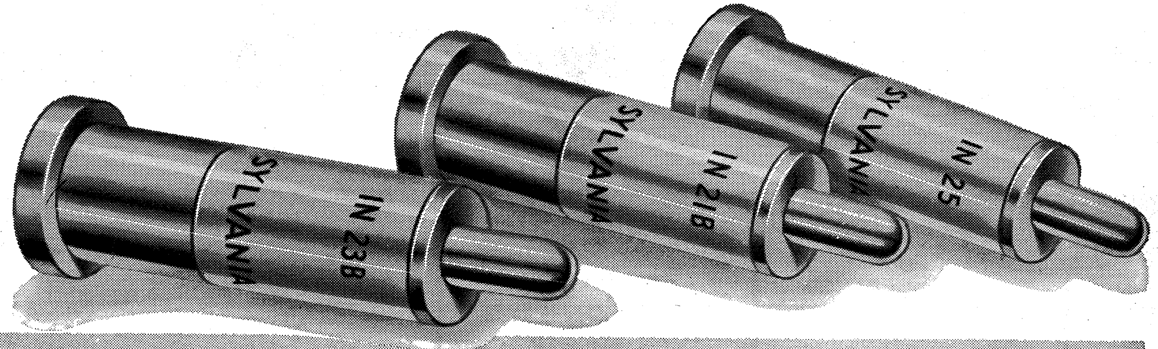
For Ease of Installation . . . Economy of Operation . . . Absolute Dependability . . . REL Units Stand Alone . . . Inquiries Are Invited for Specific Applications.

***T.M.Registered**

REL

RADIO ENGINEERING LABORATORIES, Inc.

36-40 37th Street, LONG ISLAND CITY 1, N.Y.



SYLVANIA PLUGS THE 16,000 MC GAP

with the new 1N78 Silicon Crystal Mixer

Sylvania adds another to the world's widest Silicon Crystal Mixer line—the 1N78 for 16,000 mc, one of the newest SHF bands. This new diode is the latest product of Sylvania's continuing exploration into frequency conversion in microwave regions.

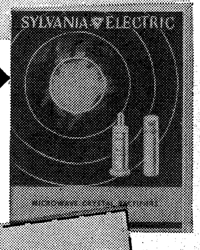
Better and better performance at existing frequency bands and new designs for tomorrow's frequencies are both to be expected of Sylvania's advanced research and long experience in Silicon Diode technology.

Sylvania also makes Silicon Crystal Video Detectors for use as microwave detectors in receivers of non-heterodyne type. Other Sylvania products engineered for radar and SHF receivers include magnetrons, TR tubes, ATR tubes and hydrogen thyratrons.

Sylvania Silicon Mixer Diodes

Type	Construction	Design Frequency (Approx.)
1N25	Cartridge	1000 mc.
1N21B	Cartridge	3000 mc.
1N23B	Cartridge	10,000 mc.
1N78	Coaxial	16,000 mc.
1N26	Coaxial	24,000 mc.
1N53	Coaxial Miniature	Above 30,000 mc.

Write for this 16-page book, "Microwave Crystal Rectifiers," including the new 1N78 characteristics and ratings.



Sylvania Electric Products Inc.
Dept. E-1009, Emporium, Pa.

Please send me the "Microwave Crystal Rectifiers" booklet, including data on the 1N78.

*Name*____________________

*Company*____________________

*Street*____________________

*City*__________ *Zone*____ *State*______

SYLVANIA

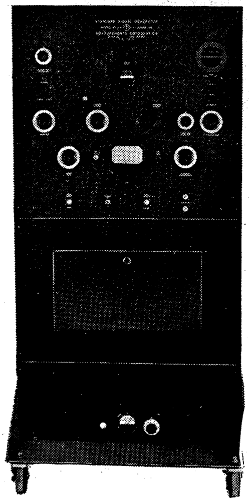
SYLVANIA ELECTRIC PRODUCTS INC. 1901-1951 50th ANNIVERSARY FIVE DECADES OF PROGRESS

ELECTRONIC DEVICES; RADIO TUBES; TELEVISION PICTURE TUBES; ELECTRONIC TEST EQUIPMENT; FLUORESCENT TUBES, FIXTURES, SIGN TUBING, WIRING DEVICES; LIGHT BULBS; PHOTOLAMPS; TELEVISION SETS

Laboratory Standards

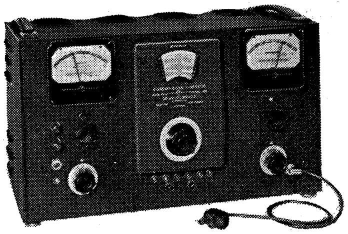
STANDARD SIGNAL GENERATORS

MODEL	FREQUENCY RANGE	OUTPUT RANGE	MODULATION	DIMENSIONS
65-B	75 Kc.-30 Mc.	0.1 microvolt to 2.2 volts	AM. 0 to 100% 400 cycles or 1000 cycles External mod., 50-10,000 cycles	11" high x 20" wide x 10¼" deep Weight: 55 lbs.
78	15-25 Mc.; 195-225 Mc. 15-25 Mc.; 90-125 Mc. other ranges on order	1 to 100,000 microvolts	AM. 8200-400 cycles 625-400 cycles Fixed at approximately 30%	10" high x 13" wide x 7" deep Weight: 22 lbs.
78-FM	86 Mc.-108 Mc.	1 to 100,000 microvolts	Deviation 0-300 Kc., 2 ranges FM. 400-8200 cycles External modulation to 15 Kc.	10" high x 13" wide x 7" deep Weight: 25 lbs.
80	2 Mc.-400 Mc.	0.1 to 100,000 microvolts	AM. 0 to 30% 400 cycles and 1000 cycles External mod., 50-10,000 cycles	10¾" high x 19" wide x 9½" deep Weight: 45 lbs.
82	20 cycles to 200 Kc. 80 Kc. to 50 Mc.	0-50 volts 0.1 microvolt to 1 volt	Continuously variable 0-50% from 20 cycles to 20 Kc.	15" high x 19" wide x 12" deep Weight: 50 lbs.
84	300 Mc.-1000 Mc.	0.1 to 100,000 microvolts	AM. 0 to 30%, 400, 1000 or 2500 cycles. Internal pulse modulator. External mod., 50-30,000 cycles.	12" high x 26" wide x 10" deep Weight: 135 lbs.
90	20 Mc.-250 Mc.	0.3 microvolt to 0.1 volt	Continuously variable, 0 to 100% Sinusoidal modulation 30 cycles to 5 Mc. Composite TV modulation	58¾" high x 28¼" wide x 25½" deep Weight: 302 lbs.



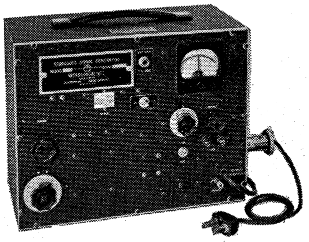
MODEL 90

TELEVISION SIGNAL GENERATOR



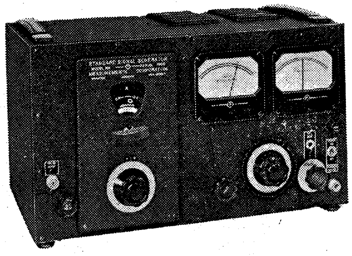
MODEL 65-B

STANDARD SIGNAL GENERATOR



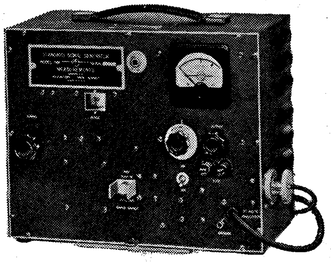
MODEL 78-FM

FM SIGNAL GENERATOR



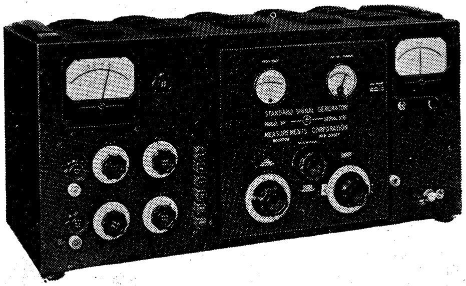
MODEL 80

STANDARD SIGNAL GENERATOR



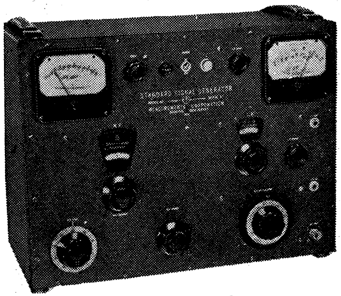
MODEL 78

STANDARD SIGNAL GENERATOR



MODEL 84

STANDARD SIGNAL GENERATOR



MODEL 82

STANDARD SIGNAL GENERATOR

MEASUREMENTS CORPORATION BOONTON · NEW JERSEY

STANDARDS ARE ONLY AS RELIABLE AS THE REPUTATION OF THEIR MAKER

SQUARE WAVE GENERATOR

MODEL	FREQUENCY RANGE	WAVE SHAPE	OUTPUT	DIMENSIONS
71	Continuously variable 6 to 100,000 cycles	Rise time less than 0.2 microseconds with negligible overshoot	Step attenuator: 75, 50, 25, 15, 10, 5 peak volts fixed and 0 to 2.5 volts continuously variable.	7″ high x 15″ wide x 7½″ deep Weight: 20 lbs.

U.H.F. RADIO NOISE and FIELD STRENGTH METER

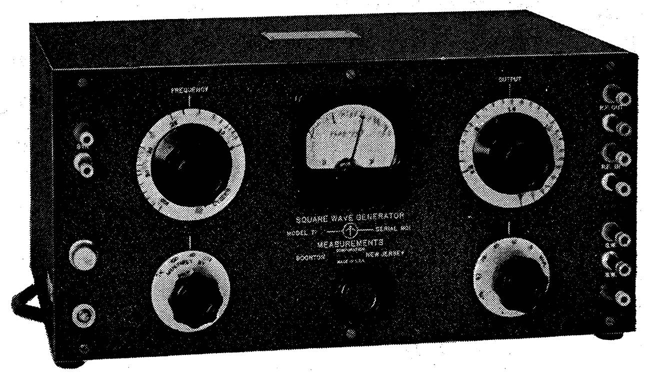
MODEL	FREQUENCY RANGE	SENSITIVITY RANGE	DIMENSIONS
58	15 Mc. to 150 Mc.	1 to 100,000 microvolts in antenna. 1 to 100 microvolts on semi-logarithmic output meter; balanced resistance attenuator with ratios of 10, 100 and 1000 ahead of all tubes.	16″ wide x 9″ high x 11″ deep Weight: 35 lbs.

PULSE GENERATOR

MODEL	REPETITION RATE	PULSE WIDTH	OUTPUT	DIMENSIONS
79-B	60 to 100,000 pulses per second	Continuously variable from 0.5 to 40 microseconds PULSE SHAPE Rise time: approx. 0.5 microseconds. Decay time: no more than 25% of pulse width at a 25% duty cycle.	Approximately 150 volts positive with respect to ground. "Sync. Output" 75 volts positive with respect to ground	10″ high x 13½″ wide x 10½″ deep Weight: 31 lbs.

MODEL 79-B

PULSE GENERATOR



MODEL 71

SQUARE WAVE GENERATOR

MODEL 58

U.H.F. RADIO NOISE and
FIELD STRENGTH METER

MEASUREMENTS CORPORATION BOONTON · NEW JERSEY

Laboratory Standards

MEGACYCLE METER

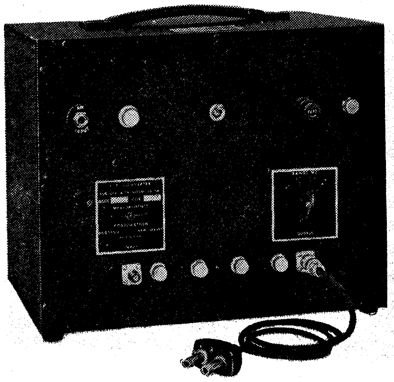
MODEL	FREQUENCY RANGE	FREQUENCY ACCURACY	MODULATION	DIMENSIONS
59	2.2 Mc. to 400 Mc.	Within ±2%	CW or 120 cycles fixed at approximately 30%. Provision for external modulation	Oscillator Unit 3¾" diameter; 2" deep. Wt. 1 lb.

I. F. CONVERTER

MODEL	FREQUENCY	OUTPUT	MODULATION	DIMENSIONS
M-275	4.5 Mc., 10.7 Mc., 21.7 Mc.	10 microvolts to 1.0 volt when used with Model 78-FM	Up to approximately 80% AM combined with or exclusive of FM	10" high x 13" wide x 7" deep Weight: 15 lbs.

VACUUM TUBE VOLTMETERS

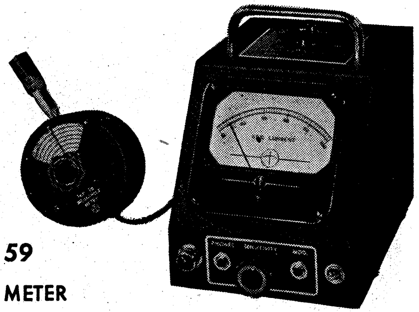
MODEL	VOLTAGE RANGE	FREQUENCY RANGE	INPUT IMPEDANCE	DIMENSIONS
62	0-1, 0-3, 0-10, 0-30 and 0-100 volts AC or DC	30 cycles to over 150 Mc.	Approximately 7 mmfd.	6" high x 4¾" wide x 8½" deep. Wt. 8 lbs.
62-U.H.F.	0-1, 0-3, 0-10, 0-30 and 0-100 volts AC or DC	100 Kc. to 500 Mc.	Approximately 2 mmfd.	6" high x 4¾" wide x 8½" deep. Wt. 15 lbs.
67	.0005 to 300 volts peak-to-peak	5 to 100,000 sine-wave cycles per second	1 megohm shunted by 30 mmfd.	7½" high x 7" wide x 8½ deep. Wt. 10 lbs.



MODEL M-275

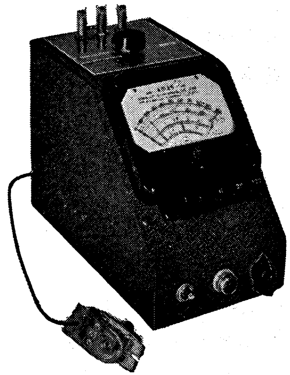
I. F. CONVERTER

For use with Model 78-FM FM Signal Generator



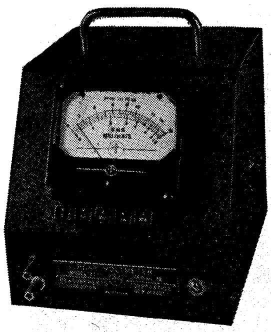
MODEL 59

MEGACYCLE METER



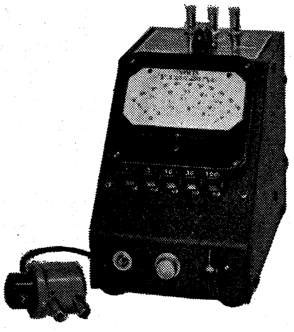
MODEL 62 U.H.F.

VACUUM TUBE VOLTMETER



MODEL 67

PEAK VOLTMETER



MODEL 62

VACUUM TUBE VOLTMETER

MEASUREMENTS CORPORATION BOONTON · NEW JERSEY

STANDARDS ARE ONLY AS RELIABLE AS THE REPUTATION OF THEIR MAKER

U.H.F. OSCILLATOR

MODEL	FREQUENCY RANGE	OUTPUT RANGE	OUTPUT IMPEDANCE	DIMENSIONS
112	300 Mc.—1000 Mc.	Maximum varies between 0.3 volt and 2 volts Adjustable over 40 db	50 ohms	12½" high x 13½" wide x 8" deep Weight: 22 lbs.

INTERMODULATION METERS

MODEL	INTERMODULATION RANGE	FREQUENCIES	ANALYZER INPUT VOLTAGES	DIMENSIONS
30	0.1% to 30%	LF: 40, 70, 100 cycles HF: 2000, 7000, 12,000 cycles.	.0001 to 100 volts (—70 to + 40 DBM)	9¾" high x 19" wide x 13" deep Weight: 30 lbs.
31	0.5% to 30%	LF: 60 cycles HF: 3000 cycles	Full scale ranges of 3, 10, 30 volts RMS	8" high x 19" wide x 9" deep Weight: 16 lbs.

CRYSTAL CALIBRATOR

MODEL	FREQUENCY RANGE	FREQUENCY ACCURACY	HARMONIC RANGE	DIMENSIONS
111	.25 Mc. to 1000 Mc.	0.001%	.25 Mc. Oscillator: .25-450 Mc. 1 Mc. Oscillator: 1-600 Mc. 10 Mc. Oscillator: 10-1000 Mc.	8" high x 6" wide x 5" deep Weight: 4 lbs.

MODEL 112
U.H.F. OSCILLATOR

MODEL 31
INTERMODULATION METER

COMPLETE descriptive literature on any of our instruments will be sent to you on request.

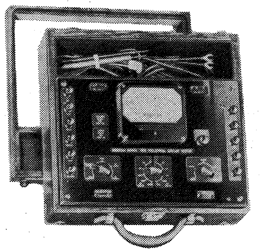
Consult us regarding your requirements for special electronic measuring equipment.

MODEL 111
CRYSTAL CALIBRATOR

MEASUREMENTS CORPORATION BOONTON · NEW JERSEY

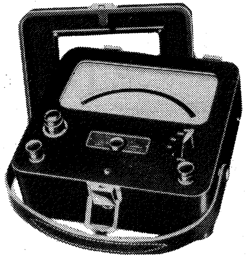
Dependable Electrical Measurements

FOR LABORATORY—PRODUCTION TESTING—MAINTENANCE



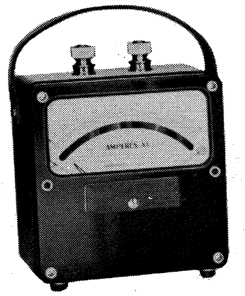
INDUSTRIAL CIRCUIT TESTER

(Model 785) A multi-range, multi-purpose, ultra-sensitive analyzer, for laboratory and industrial checking of electrical and electronic circuits. Has 28 practical scale ranges; measures d-c and a-c voltage, d-c and a-c current, and resistance. Accessories available to extend ranges. Compact and portable; furnished in either oak or steel case.



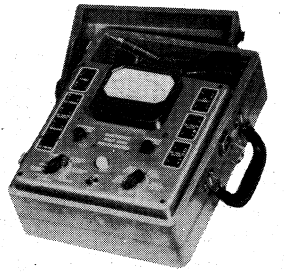
ULTRA SENSITIVE PORTABLE TEST INSTRUMENTS

(Model 622) High sensitivity, large scale A-C and D-C instruments of double pivoted type, ideal for precision measurement of low potentials and minute current. Available in single or multi range as D-C voltmeters, millivoltmeters, milliammeters, microammeters; and milliammeters and voltmeters in thermo and rectifier types for RF and A-C.



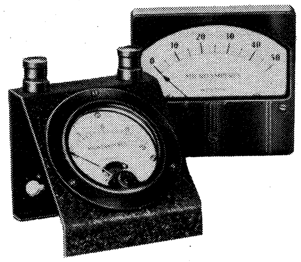
A-C AND D-C PORTABLE TEST INSTRUMENTS

(Model 931) Comprising a group of matched a-c and d-c portable test instruments; the universal favorites for highly accurate general test work. Available as voltmeters, ammeters, d-c and single phase a-c wattmeters—a-c and d-c voltmeters. Long, readable hand-calibrated mirror scales. The d-c instruments have new self-shielding mechanism.



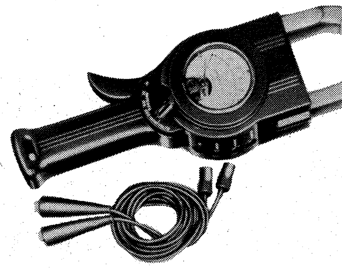
HIGH FREQUENCY ELECTRONIC ANALYZER

(Model 769) A three-in-one instrument providing a self-contained Volt-Ohm-Milliammeter, a high impedance electronic D-C Volt-Ohmmeter, and a probe type vacuum tube voltmeter for use to 300 megacycles. Exceptionally stable and accurate. Has specially designed extremely small RF and D-C probes.



PANEL INSTRUMENTS OF HIGH SENSITIVITY

In all types, sizes and ranges for all critical measurements encountered in electronics, nuclear research and general testing. All D-C, rectifier type A-C and RF instruments have built-in magnetic shunts for minor adjustments if necessary.



A-C CLAMP VOLT-AMMETER

(Model 633, Type VA-1) For convenient and rapid measurement of a-c voltage and current without breaking the circuit. Jaws take insulated or non-insulated conductors up to 2" diameter. Safe, rugged, versatile. Also available as a-c clamp ammeter, without voltage ranges.

Literature on these and other WESTON instruments for electrical and electronic requirements sent on request.
WESTON Electrical Instrument Corporation, 587 Frelinghuysen Avenue, Newark 5, New Jersey
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