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TELEVISION TRANSMITTING ANTENNA FOR EMPIRE STATE BUILDING

BY

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Summary—*The theory for a new approach to the design of wide-band radiators for vision antennas is set forth and demonstrated in its application both to models and to the television antenna recently installed on top of the Empire State Building in New York City. A band width was obtained which is several times that required to meet television standards in the United States.*

The design principles of a sound antenna capable of being located close to the vision antenna without causing mutual interference, are described and demonstrated.

The paper also deals with important transmission line phenomena and circuit problems, particularly as they occur in connection with this antenna.

A short description of mechanical and auxiliary details is included.

INTRODUCTION

HIGH-DEFINITION television has forced the development of circuit features which only recently seemed unattainable. Providing transmitting antennas, of the broadcasting type, having constant characteristics over a frequency band sufficiently wide to accommodate the full band spectrum of high-definition television, becomes increasingly difficult with decreasing carrier frequency.

The recently installed antenna on top of the Empire State Building in New York City represents a successful attempt to remove these difficulties from the frequency regions with which television is concerned. This antenna, Figure 1, consists of two separate, independent radiator systems for vision and sound transmission supported by a common column. The radiators at the middle of the column constitute the vision antenna and the top radiators constitute the sound antenna.

In the vision antenna is represented the application of a new approach to wide-band antenna development by which the radiating elements themselves, uncombined and unaided by any form of compensating circuit, present input impedances which are resistive and constant over a frequency band considerably wider than that required by present television standards.

Rational carrier-frequency allocation[1] and certain receiver requirements call for closely adjacent carrier frequencies for vision and sound transmission. In spite of this close frequency spacing, it was found possible to design the vision and sound radiator systems so that negligible coupling effect was obtained even when the two systems are located no further apart from each other than the distance of a half wave, Figure 1. The arrangement represents a clean and efficient method of preventing mutual interference between vision and sound transmitters.

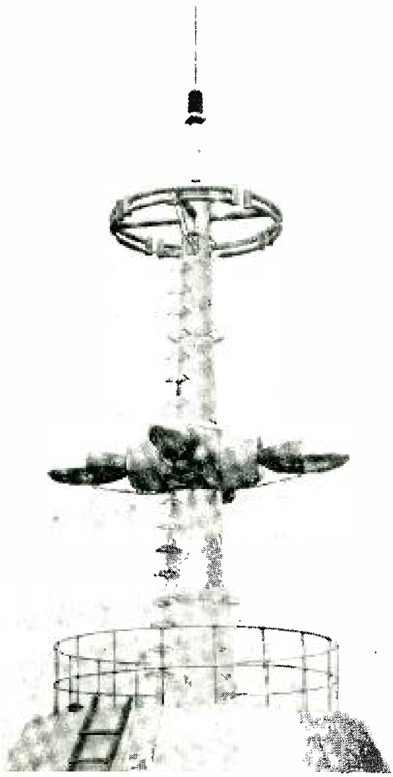

Fig. 1—Television antenna for Empire State Building, as erected for test at Rocky Point Laboratory.

While the radiation from each of the radiating systems is horizontally polarized, the vertical directivity of each approximates that of a vertical dipole. Propagation tests have fairly well established the superiority of horizontal polarization[2] which for this reason has been recommended by RMA as standard.

General Principles—Vision Antenna

It is a well-known principle that the total impedance of circuits having inductance and capacitance, becomes independent of frequency if these components are associated with resistive loads of values equal to the square root of the inductance-capacity quotient, Figure 2. An equivalent version of the circuit in Figure 2b is shown in Figure 3a. By composing a radiator from coaxial transmission-line sections as shown in Figure 3b, it is possible to make it consist of an inductive component *A* and a capacitive component *B* of such relative lengths that they carry equal portions of the radiation load. By properly proportioning the transverse dimensions of radiator components *A* and *B*, the square root of their inductance-capacitance quotient may be made equal to the radiation resistance of each component. The characteristics of the circuit in Figure 3a will thus be simulated. However,

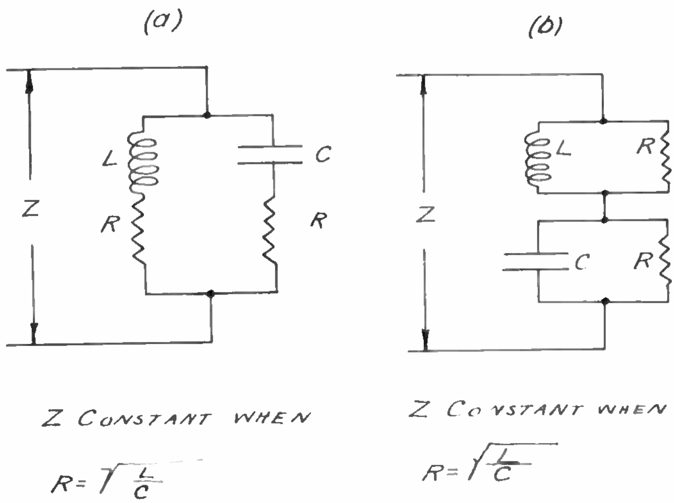

Fig. 2—Circuits for constant impedance at variable frequency.

due to the fact that the electric and magnetic high-frequency fields of the radiator in Figure 3b are distributed and thus dependent upon frequency, it is evident that inductance, capacitance and radiation resistance all become variables with frequency. The precise requirement for the ideal constant-impedance circuit—as in Figure 3a which calls for fixed values of inductance, capacitance, and resistance—is, therefore, not fulfilled. An input impedance independent of frequency could, therefore, not be expected. It was, however, considered reasonable to expect appreciable benefit when the proportions of a composite radiator are such that the requirements for the constant-impedance circuit are fulfilled at mid-spectrum frequency.

Proportioning of the inductive and capacitive radiator components required much experimenting. Steps in this experimental evolution of proportions toward optimum dimensions may be followed by the samples shown in Figure 4. The effects of shape as well as of dimensions were studied. A generally elliptic shape of all the radiator surfaces, Figures 4 and 5, seemed to permit the fullest utilization of the constant-impedance principle. Gradual expansion of the transmission line into the radiator also appeared as the only satisfactory method of connection. It is known that expansion of a radiator away from its feed point is an aid to impedance constancy at variable frequency. This coincidence is no doubt an aid to the characteristics of the composite radiator.

As may be concluded from the band-width curves shown in Figure 4, the band width increases rapidly with transverse radiator dimensions

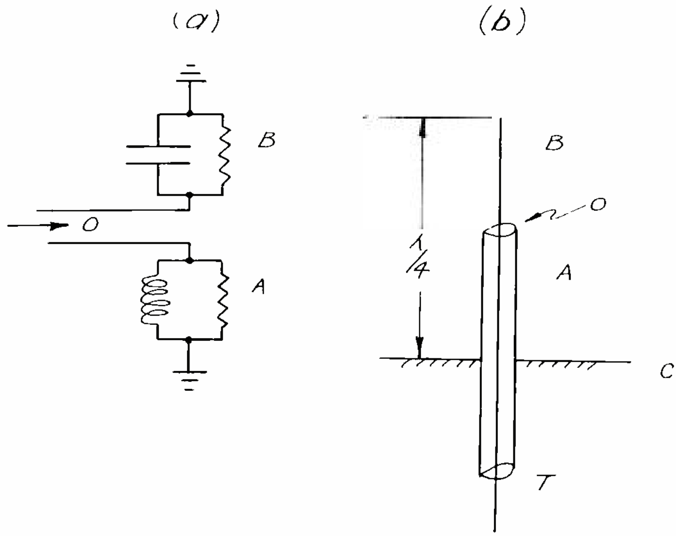

Fig. 3—Application of constant-impedance circuit to composite radiator.

as proper proportions are approached. When these dimensions are exceeded the band width again diminishes. This fact seems to indicate that the theory set forth in connection with Figure 3 represents the basic principle upon which the performance of the composite radiator depends.

The elliptically shaped extension of the center line conductor, which represents the capacitive component *B* of the composite radiator Figure 5, will hereafter be referred to as the "ellipsoid". The expanding portion of the outer conductor of the transmission line, which has elliptic curvature, will be referred to as the "throat". The surface which forms the continuation of the throat into the radiator, constitutes the inductive component *A* of the composite radiator. The curving of this surface is also elliptic and this radiator component will be referred to as the "collar".

In order that the protruding portion of the ellipsoid and the collar might be equally loaded by radiation, their relative lengths, Figure 5a, should be as 7 to 5. The input impedance, governed by this ratio, is in the order of 110 ohms. The best ratio between major and minor axis of the ellipsoid was found to be 15 to 6. The best ratio between

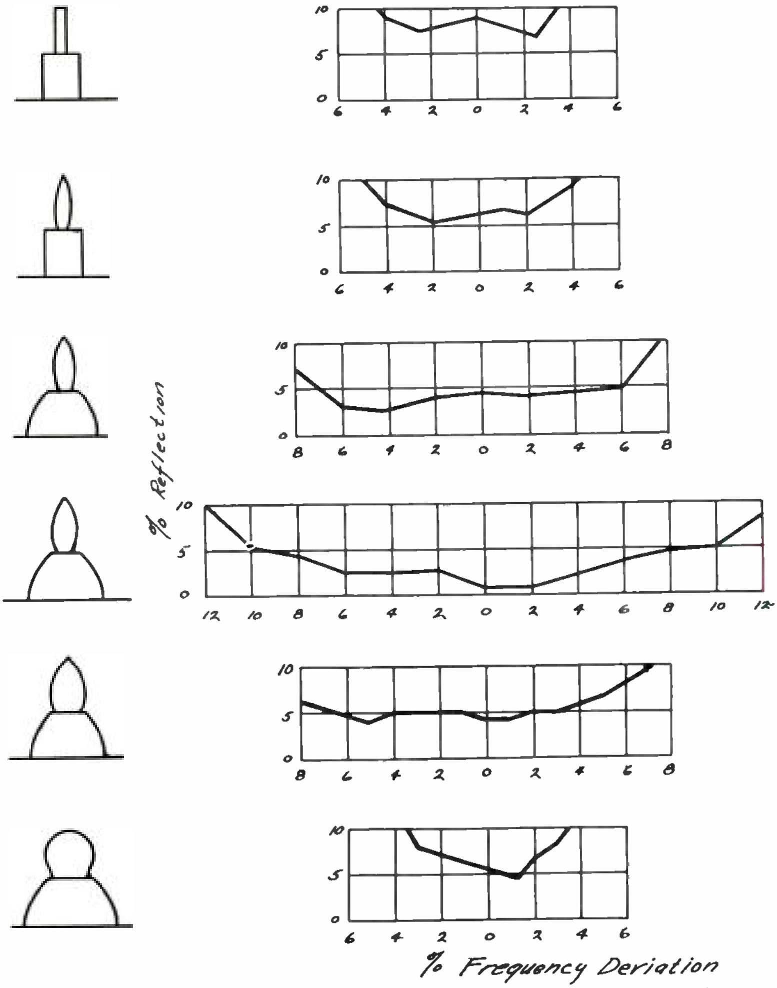

Fig. 4—Experimental steps in determining optimum proportions for maximum band width of composite radiator.

the mean outside diameter of the collar and the minor axis of the ellipsoid appeared to be 3 to 2.

The band width, within which reflection of waves from the radiator back over the transmission line was less than 5 per cent, with a composite radiator Figure 5a, vertically mounted from a horizontal metallic surface, was 20 per cent, Figure 4. This result was obtained with small scale models designed for a mean frequency of 150 Mc.

As can be understood from Figures 3 and 5a, the transmission line is in effect connected in series with the two components of the composite radiator. Since the currents on the center conductor and the throat surface are equal and opposite there is substantially no radiation from this portion of the system. However, after the current on the throat surface has passed over the joining edge to the collar surface its direction will be the same as that of the center conductor current. The currents in the collar and the ellipsoid having the same direction, the two radiator components will act like one continuous radiator. At

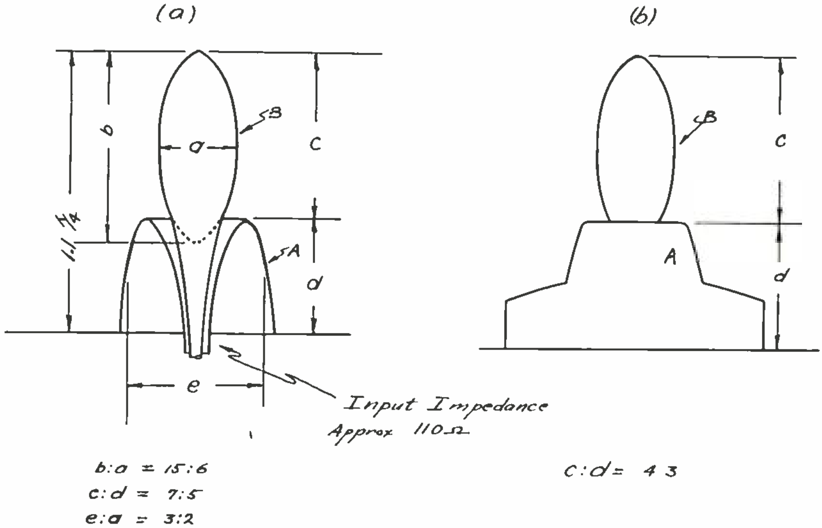

Fig. 5—General dimensions of composite radiator as modified to permit use of supporting bracket.
(a) Single radiator over a conductive surface.
(b) Model of radiator for turnstile antenna combination.

mid-frequency the total length of the composite radiator exceeds that of a quarter wave by about 10 per cent, Figure 5a.

In the throat expansion of the early models, constant ratio was maintained between the outer and the center conductors. In other words the characteristic impedance of the throat expansion was held constant. For practical purposes, however, such as mechanical strength and to obtain immunity against lightning, some form of metallic bracket, Figure 7, between ellipsoid and collar seemed indispensible. The capacitance required to balance the inductive shunt introduced by the bracket was most effectively obtained by gradual decrease of the characteristic impedance of the throat section in the direction of expansion.

In the development of the bracket only experimental methods were practical. There were many variables such as self-inductance, mutual inductance, size, length, and position which all had to be coordinated to secure the maximum mechanical strength obtainable without sacrifice of band width. By not compromising on the band width, the dimensions of the bracket were limited, but fortunately these dimensions provided adequate mechanical safety factor. The bracket as finally

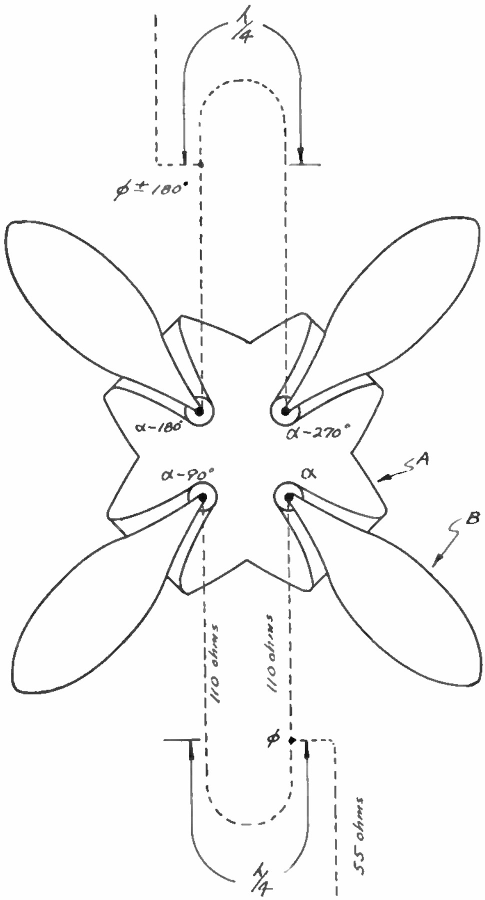

Fig. 6—General arrangement of composite radiator turnstile antenna.

developed for the full scale antenna is shown in Figures 1, 12, 15, and 17.

For broadcasting horizontally uniform, vertically polarized radiation, a single vertical composite unit as shown in Figure 5a is sufficient. However, since horizontal polarization was required, several horizontally located composite units had to be combined to produce horizontally uniform radiation, or the equivalent. A simple method of combining, which was adopted, is to arrange four composite radiators into a "turnstile" antenna. A turnstile antenna, in its simplest form, consists of two dipoles crossed at right angles carrying high-frequency currents in phase quadrature.[3] Although the radiation in the plane of each dipole is not uniform, the combination of the two, by virtue of their phase and geometric quadrature, results in a rotating electromagnetic field which produces the equivalent of uniform radiation in the plane of the two dipoles. Two composite radiators of the type shown in Figure 5a pointing in opposite directions, joined at the base of their collars and fed in phase opposition, form a composite dipole. Two such dipoles may then be combined in the horizontal plane into a composite turnstile antenna. The four collars in such a turnstile form

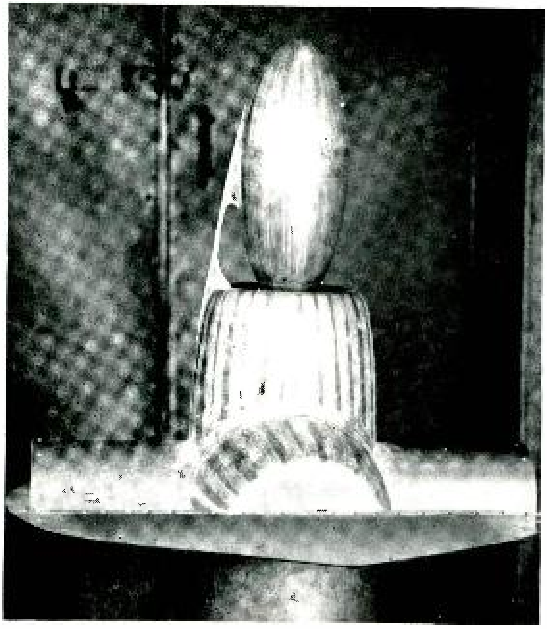

Fig. 7—Model of radiator for 100-150 Mc turnstile antenna combination.

a central cross-shaped hub from which the four ellipsoids protrude radially, Figures 1, 6, 8, 9, and 12.

Before building a complete 100-150 Mc turnstile model, Figures 8 and 9, a half section of such a model as shown in Figure 7 was made in order to expedite determination of the new relative collar length resulting from the alterations in the collar characteristic when four collars are combined into a hub. The new length ratio between the protruding portion of the ellipsoid and the collar became 4 to 3 as shown in Figure 5b.

A complete 100-150 Mc turnstile model was then built and tested, Figures 8 and 9. From the foregoing discussion it is clear that the four composite radiators of this turnstile were so connected that the currents they carried were equal and in progressive phase quadrature, Figure 6. That means that the currents in any adjacent pair of radiators were in phase quadrature while the currents in any opposite pair were in phase opposition.

The input resistance of each radiator in the turnstile was 110 ohms. The feeders to the radiators were, therefore, designed to have a characteristic impedance of 110 ohms. In this way the characteristic

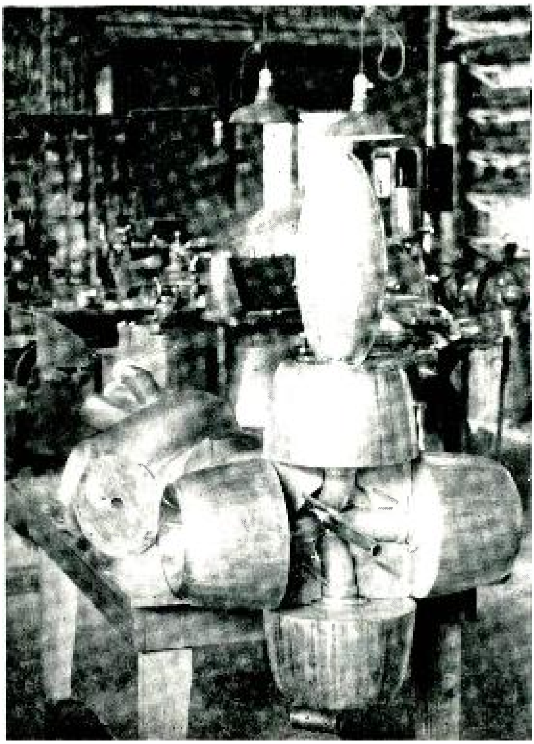

Fig. 8—Shop view of 100-150 Mc turnstile antenna.

impedance of each feeder was "matched" by the radiator, and high-frequency waves arriving over the feeder were radiated into space substantially without reflection back toward the transmitter.

Since the waves on a reflection-free line travel only in one direction, the phase of the high-frequency current it carries becomes a linear function of the distance along the line from any point of reference. The phase between the currents at points a quarter wave apart thus becomes 90 degrees. Progressive phase quadrature between the currents of the four radiators might then have been obtained by providing four separate matched feeders progressively differing in length by a quarter wave.

Since, in a progressive phase-quadrature system, the currents in opposite radiators are in phase opposition, Figure 6, it was more practical and economical to provide two equal 55-ohm main feeders coupled in phase opposition to the transmitter. Each was split into two 110-ohm branches near the antenna to feed a pair of adjacent radiators. These 110-ohm branches were made to differ in length by a quarter wave. In this way, the feeder branches of two opposite radiators could be

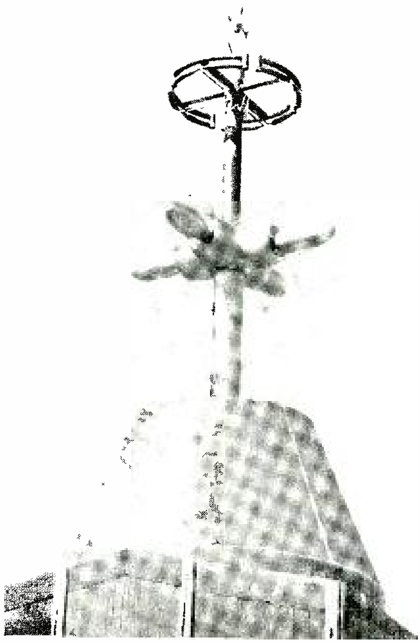

Fig. 9—Complete 100-150 Mc model of combined vision and sound antennas.

made equal in length if they were connected to main feeders of opposite phase. Progressive phase quadrature was thus obtained by using only two different lengths of branch feeders.

When two equal high-frequency loads, such as an adjacent pair of radiators in a turnstile, are supplied with power from a common branch junction through sections of transmission lines, a quarter wave different in length, the input impedance to such a system, as a whole, is more constant over a wide frequency band than the input impedance into each branch line.

If equal reflections are produced by the two radiators, the input impedances of the branch lines, at their junction, become reciprocals with respect to the characteristic impedance of the feeders. These reciprocal impedances are brought into parallel combination at the branch junction. If the differences between the reciprocal values is moderate, the resultant parallel impedance is very little different from the characteristic impedance of the main feeder, which has been designed to match the combined characteristic impedance of the branch feeders. That is, if the reflections on the feeder branches are moderate, these reflections will nearly compensate each other at the junction. When the branches differ in length by an odd multiple of a quarter

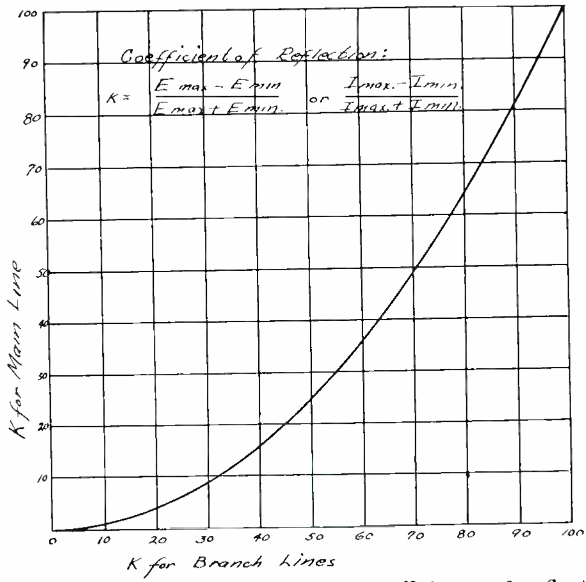

Fig. 10—Curve indicating relation between coefficients of reflection of main feeder and parallel quarter wave, differential length, branch feeders.

wave and the reflections are equal, the coefficient of reflection on the main line becomes equal to the square of the coefficient of reflection on each branch. This relationship between coefficient of reflection on branches and main feeder is shown by the curve in Figure 10. It is evident from the curve that the ratio of compensation is large for low reflections and decreases toward unity as the reflection on the branches approaches 100 per cent. Thus, while the compensation is tenfold at 10 per cent reflection on the branches, there is no compensation when the reflection is 100 per cent.

It should be clear that the compensation becomes less and less effective as departure is made from midfrequency since the difference in length between the branches is then no longer a quarter wave. This is an important consideration in the case of circuits for wide frequency bands as required for television. The composite radiator possesses the important virtue that its input impedance departs from non-reflecting values very slowly with departure from midfrequency. This type of radiator is, therefore, capable of taking fullest advantage of the compensation intrinsic in phase-quadrature combinations.

It may be realized that antenna loads which closely match the main lines throughout the operating frequency band, become an increasing necessity as the distance between transmitter and antenna becomes large in terms of wave length. A standing wave, resulting from a certain mismatching, will shift in its location in proportion to frequency and also in proportion to distance, in terms of wave length, between point of reflection and point of reference. The biggest shift will, therefore, occur at the input end of the line. If the shift is large enough the input impedance of the line will pass through both a maximum and its reciprocal with respect to the characteristic impedance of the line. This means that for a certain frequency shift, the longer the line, the faster will be the recurrence of maxima and minima. For a

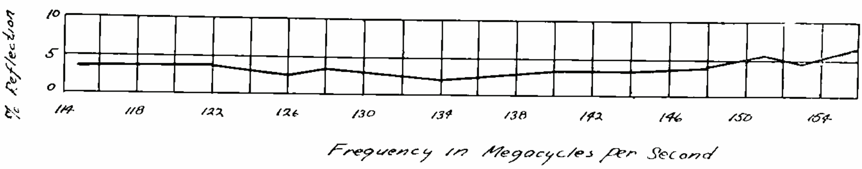

Fig. 11—Frequency-response curve of 100-150 Mc model vision antenna.

long line it is, therefore, entirely possible to have several impedance maxima and minima occur within the operating frequency band. Since the impedance passes through both maxima and minima it is evident that the input-impedance variation assumes considerable proportions even if the mismatch at the load end is small. If the line is sufficiently short, such large variations as from a maximum to a minimum will not occur within the operating band. The impedance variation will, therefore, be smaller.

It is of theoretical interest to note that impedance variations at the transmitter end of a line will not necessarily cause variations in the power output of a transmitter. If the internal impedance of the transmitter matches the characteristic impedance of the line, the power output at a fixed reflection will not vary with frequency. In practice, however, variable impedance is a serious factor.

The 100-150 Mc model antenna, Figures 8 and 9, was found to possess a band width well in excess of 30 per cent, Figure 11, which was over 50 per cent more than the band width of a single composite radiator unit. The "band width" is defined as the frequency spread within which reflection of waves back on the main feeders is less than 5 per cent. Coefficient of reflection is the quotient of the difference and the sum of maximum and minimum voltage or current on a line, Figure 10. Percentage reflection is obtained by multiplying this quotient by 100.

The important dimensions, which had been determined with the model, were then reproduced to scale in a full-size turnstile antenna having a midfrequency of about 45 Mc which was constructed for installation on top of the Empire State Building, Figure 12. When the antenna was tested it was found to have the truly remarkable band width of over 60 per cent, Figure 13. This is six times the frequency spread of a single television channel for vision and sound. The band is from 6 to 10 times that obtainable with heretofore conventional designs combined with complicated correction networks.

The discrepancy in band width between model and full-size antenna

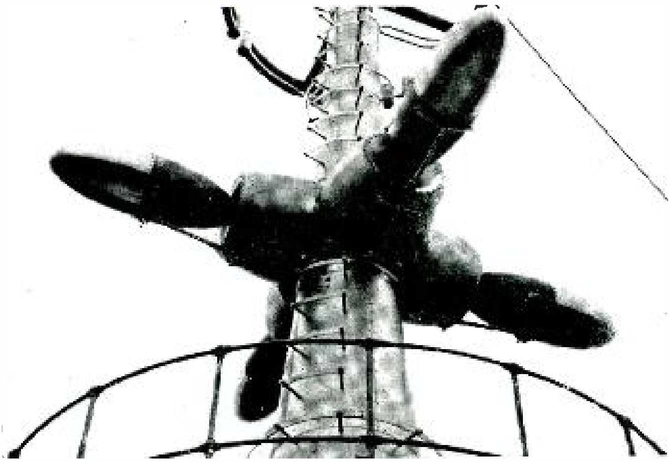

Fig. 12—Empire State Building vision turnstile radiator system.

must be attributed to greater accuracy in construction and perhaps also to better testing equipment. Up to the present, no opportunity has been available to study the band width of such large, single composite units as used in the antenna for the Empire State Building installation. It is possible that the 20 per cent band width obtained with 150-Mc models of single composite radiator units may be exceeded under the more accurate conditions possible with larger models.

When the full-scale antenna was set up and tested at Rocky Point, Figure 1, it was found possible to provide a metal rail around the testing dome on which it was mounted, by allowing the reflection to rise to a 6 per cent peak at 54 Mc. This slight marring of the band characteristic was, of course, of no consequence since this peak lies outside the operating band. After using an even more substantial rail and adding a non-symmetrically located ladder step near the radiators when installing the antenna at the Empire State Building, it was found that this peak shifted to 53 Mc and rose to a value of about 8 per cent.

Although a single turnstile antenna has less vertical directivity than a vertical dipole the difference was made up for by locating it a half wave above the roof of the dome.

It may be of interest to note that in the transmission lines used in connection with the full-scale antenna, the suspension insulators of the center conductor were located in pairs with a quarter-wave spacing between the insulators of a pair. In this way, use was made of the phase-quadrature principle for the purpose of reducing insulator effects.

Likewise, it may be interesting to note that the 55-ohm main feeders, which are several hundred feet long, were so made that in them

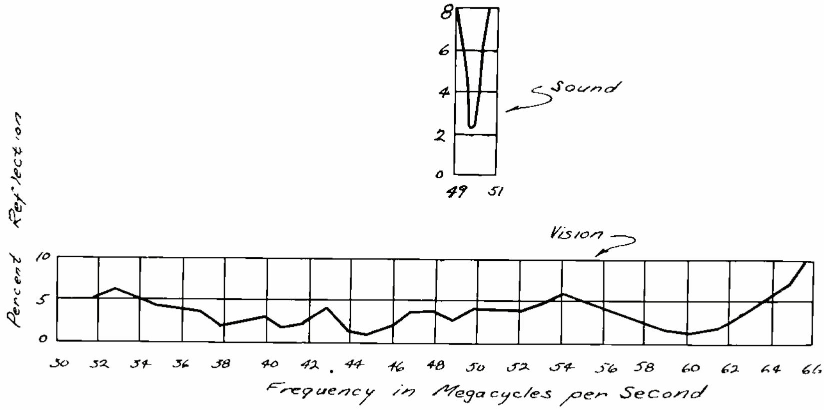

Fig. 13—Frequency-response curve of full-scale vision and sound antennas.

is represented an economic compromise between lines having a characteristic impedance of 77 ohms for securing maximum power-transfer efficiency and lines having a characteristic impedance of 30 ohms for securing maximum power-carrying ability for any given diameter of outer conductor.[4] The diameters of the outer and inner conductors of the main feeders were made 2.5 inches and 1 inch respectively.

Sound Antenna

In choosing the general design principles for an antenna for sound transmission to be mounted on the same structure as the vision antenna, minimum coupling between antennas is found to be the chief consideration.

Low coupling may be obtained in two different ways. Since the vision antenna produces a rotating field, the sound antenna could be built on the same principle, but he connected for phase rotation in the opposite direction. However, it may be readily understood that such a system would provide minimum coupling for only one frequency, at which the phase quadrature in both the sound and the vision antenna would be perfect. Since, in addition, it is desirable that each antenna have perfect phase quadrature at midfrequency, the condition cannot

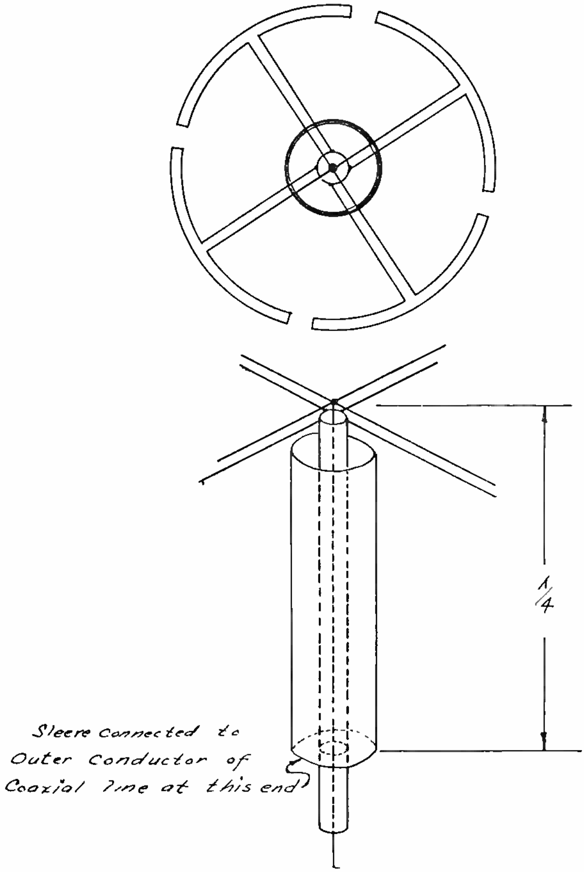

Fig. 14—Schematic diagram of sound antenna connections including line balance converter.

readily be fulfilled. Fortunately other means were available by which it was possible to furnish a design which provided low coupling for all operating frequencies.

A circular loop antenna of uniform characteristics and distribution, if located in a plane parallel to the horizontal plane of a turnstile antenna and on a common, central, vertical axis, would have no coupling to the vision antenna at any frequency. The diametrically located radiators of the vision antenna could not induce voltages giving rise to circulating currents in the loop and vice versa.

A simple way of providing the approximate equivalent of a loop antenna is to locate a number of dipoles in a circle. Cancelling effect between the pick-up of the segments or dipoles is in this case obtained at the point of connection to the common transmission line feeding the segments in parallel. This may best be understood by referring to Figure 14. The common feed line is so connected to the feeders of the individual radiators that the currents in opposite radiators become opposite. Currents, which on the other hand are induced by the turnstile antenna, are of same direction in opposite radiators. They must

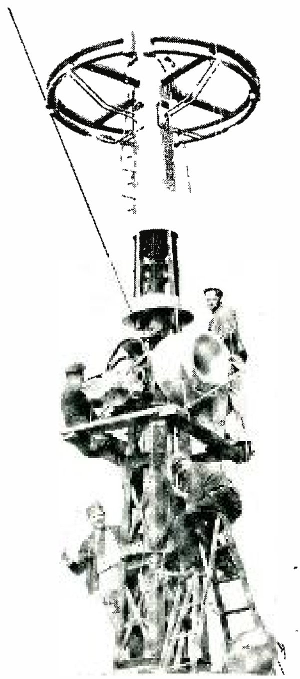

Fig. 15—Television antenna for Empire State Building during erection for test at Rocky Point Laboratory.

therefore balance each other without entering the common feed line. In the present case, the antenna was made up of four folded dipoles bent into circular segments, Figures 1, 9, 15, and 19. Among the dimensions of folded radiators which, at a given frequency, result in resistive input impedance, only the two smallest dimensions are of any interest in this case. At the larger of these dimensions the distance between the folding points of the radiator is approximately a half wave. The folding points coincide with maximum potential and the currents in the parallel conductors flow in the same direction. The distance between the folding points at the smaller dimension is only about a quarter of a wave. The input terminals are at maximum potential and the currents in the parallel conductors flow in opposite directions. A ring antenna of this latter type need only be about half the size of that required by the first type. The smaller dimensions further reduce the possibility of undesirable mutual effects between the sound and the vision antennas and reduce the mechanical problems. For these reasons, the small-type folded dipole was chosen.

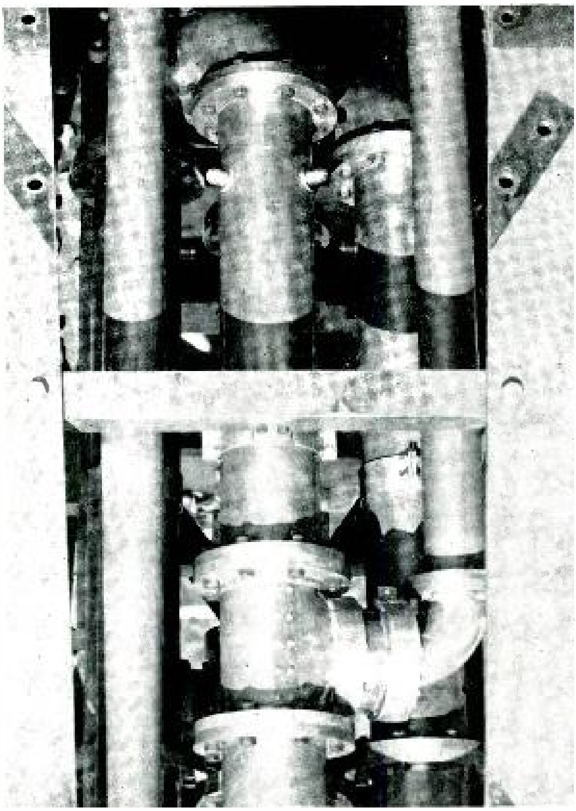

Fig. 16—Main line, junction and branch feeders inside the central supporting structure of the full-scale antenna. Note the quarter-wave branch feeder loops.

In the small-type folded dipole, the current folds with the conductor and tapers toward the input terminals where it becomes a minimum. On account of the voltage distribution thus obtained in this radiator, the capacitance between the parallel branches provides a path for circulating currents which lower the power factor. The resulting increase in frequency selectivity in the sound antenna further decreases the possibility of energy transfer from the vision antenna into the sound transmitter.

The input impedance of a single folded dipole, operating as described, would normally provide very high input impedance. However, when combining four such radiators into a ring-shaped antenna, by properly spacing adjacent folding points, it was found possible to influence the characteristics of the radiator so that the input impedance of each radiator in the combination was reduced to 220 ohms. This impedance was desirable since by parallel connection an impedance of 55 ohms could be obtained, without impedance transformation, which would equal the characteristic impedance of a coaxial line similar to the main feeders of the vision antenna. The four dipoles were then connected by balanced open feeder pairs, of 220-ohm characteristic

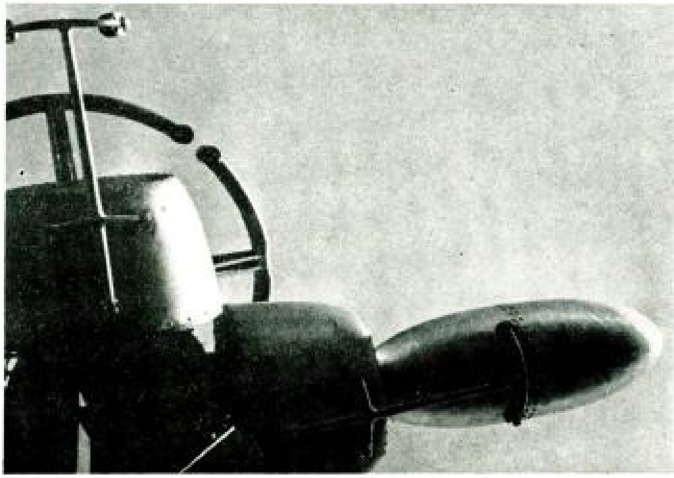

Fig. 17—Position of supporting bracket with respect to collar and ellipsoid.

impedance, through a balance converter to a 55-ohm coaxial line, Figure 14.

The line balance converter, Figure 14, involves the principle that the outer conductor of a coaxial line must be made electrically free from its surroundings in order not to destroy the balance of a system to which it is connected. This is done by surrounding the end section of the outer conductor with a concentric sleeve and by connecting the two together at a point a quarter wave from the end of the line, Figure 14. The impedance between the end of the outer conductor, of the coaxial line, and the sleeve which is bonded to the surrounding supporting structure has by this procedure been made very high. The current flowing on the inside of the outer conductor of the coaxial line can, therefore, become the continuity of the current in the connected load circuit. There is no longer any shunting effect to divert some of this current from its desired path.

This line balance converter, which was developed some time previous to the design of this antenna has found many varied uses and has made possible a number of coaxial-line combinations not otherwise possible.

Models of the sound antenna designed for 150 Mc were made and studied. The final model was located a half wave above the 100-150 Mc vision antenna model and the two antennas were supported by a common column, Figure 9.

The full-scale sound antenna, similarly mounted with respect to the full-scale vision antenna is shown in Figure 1. Its frequency response is shown in Figure 13. The power transfer from one antenna

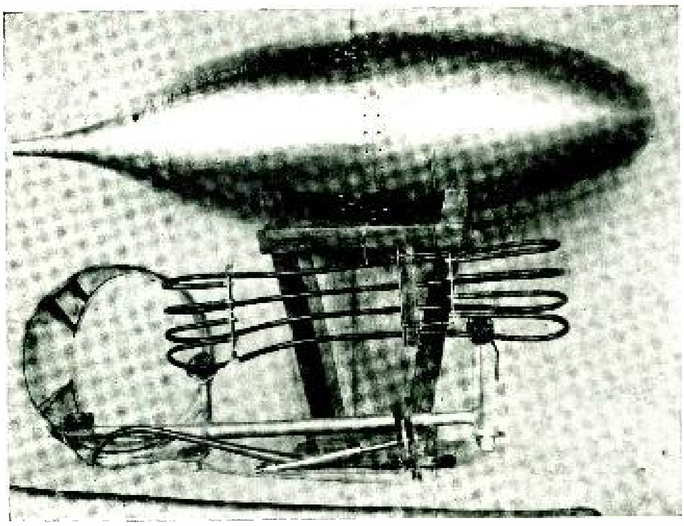

Fig. 18—Ellipsoid and associated parts.

to the other was found to be less than one part in a million of the input power.

Mechanical Design, Ice Removal, Etc.

The mechanical design of the antenna provides for a safety factor of 5 at a wind load of 40 lbs. per square foot, which corresponds to a true wind velocity of about 130 miles per hour. The weight of the total structure amounts to approximately 6,000 lbs.

The central supporting structure consists of a square lattice tower, Figure 15, which contains the necessary feeders, quarter-wave loops, Figures 6 and 16, and the balance converter, Figure 14. This tower, which is tapered, is covered with a circularly bent sheet metal skin, Figures 1 and 15. The four collars of the vision antenna at the middle of the tower have internal steel construction which is bonded to the tower itself, Figures 15 and 20. Figures 1 and 15 show how the collar and throat surfaces, which are made of spun sheet metal, are fitted over this frame. Attached to the inside framework of the collars are the supporting brackets for the ellipsoids. These brackets, which are shown in Figures 1, 15, and 17, are made of thick-walled Shelby steel tubing. The components of the brackets are welded together. The brackets are slightly tapered and have a bore through which runs the connection to the heating units located within the ellipsoids for ice removal. The ellipsoids are made in two halves joined together at the

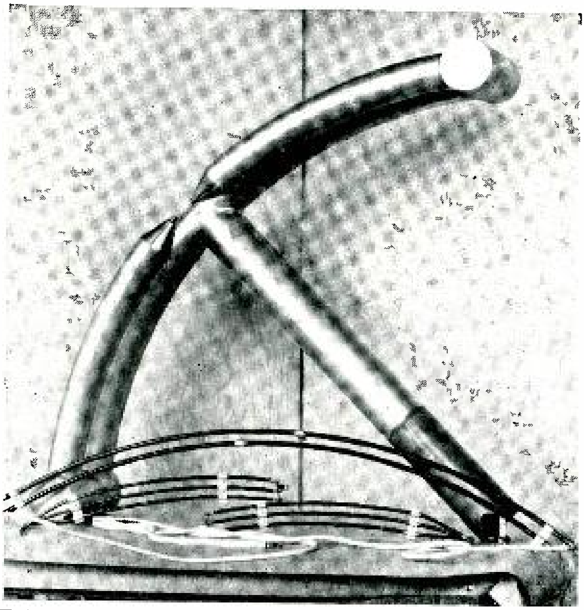

Fig. 19—Folded dipole unit of sound antenna with its associated parts.

middle by an aluminum ring, Figure 18, which is also bolted to the two ends of the bracket bridle, Figure 17.

The anchor-shaped or circular segments of the sound antenna, Figures 1, 15, and 19, are also made of steel tubing with their various components welded into one unit. The center T-member of each radiator slides into the supporting cross which is integral with the top of the tower, Figures 1 and 15. The sound antenna radiators are made up of different sizes of tubing. The different sizes are 3, 4, and 5 inches, Figure 19.

Calrod heating units, bent into convenient shapes, are installed in the radiator sections of the sound antenna, Figure 19, as well as in the ellipsoids, Figure 18, and the collar members of the vision antenna, Figure 20. The heaters of each radiator section are connected by individual leads to a distributing panel in the building dome under the antenna. Each section of the sound antenna, each ellipsoid, and each corresponding collar require normally about 1.5 kw of heating power for ice removal. Since there are four of each of these elements, each

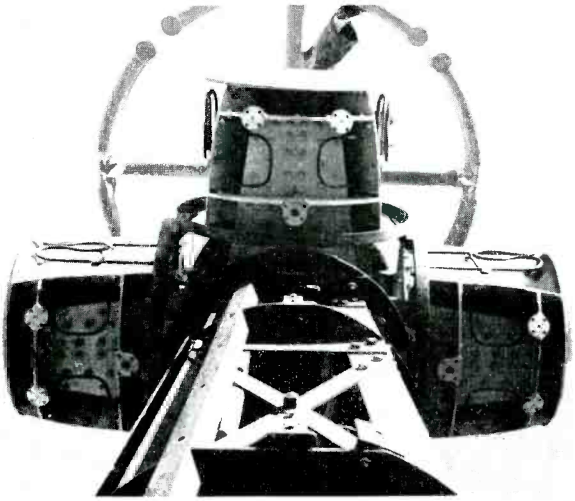

Fig. 20—Internal structure of collars. Note the Calrod heating units.

group of four will normally require 6 kw. There are three such groups which are then connected in three phase. Voltage tapping is arranged

Fig. 21—Ice formation on antenna at Empire State Building. Note the effect from the application of heat to the radiators.

to suit conditions so that the total three-phase power may be varied up to a maximum value of 27 kw. The photograph, Figure 21, shows ice conditions recently encountered. Application of heat keeps the radiators clear of ice and the electrical characteristics of the antenna remain unchanged.

The insulation throughout the transmission-line system is quartz. Three radially placed quartz rods form a suspension point for the center conductor. An exception to the use of quartz is the pressure-sealing insulators of the lines and the lead-in insulators for the sound antenna, Figures 1 and 15, which are made of high-grade porcelain. The throats of the vision antenna are closed by quartz windows which in turn are shielded from water creepage by mica guard rings located further out on the throat surface and on the ellipsoid connector.

Fig. 22—Lightning striking Empire State Building.

Except for the sound antenna and the brackets of the vision antenna, all exposed surfaces of the antenna and supporting structure are chromium-plated copper. There are no electrical reasons prohibiting the use of non-corrosive metals such as stainless steel when and if the cost of using this material can be tolerated.

Wind-velocity and wind-direction instruments are mounted at the top of the structure, also a lightning pick-up rod, Figure 1, to facilitate the lightning research carried on at this location[5], Figure 22.

ACKNOWLEDGMENT

The work described in this paper would have been impossible without the wholehearted cooperative spirit, power of imagination and skill of the entire personnel of the laboratory and shop.

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GAMMA AND RANGE IN TELEVISION*

BY

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Summary—*The similarity between television and other methods of pictorial reproduction is outlined and general principles of monochromatic pictorial reproduction are presented. A basis for the subjective aspect of the problem is given and works of Nutting and Abribat are reviewed. Concepts of gamma and range as well as perspective and density are defined and rules for handling of these variables for attaining desired pictorial effects are given.*

The peculiarities of television systems especially in respect to gamma and range are then presented. Methods of measurements and data on gamma and range characteristics of a typical television system are given. The need for standardization of the transmitted signal in respect to gamma characteristic is outlined, and established European standards are described together with the reasons for their establishment.

A REPRODUCTION of a scene, be it either a photograph or a painting, or a television image, can never be like the view of the original scene. One cannot reproduce reality, but only an appearance of the reality. The appearance of reality may be had either in shades of different colors or in shades of one color. The latter method is the most frequently used in still photography, in motion pictures, and in television. Even artists such as etchers frequently use the monochromatic method.

In general, a monochromatic reproduction of a scene containing a combination of regions of shade of several colors and of respective brightnesses

$$B_1, B_2, B_3, \cdots\cdots B_n$$

is accomplished by producing on a plane a combination of areas similar in geometry to the original scene and of respective brightnesses

$$B_1', B_2', B_3' \cdots\cdots B_n'$$

of one color. The degree of faithfulness of the reproduction is measured by the extent to which the reproduction gives an illusion of the original.

* This article was presented as a technical paper before Rochester Fall Meeting of IRE on November 15, 1938.

Two photographs of a given scene, placed side by side and identically illuminated may convey two very different impressions. One may give an illusion of the scene in daylight while the other may convey an impression of the same scene in moonlight. A closer examination will show that the brightness of the darkest dark in one print is equal to that in the other. Similarly the brightest bright in one print is equal in brightness to that of the second print. Both the room illumination and the illumination of the photographs may be varied within wide limits without a marked change of impressions given by the respective photographs.

What was just said simply means that with the same photographic paper having a limited range of brightness of about 50 to 1 one may

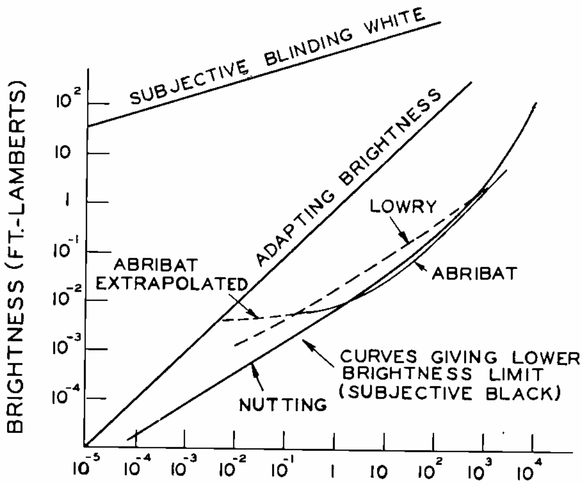

Fig. 1

successfully reproduce, or convey subjectively, the impressions of scenes the brightness of which may vary as much as 50,000 to 1. Moreover, the prints may be viewed with the incident illumination as well as background illumination varying within very wide limits with very little change in the impressions given by the photographs. Laws of physiological optics give the answer to the question: why is all this possible? One of these laws may be stated as follows: a human eye, if given sufficient time, proportions (adapts) its sensibility to the value of excitation.[1] The other law may be stated by saying that, to each value of adaptation excitation, or adaptation brightness, corresponds a

[1] A great deal of the information contained in this paper pertaining to the general subject of pictorial reproduction has been taken from an excellent and up-to-day discussion of the matter by Marcel Abribat (*Science et Industries Photographiques*, Vol. 6, p. 177, June 1935).

value of the lower limit and the higher limit of brightnesses visible. The concept of the lower and the upper limits of visible brightness is simple enough, but the adaptation brightness concept is rather complex.

When viewing a combination of brightnesses it is certain that the adaptation brightness is not a function of the brightness of any small region, but a function of some sort of an average of the brightnesses of the whole field of vision. When a small object on a uniformly lighted background is viewed, the background brightness may be safely taken as the adapting brightness. In Figure 1 are shown several curves of lower and higher visible brightness limits as functions of background brightness. These curves were compiled by Hopkinson[2] from works of Nutting, Lowry, and Abribat.

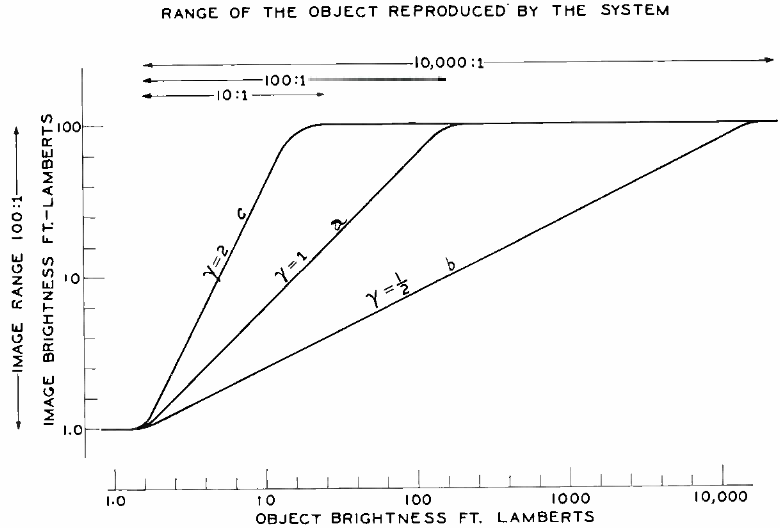

Fig. 2

On the basis of the above facts, a statement may be made that for a successful pictorial reproduction, be it photography, television or moving pictures, one need not reproduce the whole range of brightnesses of the original scene, but only the range perceptible to the eye. In general, none of the methods of pictorial reproduction fulfill even these reduced requirements, but when handled with skill derived from either art or science, most of them can cover a sufficient range for the illusion of the original scene to be conveyed faithfully and without great loss of the visible detail.

Most of the authorities on the subject of monochromatic pictorial reproduction agree that to give a faithful or a desired appearance of the original the following four conditions must be fulfilled.

[2] R. G. Hopkinson, *The Photographic Journal* (British), Vol. 77, p. 563, Sept. 1937.

1. Proper Perspective
2. Proper Gamma
3. Proper Range
4. Proper Density

The first condition is quite obvious. To give a true perspective the image should be viewed from such a distance that the angle subtended by the image shall be the same as the angle subtended at the eye by the original scene. Pictorial reproduction permits many artistic effects to be achieved by modifying this angle at the image. In such a case, it is not "faithful" reproduction, but a "desired" illusion. The second condition concerns gamma or contrast. Here the word "contrast" is used in the same sense as photographers use it in describing photographic printing papers as of high contrast, of medium contrast, and of low contrast or flat. In general, the overall gamma or contrast of a system (photography, television, etc.) is defined as the slope of a straight-line portion of the characteristic curve of the logarithm of image brightness versus the logarithm of object brightness.

In Figure 2 three typical characteristic curves of pictorial reproducing systems are shown. The curve a is for a contrast or gamma of unity while curves b and c are for gamma of one-half and gamma of two respectively. The reproduced brightness B_m' of any point of the image may be expressed in terms of the brightness of the corresponding point of the original B_m by the following relation

$$B_m' = k\, B_m^{\gamma}$$

throughout the region of brightness perceptible to the eye B_1, B_2, $B_3 \cdots\cdots\cdots B_n$, at the object. Since no methods of pictorial reproduction can quite cover most of the brightness ranges encountered in practice one must know the range of the system. The available brightness range in a reproduced picture is defined as the ratio of brightnesses of the image corresponding to the regions of the contrast curve where it is parallel to the horizontal axis. In photography the logarithm of this ratio is often used such as a range of 2 or 100 to 1.

Proper density concept applies directly to photography and when used in connection with television is somewhat a misnomer. Proper intensity would be a better term in television and it means that the intensity of some reference level must be properly chosen. This reference level may be either the subjective black or the brightest bright or the average brightness of the picture, depending on the method or rule by which "the proper" values for the four conditions mentioned above are determined.

Photography is very rich in rules for correct or proper reproduction. The oldest, and the most frequently referred to in handbooks on photography, is the "maximum range rule". By this rule the highlights of any picture should be represented by the lowest density and the darkest shadows by the highest density of the printing paper; gamma is adjusted to make it possible to meet these requirements. In other words according to this rule the contrast or gamma is varied depending on the range either of the negative or of the object. This rule, while certainly not a rule for obtaining a faithful reproduction, permits of achieving certain artistic effects and is widely used.

Another rule which is often called the correlation of highlights method, states that correct rendering is obtained by printing the highlight of the picture to a just perceptible density, while proper gamma or contrast is determined by a somewhat involved process. This rule works very satisfactorily over a limited range of brightnesses; namely high average brightnesses such as daylight landscapes, but seems to fail in some other cases.

There is a rule calling for correlation of subjective blacks plus modifying gamma and another one, recently proposed by Hopkinson, of correlating the subjective blacks while holding unity contrast.

The first of the two last rules is based on theoretical and experimental evidence and seems to give satisfactory results even at extreme conditions. The second, calling for unity gamma, gives correct and true representation at low levels of brightnesses of the original and has been successfully used in connection with making photographic records of street lighting installations for engineering purposes.

Correlation of subjective blacks calls for so adjusting the intensity of the reproduction that the parts of the original which appeared black are rendered black in the reproduction.

One of the newest rules has been statistically derived by Tuttle,[3] of the Kodak Research Laboratories, through cooperation with a number of amateurs. In essence, this new rule may be stated as follows: the developing and printing should be so controlled that overall subject-to-print gamma is slightly higher than unity, while the time of exposure of the paper is determined by a simple formula

$$t = K \frac{1}{T}$$

where t is the time of exposure, T is total transmission of the negative and K is a constant for any particular paper. If this rule, which

[3] C. M. Tuttle. Photoelectric Photometry in the Printing of Amateur Negatives. *Journ. Franklin Institute*, Vol. 224, p. 315 (Sept. 1937).

allows automatic printing, is followed the greatest number of customers are satisfied with the prints made of their negatives.

In black-and-white moving picture technique, it became a standard practice to make the overall object-to-image contrast between 1.4 and 2 in order to compensate for the lack of color. Television is also a monochromatic system and, therefore, has either to follow the experience of the older art and produce pictures with a similar increase in contrast, or to be content with photographically "flat" pictures. The expansion or increase in contrast may be applied either at the transmitter or at the receiver, but the interference consideration makes it more desirable to expand at the receiver. At the transmitter either linear or even logarithmic characteristics seem to be preferable. An inherent property of the eye is the cause of this preference. A small

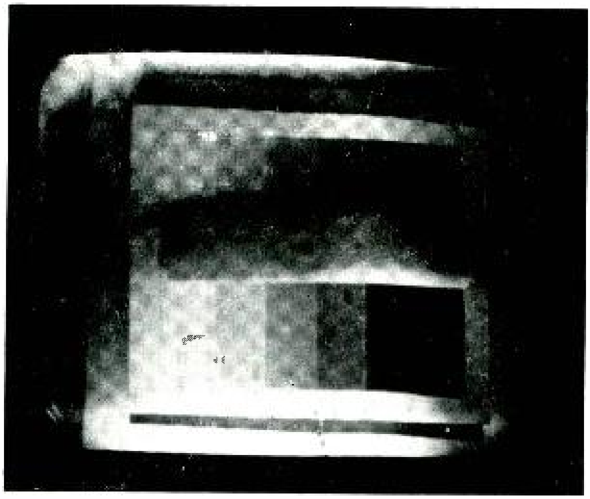

Fig. 3

increment in light intensity is more noticeable to the eye in dark parts of a picture than is the same increment in bright parts of the picture. When the picture at the receiver is expanded, the high lights are over-emphasized and shadows are under-emphasized, and in this way a greater amount of interference may be tolerated. Stating the same point differently; by use of either linear or, better still, logarithmic characteristic at the transmitter the ratio of demodulated signal-to-noise ratio is either maintained constant through the brightness range or increased in the shadows and decreased in the highlights. By the use of expansion at the receiver, the ratio of demodulated signal-to-noise is further increased in the shadows and decreased in the highlights. As was stated above, a smaller signal-to-noise ratio will pass unnoticed by the human eye in the highlights than in the shadows.

Gamma characteristics of a complete television system have been measured by the following method. A test object made of vertical strips of known reflection coefficients is placed before the Iconoscope camera. With normal studio lighting and the rest of the television system operating normally the brightnesses of the test strips are measured at the receiver. The appearance of the received test pattern is shown in Figure 3. Now if the brightness of the image is plotted against the brightness of the object on logarithmic coordinates a typical gamma characteristic is obtained. In Figure 4 a plot of such characteristics for a television system having linear amplifiers is shown.

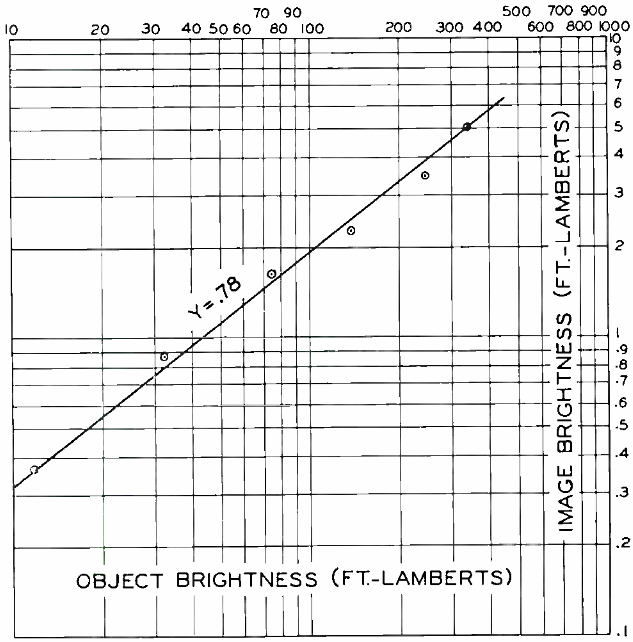

Fig. 4

Gamma characteristics of the Iconoscope itself may be obtained by taking an oscillogram of a single scanning line of a television signal obtained from the mentioned test pattern. The spurious signal or black spot signal of the Iconoscope requires the separation of a single line. This is accomplished by the use of an oscillograph equipped with a special circuit which is called "the line selector circuit". In Figure 5, a typical oscillogram of one line of a television signal is shown.

It has been found that, in general, the Iconoscope by itself is a low-gamma device. The value of gamma of the Iconoscope varies between 0.7 and 0.9 for most cases encountered in practice. Since a standard practice in the rest of the television transmitting circuit is to maintain a linear amplitude characteristic, the signal transmitted contains a certain amount of compression or, in other words, possesses a logarithmic characteristic to a certain degree, or γ under one. The amplifiers in the receiver are usually but not necessarily linear, so it is left to the Kinescope to expand the brightness scale to the desired exponential law. From the standpoint of its electrical characteristics, the Kinescope is a conventional vacuum tube and the three-halves-power law approximately expresses its current-voltage characteristic. Therefore, it may be said to possess a contrast or gamma of around 1.5. The light output vs. bombarding current characteristics of all known phosphors is linear only for a small region corresponding to the very low-current densities. Beyond this region, it shows a saturation effect which reduces the average contrast of the Kinescope to a value in the neighborhood of 1.2. The overall gamma of the system then for the type of Kinescope mentioned is around unity. Unity overall gamma of a television system is quite suitable for televising moving picture

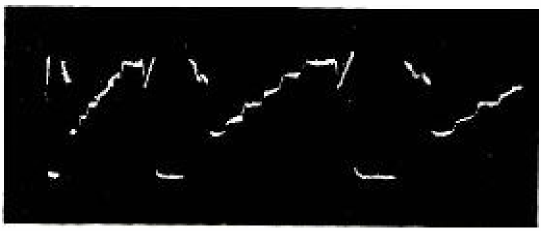

Fig. 5

films in which the contrast has been already raised to a proper value by an experienced photographer. Unity gamma is not sufficient for transmitting studio and outdoor pickup and for such occasions the contrast should be raised. It seems reasonable that by means of a logarithmic electrical amplifier the signal from the film be brought down to the same overall contrast say 0.8, and the Kinescope characteristic be raised to a gamma of say 2, with the resultant overall contrast of 1.6. Experimental Kinescopes with gamma of 2 or 3 have been built so that the above suggestion is not unreasonable. The RMA Standardizing Committee on Television has not as yet proposed any standard regarding the gamma characteristic of the transmitted television signal. In England it is intended that the transmitted signal will have unity gamma and all the desired correction will be made at the receiver.[4]

The brightness range in television images may be limited in various portions of the system. The present limit, however, is in the Kinescope. There is a limit in the signal produced within the Icono-

[4]Blumlein, Browne, Davis and Green. *Television and Short-Wave World*, Vol. 11, p. 285 (May 1938).

scope, but this is not a limiting factor at present. For certain marginal conditions of scene illumination, the noise produced in the Iconoscope output circuit may affect the range in the reproduced image. However, for optimum conditions of operation, this noise will not affect range, but may affect half-tone gradations. Again for marginal conditions of low field strength at the receiver, the noise produced in the input circuits may modify the range. Within the true primary-service area of the transmitter, this however is not likely to be the case.

The bulb shape of the Kinescope is determined by the physical characteristic needed to withstand atmospheric pressure. This results in a curvature of the portion of the tube on which the image is viewed. The luminescent screen is deposited upon this curved glass wall and is viewed through the glass. The curvature of the glass wall, the thickness of the wall, and the disposition of other portions of the tube permit the illuminated portions of the screen to throw light onto the non-illuminated parts. In addition to these reflections from the interior walls, a certain amount of light is totally reflected from the glass-air boundary on the outside surface, causing halation and reducing the range in details. Means have been found for reducing this effect in the conventional Kinescope. These means include control of the curved bulb walls, blackening of the walls, introducing a small amount of light-absorbing material in the glass wall under the luminescent screen, etc. A cathode-ray tube designed without special consideration of these factors has a range of about 50 to 1 between large areas and 10 to 1 in adjacent details. With careful design, these ranges can be increased. In some experimental tubes that have been built for a study of the effect of range on the television image, a thin piece of glass or mica has been used as the support for the luminescent material, and this screen has been mounted within a large glass envelope. With these special tubes, range has been improved to more than 100 to 1 between large areas and more than 50 to 1 between adjacent details.

In conclusion, it may be stated that the television signal is different from the sound signal in respect to amplitude and phase distortion. In sound transmission, phase distortion can be tolerated while amplitude or harmonic distortion can not. In television this situation is reversed; amplitude or harmonic distortion can be tolerated and even invited, but phase distortion must be eliminated to the greatest possible degree.

TRANSMISSION LINES AS COUPLING ELEMENTS IN TELEVISION EQUIPMENT

By

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Summary—This article deals with certain practical aspects associated with the operation of co-axial transmission lines, with particular emphasis on their television applications. The subject matter is divided into three major parts, of which the first contains a compilation (in simplified form) of all the important formulas required for use in transmission-line computations. The second part includes descriptions of simplified experimental procedures for the determination of the principal transmission-line characteristics, among which are surge impedance, attenuation factor and line loss, time delay and velocity of transmission. The third part is an analysis of means for coupling from amplifier chains or other vacuum-tube circuits into low-impedance lines at television video frequencies.

THE wide frequency band required for proper reproduction of television pictures, and the importance of phase (or relative time delay) considerations for the several wide-band amplifiers of a television system, have led to a new set of ideals for circuit performance as compared with sound reproduction systems. Not less important than the broad-band amplifiers, however, are the transmission lines which find extensive use as coupling elements between the various component parts involved in general television work.

The importance of transmission-line characteristics is readily appreciated by noting that a 35-foot length of ordinary shielded cable is of quarter-wave resonant length in the 3-to-4 megacycle region of the video range. Such a relatively short cable, when used as a video line, if not properly terminated, may give rise to reflections which nullify much of the effort expended on the amplifying networks which it connects. Other difficulties arise when either the attenuation or time delay of the line varies too much with frequency, in which cases the wave form of the video signal becomes intolerably distorted. In the main, this article is written from the standpoint of practical engineering applications, with an essentially non-mathematical approach. The standard definitions of transmission-line properties are found in Section III.

The types, applications, and various properties of transmission lines are discussed in connection with their use in television apparatus.

I. Types of Transmission Lines

There are three general types of transmission lines which probably will find application in television practice. The first is the familiar open-wire line, generally used to transmit energy to or from antennae. This type of line may be regarded as being balanced to ground. Often, the balanced arrangement is desirable for ultra-high-frequency antenna coupling lines, because of the widespread use of dipole or doublet antennae at the television carrier frequencies.

The second type of line, which also is balanced to ground, may take any of several forms. One of these is the familiar twisted pair, whose use at ultra-high frequencies is usually limited by its high loss. Another type of balanced line makes use of a unique expedient, in which two balanced and shielded concentric lines are placed back-to-back, with their shields connected together and to ground. The inner conductors of the two lines are used for coupling into the desired balanced element, usually a receiving or transmitting antenna.

The third type of coupling unit in this group is the coaxial cable, or self-shielded transmission line. This type of line, being unbalanced to ground, is useful to connect single-ended amplifiers and similar equipment. There are several types of coaxial cable. Of these, one is the "rigid pipe" which contains a small inner conductor, supported and spaced on disc-shaped insulators. Occasionally, the "pipe" is filled with nitrogen in order to deter the entrance of moisture. This type of transmission line is expensive, and usually is found in permanent installations where the line may be buried under a street or in the walls of a building, etc.

Probably the flexible cable is the type of line used most extensively in laboratory work. Flexible cable embodies a stranded inner conductor, rubber insulation, and external metallic braid. Its electrical characteristics (attenuation, time delay, etc.) are usually quite good, and its flexibility makes it desirable as regards convenience in general. Accordingly, this article is concerned chiefly with the shielded, unbalanced type of transmission line. Whether it be flexible or not is immaterial, of course.

II. Applications

The television applications of concentric cable are numerous. It is found at transmitters for connecting cameras and line amplifiers, synchronizing generators and sweep generators, and it is used as a video line from the studio or control room to the transmitter itself. The electrical length of the line may be considerable in some instances, particularly when the line carries the video range of frequencies.

The use of flexible cable as a transmission medium for test signals (video, i-f, synchronizing, and blanking) in television receiver factories is likely to be widespread. Moreover, in any establishment of presently anticipated size the electrical length of the cables will be great enough to require attention to the line characteristics.

III. Simplified Expressions for the Basic Properties of Transmission Lines

Since the physical structure of transmission lines results in a uniform distribution of inductance, capacitance, and resistance, the usual performance considerations of transmission lines involve electrical properties which are defined quite simply insofar as close approximations to fact are proper. For convenient reference, the simplified expressions are grouped in this Section. In all cases, these formulas are sufficiently accurate (usually within 1 per cent) for usual television laboratory application. Precisely correct formulas, with the derivations of these simplified expressions, occur throughout the following Sections.

The *surge impedance*, (or characteristic impedance), designated as Z_o throughout this article, is

$$\sqrt{\frac{\text{Total Inductance (Henrys)}}{\text{Total Capacitance (Farads)}}}\,.$$

Both the inductance and the capacitance may be measured with a 1,000-cycle bridge when the line is as much as several hundred feet long. Of course, the inductance is measured with the far end of the line shorted, and the capacitance is measured with the far end open. These measurements must not be made when the line is wound on a spool, or when the outer conductor (the shield) crosses itself or contacts other conductive objects, except in the case of a cable having a covering of insulating material.

The *time required for the transmission of energy from one end of the line to the other*, herefrom designated as $\overline{\Delta T}$ seconds, is

$\sqrt{\text{Total Inductance (Henrys)} \times \text{Total Capacitance (Farads)}}$. Since

$$Z_o = \sqrt{\frac{L_{\text{total}}}{C_{\text{total}}}}, \quad \text{it follows that}$$

$$\overline{\Delta T} = C_{\text{total}} \times Z_o, \text{ or, } \overline{\Delta T} = \frac{L_{\text{total}}}{Z_o}.$$

The *velocity of wave propagation*, designated here as V (meters per second), is

$$\frac{\text{Length of line in meters}}{\sqrt{L_{\text{total}} \times C_{\text{total}}}}.$$

The *phase shift* (delay) θ, in degrees, at a given frequency, is $\overline{\triangle T} \times$ frequency in cycles per second $\times$ 360.

One way of dealing with the *attenuation* caused by the inevitable resistance of a transmission line is first to consider the usual case where the line is terminated in a resistance which is equal to the surge impedance Z_0. In this case, the energy delivered by the generator is gradually attenuated as it is propagated, and all energy reaching the resistance load (Z_0) is dissipated therein without reflection back into the line. In this case of proper line termination, the voltage across the load resistance is

$$\frac{\text{Voltage impressed on the line}}{\text{Anti}\log_{10}\dfrac{(\text{db})}{20}},$$

where (db) has its usual meaning, line loss in decibels.

A generally suitable, and often preferable method of measuring the line-loss factor (db), is to cut a piece of line for quarter-wave resonant length, and then to measure its input resistance with the far end open. This can be done by inserting the near end in a tuned circuit, and by observing the effect on Q. (A detailed description of this method is given in Section VII.) Resulting from this method, the loss (in db) on a properly terminated line is 8.686 $\times$ Input resistance of quarter-wave line with far end open, divided by surge impedance.

At this point, it is interesting to note that in the case of a line having rubber dielectric, the actual length of a quarter-wave resonant line is usually from 55 to 60 per cent of a quarter-wave length in free space or of a line having air dielectric.

Having measured the loss factor (db) for a quarter-wave length of line, the loss for any other length of properly terminated line is at once determined from the fact that line loss is proportional to the length of the line.

Since the attenuation of a transmission line is not independent of frequency, the higher frequencies suffering greater attenuation than the lower, a considerable length of line has an appreciable effect on the over-all frequency-response characteristic of a television system. Ordinarily, this effect does not present serious difficulty, and may be readily compensated in various parts of the system.

Of greater importance is the matter of transmission-line reflections, when the line is not properly terminated. These reflections result in multiple images in case the line transmits video signals. Moreover, reflections make measurements (with steady-state sine-wave potentials) misleading and difficult to interpret.

Energy flowing in a transmission line, regardless of the load at the receiving end (termination), is attenuated logarithmically according to the law: $E_{out}/E_{in} = \varepsilon^{-\alpha}$. When reflection occurs at the receiving end (terminating load not a resistance equal to Z_0), the reflected energy flows back toward the sending end, and is subjected to the same logarithmic attenuation. After its return to the sending end, it may be entirely dissipated by the internal impedance of the sending "generator" (when that impedance is a resistance equal to Z_0), or, some of the energy may be reflected back toward the receiving end, suffering logarithmic attenuation again as it flows along the line. When the energy originally delivered to the line at the sending end is a simple and continuous sine wave, the energy reflected back from the receiving end may arrive at the sending end with such phase that it reinforces the impressed wave. Similarly, it may arrive in phase opposition, or it may have any intermediate phase, depending on the time required for a round trip. This is the explanation of the familiar standing waves on improperly terminated lines. However, due to the fact that television signals seldom (if ever) contain repetitious components at frequencies greater than two or three times the line frequency (13,230 cycles/sec), standing waves on video transmission lines having the order of length found in the usual television laboratory are essentially meaningless. It is preferable (and correct) to visualize the original pulse (or wave) as being propagated to the terminating end of the line, and suffering the normal attenuation effected by the entire length of the line, regardless of the type or magnitude of the terminating impedance. If a reflection occurs (due to improper termination), the reflected pulse is an electrical echo, distinct and apart from the original signal. This view clarifies the reason for the distinct multiple images on a television screen when incorrect terminations of transmission lines give rise to reflections. Obviously, when the terminating resistance of a line is of such magnitude that it absorbs all of the received energy, there can be no reflection whatsoever.

However, in case the internal impedance of the generator is a resistance which is equal to the surge impedance of the line, whatever energy is reflected from the receiving end is entirely absorbed at the sending end and, hence, cannot again travel along the line to cause a delayed (or echo) signal at the receiving end. Accordingly, multiple images are prevented by a proper terminating impedance at either end of the transmission line. Occasionally, and as a double precaution, each end of the line is terminated in a resistance equal to Z_0, but this results in decreased line voltage and is seldom necessary.

IV. Measurement of Transmission Line Surge Impedance

The rigorous expression for surge impedance is

$$Z_o = \sqrt{\frac{r + j\omega L}{g + j\omega C}} \tag{1}$$

where r, L, g, and C are respectively the series resistance and inductance of the conductors, the shunt leakage conductance, and the capacitance, all measured per unit length of line. Eq. (1), developed for a line having both shunt and series loss, shows that Z_0 varies with frequency and that it is not a pure real quantity; i.e., it has a reactive component. This may appear to present some practical difficulties, due to mis-match, in view of the well-known law which states that "a line must be terminated in an impedance equal to its surge impedance, in order to avoid reflections from the receiving end." Since Z_o varies with frequency, the question naturally arises as to what impedance should be used for proper termination, particularly as regards television where a wide range of frequency is involved.

The situation may be considered most directly by rewriting (1) in a slightly modified form, viz:

$$Z_o = \sqrt{\frac{L}{C}} \sqrt{\frac{1 + \dfrac{r}{j\omega L}}{1 + \dfrac{g}{j\omega C}}} \tag{2}$$

It can be shown that Z_o may be taken as $\sqrt{L/C}$ for all usual transmission lines, on the basis of two practical conditions. First, for all frequencies greater than (say) 10 kc, $r/\omega L$ and $g/\omega C$ are negligible in comparison with unity, in which case $Z_o = \sqrt{L/C}$. Secondly, $r/\omega L$ being small in comparison with unity above 10 kc, $g/\omega C$ in many cases is nearly independent of frequency and small in comparison with unity; the dielectric medium in the cable has a power factor which is nearly constant as the frequency is varied.

At very low frequencies (say, less than 200 cycles/sec), the resistive terms in (2) may not be negligibly small in comparison with the reactances; hence, the surge impedance (Z_0) may vary within wide limits, since at extremely low frequencies Z_0 approaches $\sqrt{r/g}$. At first thought this variation of Z_0 may seem to portend difficulties due to mis-match of impedances at the very low-frequency end of the video range, but, practically, there is no such effect, because any actual transmission line used for television purposes is electrically very short at these frequencies, (at 60 cycles/sec, one wavelength is 5,000,000 meters), and of course mis-match is relatively unimportant when a line is electrically short.

Accepting $\sqrt{L/C}$ as a suitably close approximation of Z_0, measurements of L and C are best carried out with the usual laboratory type of 1,000-cycle bridge. L is measured with the far end of the line shorted, and C is measured with the far end open.

Several precautions should be observed in order to obtain Z_0 without appreciable error. It is important to use a sufficiently long piece of transmission line to permit reasonably precise measurements of L and C. (The inductance of a typical cable used in this laboratory is approximately 0.1 microhenry per foot, and the capacitance is 25 $\mu\mu f$ per foot.) Hence, while the line capacitance can be determined quite accurately with only a short length of cable, the measurement of L requires a line of sufficient length to bring the inductance to an easily measurable value.

Another precaution to be taken involves an effect observed with a flexible cable (having a metallic outer sheath) wound on a spool. It was found that the measured inductance of the line increased four-fold at 1,000 cycles/sec when the return current was permitted to flow across the adjacent turns of the coiled cable, rather than to flow back along the outer conductor, as it would do in case the line were uncoiled and laid out with no contacting points on the outer conductor. The inductance of an uncovered cable should not be measured when the cable is wound on a spool and if the frequency of the bridge signal is less than about 10 kc. At higher frequencies, substantially all of the return current flows along the inner side of the outer conductor; hence, it is then immaterial whether the cable is coiled or not.

Another method of measuring surge impedance is perhaps more direct than the foregoing, but it requires a good oscilloscope and a signal of recurrent rectangular wave form. The transmission line under test (which should, in this case, be at least 500 feet long) is fed from a source of rectangular pulses, such as the horizontal line-frequency synchronizing signals used in television applications. The line is terminated at its far end by a variable resistance whose order of magnitude is that of the anticipated surge impedance value. The oscilloscope, which must be capable of accurately portraying the input wave shape, is connected across the input terminals of the line under test. Figure 1 shows a circuit diagram of the system.

This method of measuring Z_0 is predicated on the fact that when a transmission line is terminated by an impedance equal to its surge impedance, no reflection can occur from the thus terminated end. A pulse applied to the input of the line travels to the far end. In case some of the pulse energy is reflected (due to improper line termination), it arrives back at the input end of the line with a time displacement relative to the originally impressed pulse, and the oscilloscope shows a wave form which differs from that of the original pulse alone. When the line is properly terminated, no reflection occurs, and the oscilloscope shows only the original signal with its rectangular wave form.

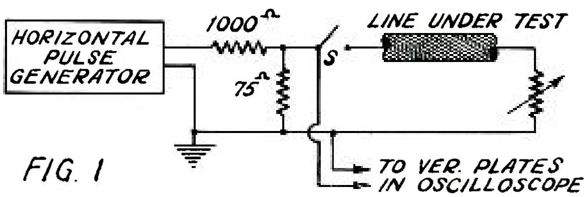

FIG. 1

Practically, the mode of procedure is quite direct. The terminating resistor is adjusted until opening or closing the switch S (Figure 1) results in no change in wave form as seen on the oscilloscope. The value of resistance which accomplishes this condition is taken as the surge impedance of the line. Different values of molded carbon resistors, or a decade resistance box, may be used for the adjustable termination. For example, it has been found that the small reactive components of impedance in a General Radio decade box (10, 1, and 0.1-ohm steps) are not sufficient to cause difficulty in precise determination of a 65-ohm line.

Note that the regulation of the source of pulses must be equivalent to several hundred ohms (see 1,000-ohm resistor in Figure 1). Otherwise, reflected pulses will not alter the wave form enough to provide an accurate determination of Z_0.

V. Measurement of Attenuation

The attenuation factor may be straightforwardly determined. It is seldom independent of frequency, and often is roughly proportional to frequency.

To measure α, steady-state sine-wave energy is applied to the line under test, and reflections are allowed to exist. When the length of line and the applied frequency are discreetly chosen, standing waves accumulate to a steady-state condition. Their strength, together with the known frequency and length of line, serve in the determination of α (for that frequency) and also for determination of the time delay $(\beta/2\pi f)$ for that length of line. β is known as the phase-delay constant, and is invariably expressed in radians; hence, β divided by the angular velocity $(2\pi f)$ is the time delay (in seconds), if it is in cycles/sec.

It seems sufficient to say that for all usual transmission lines the time required for an impulse to travel from one end of the line to the other is the square root of the product of the total inductance and total capacitance, as found in Section III. Accordingly, and since a quarter-cycle corresponds to $\pi/2$ radians, the frequency (f_4) at which a line requires the time of a quarter-cycle as the transmission time is as follows:

$$f_4 = \frac{\pi}{2} \times \frac{1}{2\pi} \times \frac{1}{\sqrt{LC}} = \frac{1}{4\sqrt{LC}}.$$

When this frequency is applied to the line under test, quarter-wave resonance obtains, and the accumulated reflections ordinarily result in a steady state voltage (at the open end of the line) which is many times greater than the impressed voltage.

It is obvious that the phase relations among the reflections must be quite precisely correct in order that the full resonant rise of voltage shall be realized. Consequently, it is important to cut the length of line correctly for the applied frequency, or to adjust the applied frequency and to measure the length of line accurately. Also, it is self-evident that the resonant rise of voltage depends upon the attenuation of the line, since a small value of α must result in correspondingly great reinforcement by the reflected waves. This dependence of resonant rise on attenuation affords a method of determining α, as follows.

Quarter-wave resonance is established, as discussed above, and by means of tube voltmeters the ratio $\dfrac{\text{Impressed Voltage}}{\text{Resonant Voltage}}$, or, $\dfrac{E_{in}}{E_{out}}$ is measured. This ratio is equal to α (within 1 per cent when α is less than 0.35), and therefore may be used for all usual transmission lines.

While the resonant rise method of measuring α may appear to be direct and simple, in practice it requires a low-impedance source of signal having very small harmonic content, and vacuum-tube voltmeters which are accurate at both high and low levels of r-f voltage.

Other methods for measuring α are based on the input impedance of a quarter-wave line. When this impedance is denoted by Z_{in}, the surge impedance being Z_o, $\alpha = \dfrac{Z_{in}}{Z_o}$ within 3 per cent when α is less than 0.3.

A preferred method for measuring Z_{in} is to insert the near end of a quarter-wave line (far end open) into a tuned circuit, and to note the resulting decrease in Q of the tuned circuit, either by direct Q measurement or by a resistance-substitution procedure. This method for determining Z_{in} is shown on Figure 2. An 1851 tube drives a

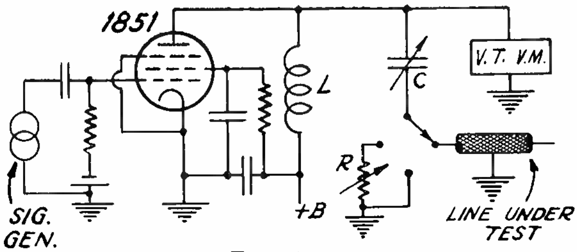

FIG. 2

tuned circuit having a low L/C ratio. An essentially sinusoidal tank-circuit current is desired. The procedure is reasonably simple, and involves resonance to decrease harmonic strengths and to increase the line input voltage to a level at which it can be measured accurately by a vacuum-tube voltmeter. The signal-generator frequency and the tank-circuit capacitance are first adjusted by trial, until the insertion of the transmission line causes no detuning, thus establishing quarter-wave resonance on the line simultaneously with tank-circuit resonance.

With f and C set, the 1851 grid voltage is adjusted for a selected deflection on the vacuum-tube voltmeter. (This deflection should be in the square-law region for precision of measurement.) The input voltage to the 1851 grid, E_o, and the deflection, d_o, are noted. The line is then inserted, and a new deflection, d_1 is noted. A check is effected by increasing the grid voltage to E_1, restoring the deflection d_o. Let R_o be the tank-coil resistance (which need not be known). Z_{in}, the input impedance of the transmission line under test, is a pure resistance (a necessary condition at quarter-wave resonance).

Then:

$$Z_{in} = R_o \left(\frac{E_1}{E_o} - 1 \right) = R_o \left[\left(\frac{d_o}{d_1} \right)^n - 1 \right] \qquad (3)$$

where $n = 1/2$ for a square-law tube voltmeter. The law of deflection should be checked if the second part of (3) is to be used. In case R_o is known from an accurate Q measurement, (3) may be used directly.

In general, Z_{in} may be more precisely determined by substituting small molded carbon resistors for the transmission line, and by noting the effects on the tuned circuit. 10-ohm, ¼-watt resistors may be used in parallel for the substituted elements, since such resistors hold their d-c values of resistance up to extremely high frequencies. Precision of measurement is favored by plotting a curve of tank-circuit voltage against substituted resistance.

The attenuation factor α may be determined from Z_{in} and Z_o by any of the following formulas:

$$\alpha = \tanh^{-1} \frac{Z_{in}}{Z_o}, \tag{4}$$

an exact expression.

$$\alpha = \frac{1}{\sqrt{\left(\frac{Z_o}{Z_{in}}\right)^2 - 1}} \tag{5}$$

within 1 per cent when $\frac{Z_o}{Z_{in}}$ is greater than 3.5

$$\alpha = \frac{Z_{in}}{Z_o} \tag{6}$$

The following table shows the departures from the correct values of α [as obtained from (4)] incurred by using (6). In each case, (6) gives a value of α which is less than the correct value.

α	*Per cent error*
0.1	0.3
0.2	1.3
0.3	3.0
0.4	5.0

α may be converted into db directly by multiplying by 8.686

VI. Measurement of Time Delay

The simplest method for determining time delay is to use the relation given in Section III, i.e., $\overline{\triangle T}$ seconds $= Z_o \times$ line capacitance in farads. However, $\overline{\triangle T}$ can be determined from the physical length of line required to establish quarter-wave resonance at a known frequency f. Since that length of line is used in the measurement of α, the following formulas may be found convenient.

Velocity of propagation along the line (meters per second) =

$$\frac{4 \times \text{Physical length* (meters)} \times \text{Velocity of Light (meters per second)}}{\text{Wave length of Signal (meters at quarter-wave resonance)}}$$

Time delay in seconds, $\overline{\triangle T} =$

$$\frac{\text{Length of line in meters}}{\text{Velocity of transmission (meters per second)}}$$

Phase delay in degrees = Time delay (sec.) $\times$ Frequency (cycles/sec) $\times$ 360.

VII. Experimental Results

Tests were made by the foregoing methods on a typical shielded cable used in this laboratory for coupling television apparatus. A brief summary of the results is included here, principally to indicate the orders of magnitude of the characteristics likely to be found for similar lines.

The time delay is essentially constant from 0.5 to 6 megacycles (the range covered by the measurements), at approximately 0.17 microseconds per 100 feet of cable.

The value of α, which, of course, varies with frequency, was found to be 0.04 nepers per 100 feet at 0.7 Mc, and 0.15 nepers per 100 feet at 5.6 Mc. These values correspond respectively to 0.35 db and 1.3 db loss, when the line is terminated in 65 ohms (correct termination). α does not increase linearly with frequency; hence, the db loss per wavelength is not constant throughout the range. α was found proportional to $(f)^{.6}$; hence, a decrease in db loss per wavelength results from an increase in frequency.

VIII. Circuits for Feeding Transmission Lines

A properly designed line-coupling stage must fulfill at least two requirements. The first is the isolation of the center conductor of the line (whose outer conductor is at ground potential for d.c.) from the d-c plate supply for the coupling stage. Secondly, the frequency and phase response of the unit must not be affected materially by the use of the tube-to-line coupling condenser.

* To yield quarter-wave resonance.

Line-coupling stages may be divided into two groups: (1) circuits in which the line is fed from the plate of the tube, in which case a blocking condenser must be used, and (2) cathode-loaded coupling tubes (also known as cathode followers) in which cases the line is generally not isolated for d.c.

The latter group (cathode followers) may be covered merely by stating that the usual considerations applied to feeding a low-impedance load from the cathode of a tube apply equally well in this case. Since no coupling condenser is used, the low-frequency response is not affected by the use of this type of line coupler.

Line-coupling stages of the first group may consist of either a low R_p tube (or tubes), or a high R_p pentode, feeding the line through a blocking condenser. The first case (low R_p triodes) is an approach to the constant-voltage drive condition, whereas the use of a pentode permits an approach to the constant-current condition. In both cases the d-c potentials are applied to the output tube plates by parallel-feed resistors.

At this point it should be noted that the low-frequency characteristics of any of these networks cannot be predicted merely on a basis of frequency response to a monotone driving signal, because the nature of the low-frequency video signals is such that a precise determination of amplitude and phase response for the frequencies contained in a typical rectangular pulse would be required. As an example of this aspect of low-frequency performance, it may be noted that a certain plate-to-line coupling unit which showed (on a monotone basis) only ¼ of 1 per cent loss in response at 60 cycles, caused a variation in the peak amplitude of a 60-cycle square wave of 15 per cent during only 1/120 second.

The criterion for proper low-frequency response may be taken as the ratio of the time constant CR of the coupling circuit, to the period of the lowest frequency to be transmitted (generally 60 cycles/sec). C is the capacitance of the blocking condenser (farads), and R is the total resistance (dynamic) seen by the blocking condenser. R is equal to the sum of the line input impedance and the parallel combination of plate resistance and feeding resistor.

In general, the coupling-circuit time constant should be 3 to 20 times the period of the lowest frequency to be transmitted. The lower CR values are permissible in cases where subsequent low-frequency compensation is used, either in the plate circuit of the tube following the transmission line, or at the output end of the cable. CR values of 20 or more times the period of the lowest frequency will be required when no such compensation is used, and when several stages with relatively small time constants are included in the same amplifier chain.

FIELD STRENGTH MEASURING EQUIPMENT FOR WIDE-BAND U-H-F TRANSMISSION

BY

R. W. GEORGE

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Summary—*Mainly in the interest of television service, some quantitative data were required on the characteristics of ultra-high-frequency propagation paths. Portable measuring equipment and methods devised to obtain these data are described.*

Transmitter systems giving constant output over the ranges of 81 to 86 Mc and 140 to 145 Mc were used, the frequency of each transmitter varied at the rate of 166 kc per second from one extreme of its range to the other. Field strength was measured at each 70-kc increment of the frequency range, a measurement being recorded automatically as the signal frequency passed through the receiver pass-band. The circuits were arranged to hold the recorder indication until changed by the following measurement. About 70 measurements were recorded in one-half minute by means of a novel automatic tuning arrangement.

The signal generator and calibration methods used are also described.

FIELD strength versus frequency measurements afford a means of studying the characteristics of propagation paths, especially where the difference between direct and indirect path lengths is relatively small. It is the object of this paper to describe some equipment with which such measurements were made during a recent study of ultra-high-frequency transmissions from the Empire State Building.[1] In general, the problem was resolved to one of measuring and recording received field strength as the transmitter frequency varied at a constant rate of change over a 5-megacycle range. Measurements over this frequency range are sufficient to indicate the existence, and some important characteristics, of indirect paths which have time delays long enough to cause distortion in television reception.

In the past, one method of making such measurements has been to adjust manually the transmitter and receiver frequency step by step for each measurement to be made.[2] This method is obviously slow and requires perfect coordination between the transmitter and

[1] "A study of U-H-F Wide-Band Propagation Characteristics" by R. W. George, *Proc. I. R. E.*, Vol. 27, January, 1939.

[2] P. S. Carter and G. S. Wickizer, "Ultra-High-Frequency Transmission Between the RCA Building and the Empire State Building in New York City", *Proc. I.R.E.*, Vol. 24, August, 1936.

receiver operators. For the purpose of making a rather extensive survey, it was considered desirable to speed up the operation to permit a greater number of measurements to be taken, and to record automatically the measurements in a form readily utilized.

One relatively simple measuring system to accomplish this result might consist of a wide-band receiver associated with a suitable recorder. Such a receiver, having constant response over a 5-megacycle range has the disadvantage that its inherently high noise equivalent would not permit satisfactory recording of weak signals. Otherwise, this method would be very convenient as no tuning of the receiver to the changing transmitter frequency would be necessary. The measuring equipment to be described was used to record a large number of measurements over the frequency range with a result similar to that which would be had if continuous measurements with frequency were made. It also embodied a conventional type of receiver which permitted measurements at low field intensities. By making the transmitting and receiving equipment conform to the following conditions, the recorded data required no corrections in order to determine the field strength at a given frequency.

The transmitting systems were so designed that the radiated power was constant as the frequency varied from 81 to 86 megacycles or 140 to 145 megacycles[3]. The transmitted frequency varied at the rate of 166-kc change per second and required one-half minute to go from one extreme of the frequency range to the other. By the use of half-wave doublet receiving antennas having substantially constant response over the desired 5-megacycle range, connected to a 75- or 100-ohm transmission line terminated with resistance at the receiver, constant voltage was delivered to the receiver for constant field strength at the antenna.

An ultra-high-frequency, triple-detection receiver having a 100-kilocycle i-f pass-band was adapted by ganging the ultra-high-frequency detector and heterodyne-oscillator controls and adjusting the circuits to extend the 5-megacycle signal range over the major portion of the control dial. The response of the receiver was adjusted to be constant over the required tuning range. It was not considered feasible to incorporate conventional automatic-frequency control of the ultra-high-frequency circuits because of the required flexibility of the measuring system. An automatic step-by-step frequency control, coordinated with the measuring system, was used which will be understood more clearly after the following description of the functions of this system.

[3] G. L. Usselman, "Wide-Band Variable-Frequency Testing Transmitters," published in this issue of RCA Review.

The most suitable time to make a measurement would be when the signal, changing in frequency, was at the mid-band of the intermediate-frequency amplifier. Thus, the signal is in the receiver about three-tenths of a second before it reaches the middle of the pass-band which gives the diode time to reach an output corresponding to the input signal. The operation of the measuring functions is indicated in Figure 1. An audio tone which was utilized to cause a measurement to be made, was obtained by combining, in a separate detector, the intermediate-signal frequency and a beat oscillator tuned to the mid-band frequency of the second intermediate-frequency amplifier. This tone, fed through an audio amplifier and associated rectifier, caused the relay R_1 to connect the diode to the grid of the first d-c amplifier tube for the duration of the tone which was about 0.12 second. The small condenser C_1 was charged to the corresponding diode voltage and maintained this voltage on the grid of the first d-c amplifier tube after the signal had passed out of the audio-frequency pass-band and the diode was disconnected. To accomplish this, it is important that the grid current of the first d-c amplifier be extremely low, a condition that was easily obtained by the use of a G.E., FP-54 type tube. It was necessary for the capacity of C_1 to be rather small, on the order of 0.01 microfarad, for it to be fully charged in the short charging time allowed. With this arrangement, over an hour was required for the grid voltage to decay to 37 per cent of its original value and, of course, no appreciable change occurred between measurements. The d-c amplifier had a linear input-output characteristic and operated a conventional recorder. This method of holding the recorder at its last measurement until a subsequent measurement is made, was advantageous in that a sequence of measurements gave an approximately smooth curve. It also gave the recorder-movement time to indicate the true amplitude of each measurement.

With the system as described up to this point, it will be apparent that if the receiver is tuned to say the low-frequency end of the signal-frequency range, a measurement will be made as the transmitter frequency, increasing from its minimum value, passes through the pass-band of the receiver. After this measurement, if the receiver is quickly tuned to a higher frequency, the increasing signal frequency will again come into the audio-frequency pass-band and cause another measurement to be made. By carrying on this procedure a complete series of measurements can be made. In advancing the receiver tuning, the receiver is tuned through the signal frequency, but this can be done so quickly that the signal is not in the pass-band long enough to cause a false measurement to be made. Tuning by hand in this manner permitted about fifty measurements to be made over the 5-megacycle range in one-half minute. These measurements were spaced with fairly equal frequency increments according to the skill of the operator. The automatic receiver-tuning means permitted about 70 measurements to be made with substantially equal frequency increments between measurements, and when the relay R_1 returned to normal signifying that a measurement was completed, it caused another relay, R_2, to be operated which in turn caused the receiver tuning to be changed by means of a modified automatic-telephone circuit-selector mechanism. This tuning device simply notched up the tuning controls to a higher frequency by approximately 70-kilocycle incre-

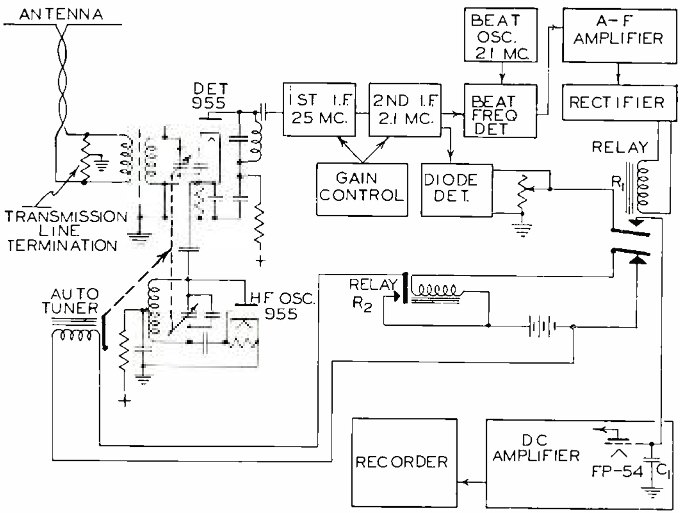

Fig. 1—Diagram showing elements and functions of the measuring system, except the signal generator.

ments. At the instant an incoming signal caused relay R_1 to operate, the holding current in relay R_2 was removed which in turn caused current to be supplied to the auto tuner. The auto tuner was so arranged that the application of current in its coil caused a pawl to engage an advanced tooth on an associated ratchet which was geared with the tuning controls. As the signal frequency increased, the audio-beat note passed out of the response band of the a-f amplifier and thus restored R_1 to normal as shown in Figure 1. It will be noted that this last operation first disconnected the diode from the d-c amplifier and then applied current to the relay R_2. This removed the current in the auto tuner, the armature of which in being restored to normal by a spring caused the receiver tuning to be changed by means of the pawl and ratchet.

As a matter of uniform procedure, the system was designed so that measurements were taken with increasing frequency. Thus, a few minutes of observation indicated the proper gain of the receiver to use for the range of signal levels available and the measurement was started by disengaging the automatic-tuning drive and tuning the receiver to the lowest frequency. The automatic-tuning drive was

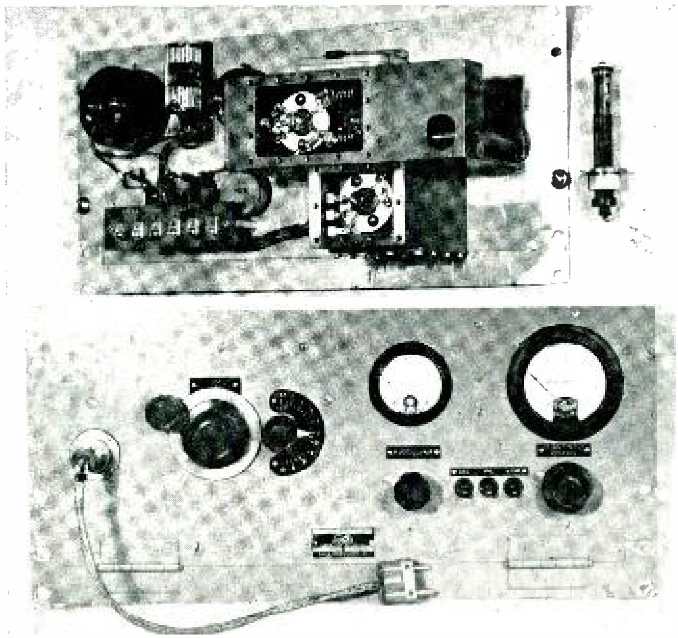

Fig. 2—Standard u-h-f signal generator. In the top view (rear) the shield covers are removed exposing the u-h-f oscillator in the upper compartment, and the voltmeter in the lower compartment. Through the circular opening can be seen part of the inductance, L_1. At the extreme right is shown the movable element of the attenuator.

then re-engaged and the series of measurements started when the signal began increasing in frequency after reaching its lowest frequency.

With this system, it was only necessary to keep the transmitter-frequency sweep in continuous operation and to observe a predetermined schedule for polarization of the transmission. Maximum stability of the measuring system was obtained by the use of voltage regulated-power supplies. Normally the equipment was operated from 110-volt a-c sources.

Calibration of Equipment

For all such measurements, some means of comparing data with a common reference is necessary. The signal generator used as a reference also provided a fairly accurate means of measuring field strengths. This means was embodied in the 74-ohm output circuit of the signal generator which is known to be substantially equivalent to the impedance of a half-wave doublet antenna. Thus, by substituting the signal generator for the antenna, the equivalent voltage in the antenna was known, which with its known effective height provided the corresponding field strength indicated by the receiver output. As it was somewhat cumbersome to check the overall gain of the measur-

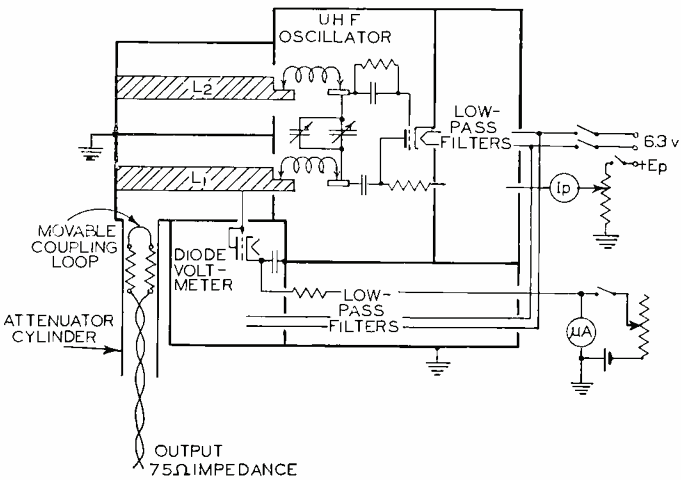

Fig. 3—Diagram of u-h-f signal generator.

ing system frequently by this method, it was found practical to measure the transmission-line loss and calibrate the measuring system by connecting the signal generator to the receiver-input terminals, including the transmission-line termination, whenever a calibration was necessary.

The signal generator also provided means of checking the input-output response and the flatness of response over the frequency range of the measuring system. Final evidence that the entire system had constant response over both frequency ranges was obtained by making measurements in an open, unobstructed field where no serious indirect-path waves, with the exception of the wave reflected from the ground, were expected. The path length of the wave reflected from the ground was so nearly equal to the direct wave-path length that it caused no appreciable field strength change over the 5-Mc range with the result that the recorder indicated substantially constant field strength.

Details of the ultra-high-frequency signal generator, Figure 2, are given in the diagram, Figure 3. It will be seen that two sections of concentric line, L_1 and L_2, comprise part of the inductance in the ultra-high-frequency oscillator circuit. The voltage across a portion of the inductance, L_1, can be set to calibrated values by means of the vacuum-tube voltmeter and plate-voltage control of the oscillator. Output from the oscillator is taken by inductively coupling a small loop with the maximum-current end of the inductor L_1. The coupling

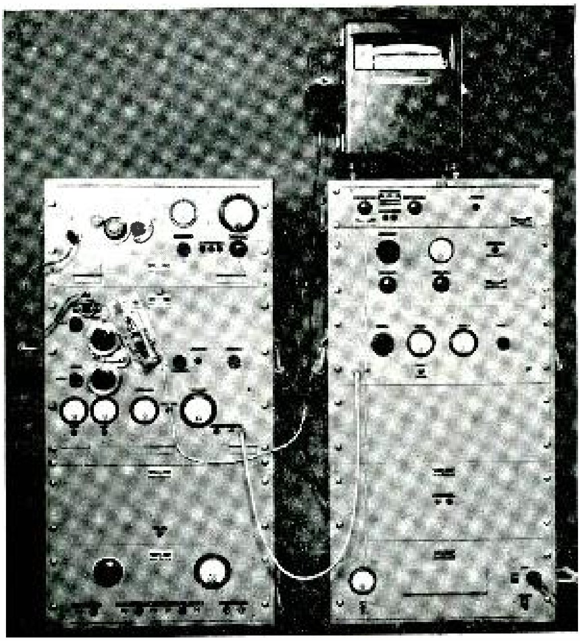

Fig. 4—Measuring equipment in portable racks. The units are, beginning at the top of the left-hand rack, standard signal generator, u-h-f receiver with the automatic-tuning device attached, power-supply unit, and spare audio amplifier. In the right-hand rack are, the recorder (on top), final d-c amplifier, FP-54 d-c amplifier, tone rectifier, regulated power supply, and a-c voltage regulator.

loop is connected to a short 74-ohm transmission line with a 37-ohm resistor on either side thus providing a 74-ohm output circuit. The coupling of the loop, and thus the output, is controlled by a calibrated threaded movement which moves the coupling loop inside a metal sleeve. This attenuator gives a range of 60 db with about 1¼ inches displacement, the calibration of which is substantially independent of frequency. The signal generator was calibrated by comparison with other standard-signal generators at 30 and 40 megacycles. Some satisfactory calibration checks have also been obtained with field-strength measurements at frequencies between 80 and 100 megacycles compared with field-strength measurements made with half-wave antennas incorporating a current- or voltage-measuring means at the center. A frequency range of from 30 to 200 megacycles is obtained by the use of two sets of plug-in inductance elements.

MISCELLANEOUS

Several items were found useful in facilitating measurements in the field. Simple attenuator pads made up with quarter-watt carbon resistors, were inserted in the transmission line at the receiver to obtain a loss of as much as 24 db. These were necessary in some cases

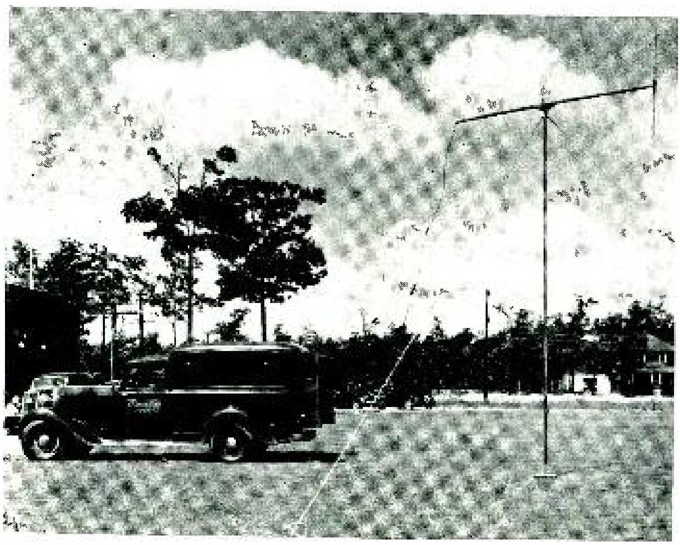

Fig. 5—Portable mast set-up, showing arrangement of antenna and transmission line.

where the received signal was several hundred millivolts. The frequency-response characteristic of these pads was found to be quite satisfactory over the required frequency range and their loss was conveniently measured by means of the signal generator.

The measuring equipment, Figure 4, excepting the recorder, was mounted in two portable racks which were about 40 inches high and accommodated standard 19-inch panel units. Suitable clamps were provided to fasten the racks in place in a ¾-ton panel truck. Measurements were made without removing the equipment from the truck whenever feasible.

The antenna-mounting means were extremely flexible in order to support the antenna satisfactorily at a large variety of receiving locations. A sectionalized mast, Figure 5, was provided which could be set up at various heights up to 22 feet. The top of the mast supported the center of a 10-foot cross-arm along which the transmission line was draped. The antenna made of two ¼-wavelength aluminum tubes was mounted by a bakelite clamp which in turn was supported on the end of a pole or at one end of the cross-arm on the mast. The cross-arm was mounted so it could be turned through 90 degrees by means of ropes, enabling easy control of the antenna polarization. The position of the antenna could also be changed over a circle of 5-foot radius by simply turning the mast, the three guy ropes being fastened to a movable collar at the top.

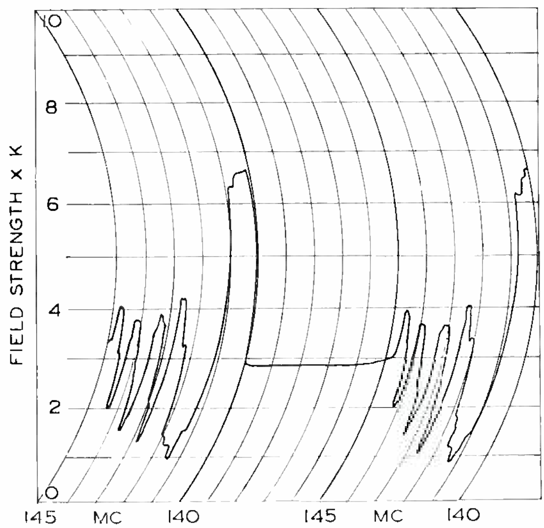

Fig. 6—Two frequency versus amplitude curves recorded under unchanged conditions.

The transmission lines used were of the twisted pair, Latox rubber-insulated type. For required lengths under 60 feet, a 74-ohm (characteristic impedance) line was available which had a loss of 10½ db per 100 feet at 83 megacycles and about 15 db per 100 feet at 145 megacycles. Longer lines sometimes used, having a characteristic impedance of 100 ohms, had a loss of 6½ db per 100 feet at 83 megacycles and 9½ db per 100 feet at 145 megacycles.

The recorder was critically damped and followed the changing signal levels quite accurately except in one case where extremely large and fast variations were encountered. To insure a uniform time or frequency scale, the chart was driven at the rate of three inches per minute by a small synchronous motor.

In general, the measuring equipment was quite satisfactory and gave consistent repeat measurements. This performance is illustrated by the two samples of recorded data shown in Figure 6 which were obtained in sequence with unchanged receiving conditions. The time interval between curves is one-half minute which was required for the transmitter frequency to return from 145 Mc to the starting frequency of 140 Mc. During this interval, the receiver tuning was changed to 140 Mc and the automatic-tuning device re-engaged so that measurements were repeated when the transmitter frequency began increasing from 140 Mc. Since only one-half minute was required in which to make a complete series of measurements, a large number of measurements could be made at each location with a variety of antenna positions.

ACKNOWLEDGMENT

The measuring methods embodied in the described equipment were developed under the guidance of Mr. H. O. Peterson. Mr. K. G. MacLean gave valuable assistance in adjusting the equipment to give the desired performance.

MEASUREMENT OF PHASE SHIFT IN TELEVISION AMPLIFIERS

By

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Summary—*The importance of phase-shift characteristics of the amplifiers in television receivers and signal generating systems has created the necessity for an accurate, convenient, and simplified method of measurement. This paper describes an expansion of the conventional oscilloscopic method. Technique for measurement of video amplifiers, picture i-f amplifiers, delay networks, and artificial lines is presented along with a description of the equipment used and a discussion of the requirements of such equipment. Also included is a short discussion of the probable magnitudes encountered in such measurements.*

THE ideal television amplifier would be one having uniform gain and linear phase shift throughout the video pass-band. The uniform gain requirement is apparent and familiar, since it is an audio amplifier requirement. In television amplifiers, however, linear phase shift is equally as important as uniform gain. Hence, equipment and technique for phase shift measurement are of prime importance in television development, design, and test work.

The phrase "linear phase shift" is synonomous with "constant time delay". This becomes apparent when it is recalled that $t = \frac{\theta}{\omega}$ or $t = \frac{\theta}{2\pi f}$. Thus, if phase shift is proportional to frequency, the time delay is constant, and all components of a complex wave have equal time of transmission. Perhaps the most straightforward method of phase shift determination is by direct observation on the cathode-ray oscilloscope. The general arrangement is shown in Figure 1.

At some frequency in the band of 5 kc to 50 kc the phase shift in most amplifiers is close to zero. Under this condition the trace on the oscilloscope is a straight line, its slope or orientation depending upon the relative amplitudes of the vertical and horizontal deflections. It is advantageous to make the sweeps approximately equal so that the trace will be symmetrical. As the frequency is increased from the zero phase-shift value, the trace on the oscilloscope passes through the configurations shown in Figure 2.

The frequencies at which the "cardinal points" (0, 90°, 180°, 270°,

kc to 50 kc at which the trace appears as a straight line. This uncertainty may be eliminated by assuming one of the possible points to be correct, and then plotting the results as shown in Figure 3. If the correct point has been chosen, the curve, if extrapolated, will pass approximately through the origin. If one of the incorrect points has been chosen, the curve, when extended, will pass through the ordinate at plus or minus 180° approximately. This is shown by the dotted lines in Figure 3. The situation will be understood when it is recalled that the phase shift in the upper video range (above 5 kc or so) is caused by capacitances shunting the load resistors in the amplifiers. This type of phase shift must be zero at zero frequency. The net phase shift is only approximately zero because of the effect of grid-coupling condensers at the low-frequency end of the video range. In summary, the

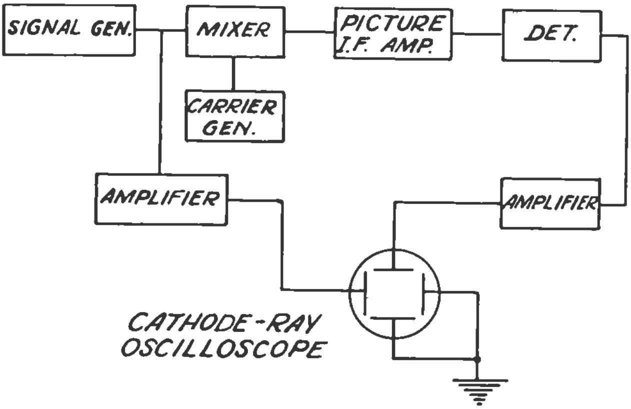

Fig. 4

zero phase-shift point is characterized by a straight-line trace on the oscilloscope, and by a phase-shift curve which passes approximately through the origin. With a little experience the matter of location is relatively simple.

Note that in Figure 3 negative phase shift has been plotted in the positive direction, for convenience. It must also be remembered that delay time is independent of the direction of phase shift; i.e., both phase advance and phase delay result in time delay. It is physically impossible to cause a time advance. This last statement may be the source of considerable confusion. However, it may be more readily understood when it is realized that we are interested in delay time from a transient standpoint. It is true that in a steady-state condition there may be an apparent time advance.

For example, consider the case of a single, parallel-tuned, resonant circuit driven by a high-impedance vacuum tube. If a carrier and two side bands are impressed upon the grid of the tube and the carrier is at the resonant frequency, it will not be shifted in phase except for the 180° displacement caused by the tube's grid-plate characteristic. However, the lower side band will be advanced in phase, and the upper side band will be retarded by approximately the same amount. But, all three components are subjected to a very definite time delay (due to electrical inertia in the resonant circuit) which is exactly equal to the phase rotation (in radians) divided by the angular velocity of the side bands with respect to the carrier. More specifically, suppose the modulating frequency to be 100,000 cycles per second and the side bands shifted $\frac{2\pi}{8}$ radians (45°) with respect to the carrier. The entire signal would

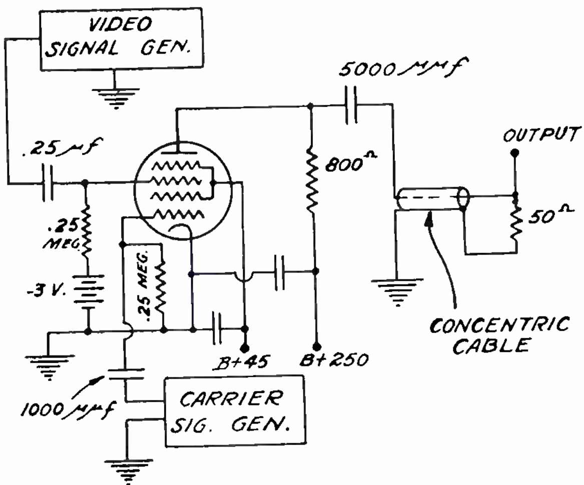

Fig. 5

then be delayed by $\dfrac{2\pi}{8 \times 100{,}000 \times 2\pi}$ or 1.25 μsec.

Picture I-F Phase Shift

When measuring phase shift in picture i-f amplifiers, it is necessary to use a carrier modulated at video frequency. The picture i-f output is then passed through a detector and all measurements made at video frequency as before (see Figure 4).

A mixer suitable for supplying the modulated carrier for picture i-f measurement is shown in Figure 5. A type 6L7 tube is used.

The problem of picture i-f phase shift in a simple parallel-resonant circuit when the carrier is set at resonance has already been discussed, although briefly. By reference to Figure 6, it can be seen that under this condition the maximum shift may be 90° per stage. However, if the carrier is set on one side of the resonance curve, the maximum phase shift may approach 180° per stage. Such would be the case when single side-band reception is attempted.

In some instances it may be desirable to use an i-f system consisting of stages alternately coupled through simple parallel-resonant circuits and overcoupled double-tuned circuits. The object of this is to have the single resonant peak compensate for the dip in the center of the resonance curve of the double-tuned stage. This usually results in a higher uniform overall gain for a given number of stages.

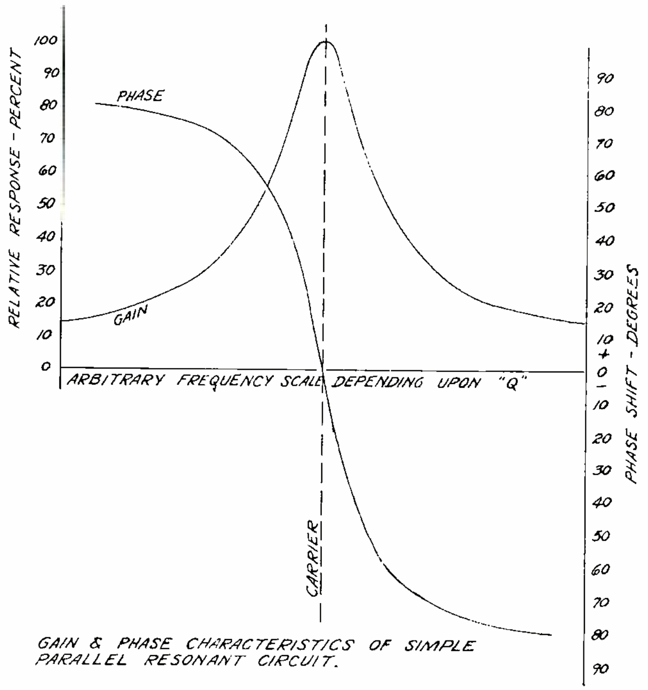

Fig. 6

The characteristics of a representative double-tuned circuit are shown in Figure 7. Again, as in the case of the simple parallel-resonant circuit, the maximum possible phase shift is doubled when the carrier is set to one side of the resonance curve for single side-band reception. It may be seen that when the carrier is symmetrically placed, the maximum phase shift approaches 180° as a limit. With the carrier at one side the maximum phase shift approaches 360° as a limit.

It is interesting to note that when the system just mentioned (alternate single- and double-tuned stages) is used for single side-band reception, the frequencies in the upper video range are essentially single side band in nature, while the frequencies in the lower range are dealt with as double side band. This may be understood more clearly by reference to Figure 8. Here, the carrier is placed approximately at the point of 50 per cent response on the high-frequency end of the pass band.

The complexity of calculation involved in predicting the phase characteristic of such a composite system may readily be appreciated.

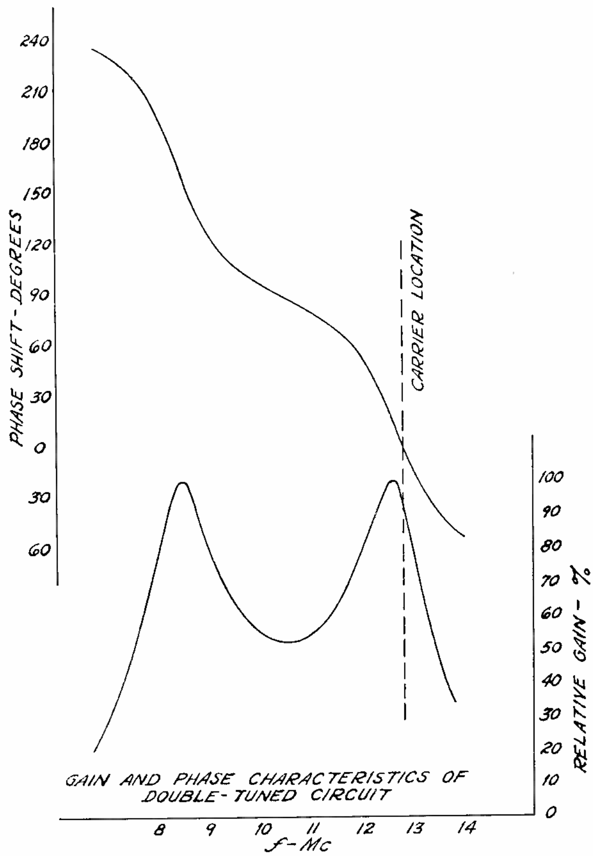

Fig. 7

Thus arises the utility of this quantitative method of phase-shift determination.

Small Phase Shifts

In the cited examples it has been assumed that the amplifier under test had a sufficient number of stages so that the total phase shift was large enough to give numerous cardinal points. When investigating amplifiers having only one or two stages, it is often desirable to increase the number of points on the curve above that obtainable by the cardinal point method. There are two other possibilities in such cases.

The first method is as follows: Adjust the gains of the horizontal and vertical amplifiers to give a good-sized trace. Turn off or otherwise remove the horizontal sweep. Place a scale beside the remaining vertical deflection and measure its length. Next, restore the horizontal sweep and then measure the distance between the intercepts of the trace at the scale. The ratio of these two distances is the sine of the phase angle. While this method permits measurement of phase angles during the transition between cardinal points, it will be found subject

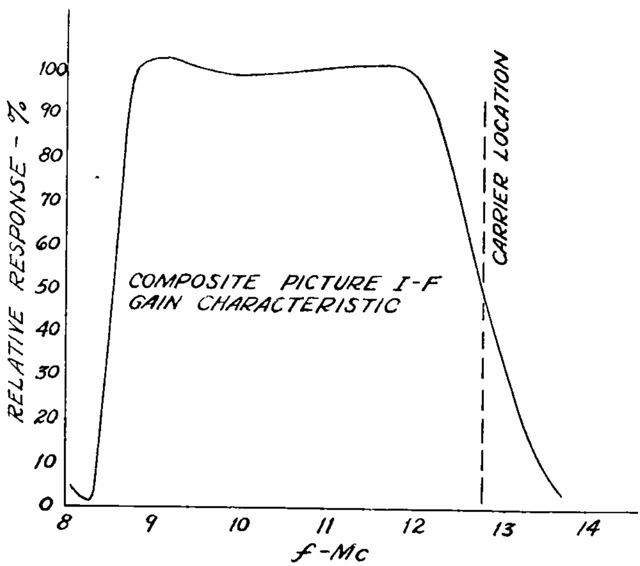

Fig. 8

to appreciable error due to the difficulty of measuring the distances accurately, particularly if the viewing screen of the oscilloscope is small.

A more satisfactory method is to place a relatively large known delay in series with the amplifier under test. Then, after determining the total delay, the known delay is subtracted and that attributable to the amplifier under test results. By application of transmission-line theory it may be shown that, in loss-free lines having uniformly distributed constants, the following relationships are true.

$$t = n\sqrt{LC}$$

$$Z_s = \sqrt{\frac{L}{C}}$$

where

$t =$ overall time of transmission (seconds)
$n =$ number of sections
$L =$ inductance/section (henries)
$C =$ capacitance/section (farads)
$Z_s =$ characteristic impedances (ohms)

Thus, the delay may be obtained by using a section of transmission line, or by using an artificial line or simulating network. While the transmission line will prove satisfactory from an electrical standpoint, it may be found quite bulky and difficult to handle. For example, one type of transmission line which has considerable use in television work has the following constants.

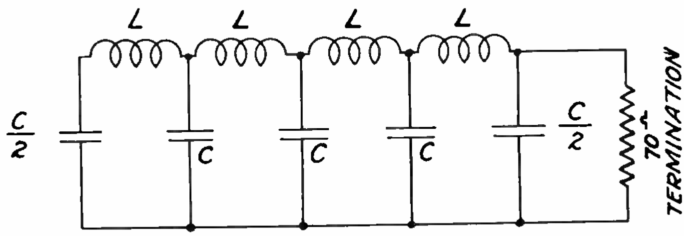

Fig. 9

$$L = 121\ \mu h/\text{ft}$$
$$C = 26.1\ \mu\mu f/\text{ft}$$
$$Z_s = 68\ \text{ohms}$$
$$\text{delay} = 0.00178\ \mu\text{sec/ft}$$
$$(562\ \text{ft for } 1\ \mu\text{sec delay})$$

This cable consists of central conductor, rubber sheath, and flexible outer shield.

The delay may be more conveniently obtained from an artificial line or simulating network. The equations given for transmission lines may be used for the design of the simulating network. The data for one type of lumped-constant delay network which proved satisfactory are given below. The circuit is shown in Figure 9.

$$L = 7.5\ \mu h/\text{section}$$
$$C = 1500\ \mu\mu f/\text{section}$$
$$Z_s = 70\ \text{ohms}$$
$$f_c = 3\ \text{Mc} = \frac{1}{\pi\sqrt{LC}}$$
$$t = 0.106\ \mu\text{sec/section}$$

In this particular network twenty sections were used, giving a total delay of about 2.2 μsec. While the delay given by these networks is approximately constant with frequency, it is desirable in many cases to have a point-to-point correction curve available. These data may be obtained by measuring the delay of the network just as if it were an amplifier under test. A typical curve is shown in Figure 10.

SIGNAL GENERATOR

The signal generator for the upper frequency range should preferably be of the beat-frequency oscillator type. Desired characteristics are—wide frequency range, direct-reading scale, relatively constant output throughout the frequency range, low output impedance, high stability, and *accurate calibration.* It is important that the frequency calibration be accurate because the ultimate accuracy of this method of phase shift measurement depends primarily upon the exactness of the frequency measurement. This cannot be stressed too strongly.

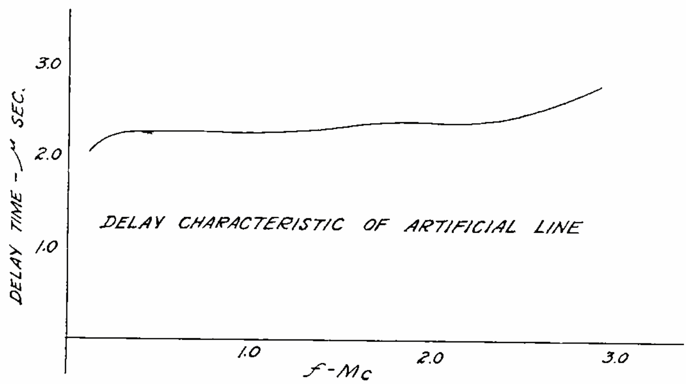

Fig. 10

SWEEP AMPLIFIERS FOR THE CATHODE-RAY OSCILLOSCOPE

In general, the amplifiers for the cathode-ray oscilloscope present no particular difficulties. The important factor in the design and construction of these amplifiers is that they must have *identical* phase characteristics. It is not necessary that their phase characteristics be linear, or that they be known. The frequency characteristic need not be absolutely flat, but it is convenient to have it approximately so in order to eliminate the necessity for changing the gain as the frequency is increased. It will be found advantageous to expend the additional effort necessary to make the frequency characteristic quite flat and the phase characteristic reasonably linear. This will be found true not because these features are necessary for phase measurement, but because they are highly desirable if the amplifiers are to be used for the many other measurements which arise in television work. The amplifiers may be checked for identical phase characteristics by applying a common signal to the inputs and observing the difference in phase shifts on the cathode-ray oscilloscope. Care should be exercised in arranging the input and output coupling circuits of these amplifiers in order to retain as much symmetry as possible.

Particular care should be taken in the arrangement of the attenuator or gain-control circuits of the amplifiers. It is not possible to use the customary high-resistance potentiometer for grid-circuit gain control as is done in audio amplifiers because of the shunting effect of the tube input capacitance at high frequencies. A choice of two alternatives is possible in such cases. Either the grid-to-cathode portion of the potentiometer may be made to have an impedance small com-

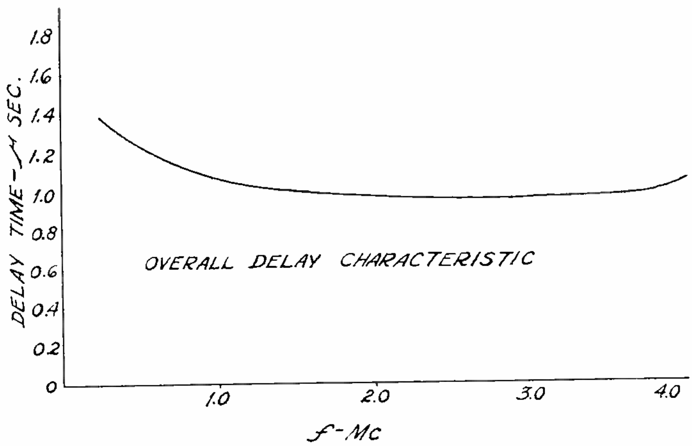

Fig. 11

pared to the input reactance of the tube, or the upper portion of the potentiometer may be by-passed with capacitance in order to balance the voltage division at the higher frequencies. Attenuators of the latter type are inherently of the step type. An additional measure of gain control may be afforded by biasing back the control grid of a remote cut-off tube. This should preferably be done in the first stage of the amplifier, where the impedance of the signal voltage source is low, so that changes in tube input capacitance with bias voltage will not cause changes in phase shift.

Probable Magnitude and Variation in Delay Time

In an average television receiver the total delay, or time of transmission, from first detector to Kinescope grid will be of the order of 1 to 1.5 μsec, as shown in Figure 11. In a 441-line picture the horizontal repetition rate is 13,230 per second. If 15 per cent horizontal return time is assumed, the scanning period of one line of the picture is 64.1 μsec. Again, assuming a 4:3 aspect ratio, 7 per cent vertical return time, and equal vertical and horizontal resolution, it is found that the scanning period of one horizontal picture element is

$$64.1 \times \frac{1}{441\,(1-0.07) \times \frac{4}{3}} = 0.117\ \mu\text{sec.}$$

Thus it may be seen that a variation in time delay in the upper frequency range of about 0.1 μsec corresponds to a horizontal displacement of one picture element. It is probable that displacements of less than this amount would be unnoticeable. The problem is further complicated by the fact that the resolution may in some cases be limited by factors other than phase shift. It is entirely possible that in some cases such causes as insufficient band width or large Kinescope spot size may mask the deleterious effects of phase shift.

However, the situation regarding time delay at low frequencies, in the vicinity of 100 cycles, is quite different. The wave components in this frequency region have little effect upon the detail of the reproduced picture, but merely supply the background or shading. In fact, even the line frequency has but little effect upon detail. Hence, it is possible to tolerate an even greater time delay at these low frequencies without its effect being too apparent in the picture. For example, a delay of 75 μsec at 60 cycles merely causes the general vertical shading to be shifted downward two lines. Since the background shading is so gradual its displacement would not be apparent in the picture.

MEASUREMENT OF BROADCAST COVERAGE AND ANTENNA PERFORMANCE

BY

WILLIAM A. FITCH AND WILLIAM S. DUTTERA
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Summary—The first two parts of this paper were presented in earlier issues of the RCA REVIEW*. *Part I covered the general aspects of "Day-time and Night-time" coverage. Part II described the method of determining the input power to and the efficiency of an antenna. This third part paper, Part III, is a continuation of the subject matter contained in Part II and includes methods of determining the one-mile signal intensity and a description of the steps necessary in selecting a new transmitter site.*

PART III

DETERMINATION OF ONE-MILE SIGNAL INTENSITY

ACTUALLY the ideal field-strength measuring conditions enumerated in Part II cannot be expected and the accurate determination of E_1 is considerably more involved than the making of a single measurement of field strength. The determination of E_1 will now be described for the general case.

As outlined in Part II the errors in measuring signal intensity near the transmitter are: (a) Errors in determining the exact distance of the measuring point from the antenna; (b) the possibility that the value of signal intensity at that particular measuring point is not representative, but is influenced by uneven terrain, proximity of overhead wires and in some cases undeground conductors; (c) inherent inaccuracy of the field-intensity meter plus the inaccuracy caused by the observer in the operation of the equipment; and (d) inaccuracies arising from an insufficient number of measurements about the antenna to show all the irregularities in the signal intensity that may occur in the different directions, and result in a non-circular pattern.

From the possible sources of error, enumerated above, it is possible to outline a measuring procedure which will minimize most of these errors. It does not seem that an abbreviation of the following procedure is ever justified for the first set of measurements, although for subsequent measurements it is generally permissible to reduce considerably the work by the selection of key measuring points which previously have been thoroughly evaluated.

* W. A. Fitch and W. S. Duttera, "Measurement of Broadcast Coverage and Antenna Performance" Part I, RCA REVIEW, Vol. II, No. 4, April, 1938 and Part II, RCA REVIEW, Vol. III, No. 3, January, 1939.

The measurements of field intensity should be made on a radial extending from about 0.6 mile to about 7 miles from the antenna. Measurements should be made very frequently on this radial. Practically, it may be impossible to obtain good measuring points spaced as close as desirable. About 6 nearly equally spaced measurements should be made between say 0.8 mile and 1.5 miles. About 6 nearly equally spaced measurements should be made between 1.5 miles and 3

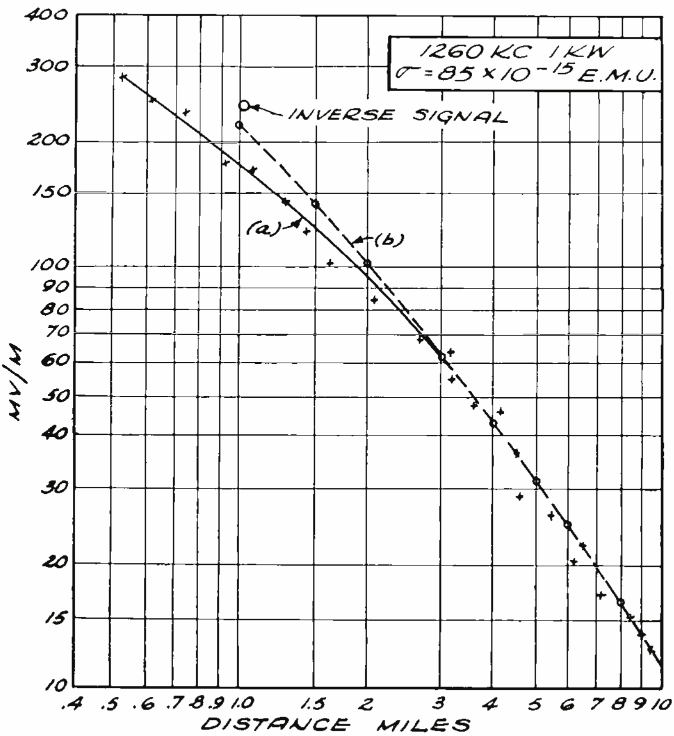

Fig. 18—Radial for determination of one-mile signal.

miles. About 6 measurements should be made between 3 and 7 miles. It may be desirable in some cases to continue the measurements out to 10 or 15 miles. The figure of 7 miles and the number of measurements are almost ideal figures and may be subject to some decrease as the individual case would seem to permit. These radials may be plotted as shown in Figure 18.

These radials should be made at least every 60 degrees and much more frequently if there is any possibility of a non-circular pattern or if the six radials mentioned above are not alike.

Quite often the one-mile signal intensity has been found by making a series of measurements around an antenna at distances of about

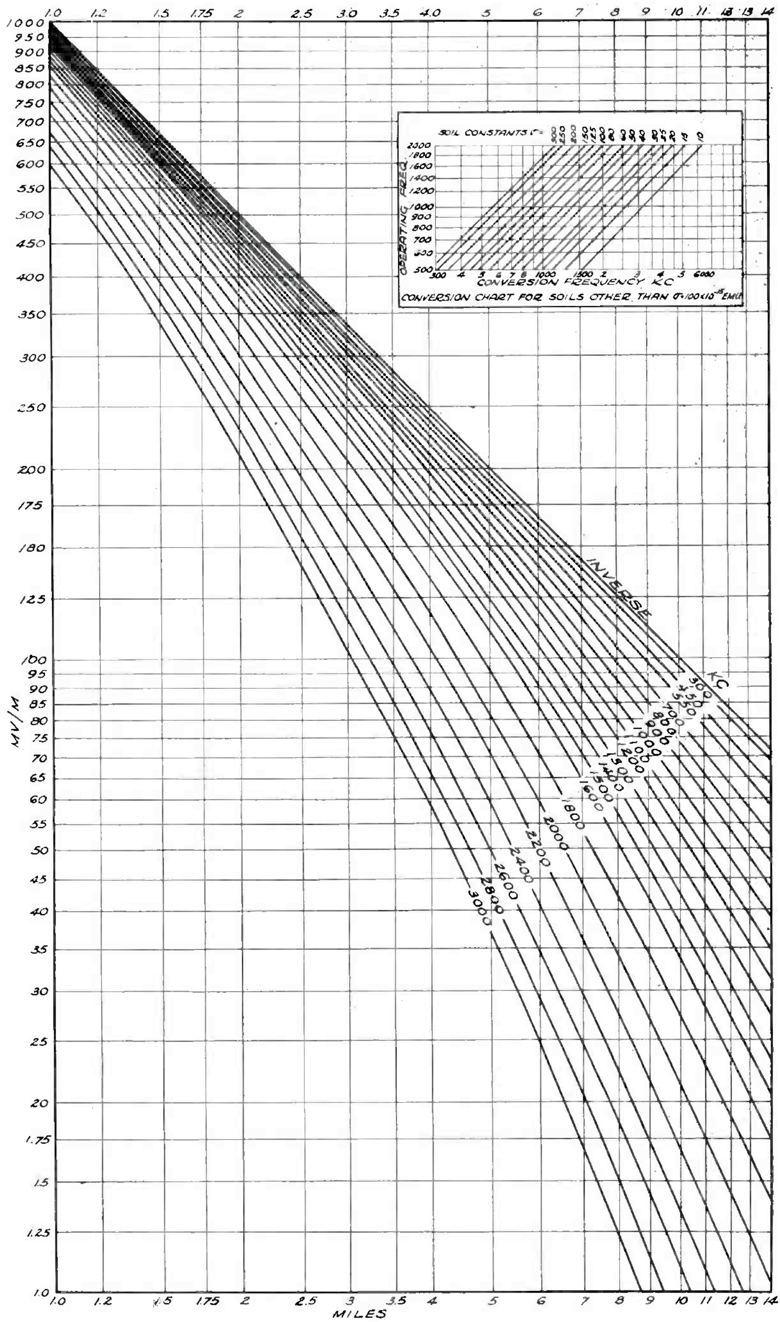
SOIL CONSTANTS σ= 300 250 200 150 125 100 80 60 50 40 30 25 20 15 10
OPERATING FREQ.
CONVERSION FREQUENCY KC
CONVERSION CHART FOR SOILS OTHER THAN σ=100×10⁻¹⁵ EMU
INVERSE
KC
MV/M
MILES

Fig. 19

one mile and corrected for one mile by multiplying the measured signal by the distance from the antenna. This method has been found to give only approximate results when about 80 measurements were made and in several isolated cases the results have been off as much as 10 to 15 per cent, which is a far greater loss of accuracy than can be tolerated or justified.

Needless to say only large-scale maps should be used in making the distance determinations. Because of the inaccuracy of the map itself and the difficulty of exactly locating the measuring point on the map, the measurements made at less than about one mile have to be regarded with suspicion. The map should preferably have a scale no smaller than one mile to the inch. Landmarks on most maps should be carefully checked for inaccuracies, as inaccurate maps are often found.

Since the antenna is not responsible for the attenuation introduced by the soil beyond the antenna system, the measured radials should be corrected for this attenuation so that the field intensity used in evaluating the antenna performance is the inverse field or unattenuated intensity at one mile rather than the measured field intensity. The determination of the inverse one-mile signal requires an accurate knowledge of the attenuation constant of the soil over which the radial has been taken.

This constant may be determined by comparing the measured attenuation with the curves on Figure 19 over the distances of about 3 to 10 miles. From the frequency of the curve which fits the measured curve, and from the operating frequency the soil constant may be found on the conversion chart of Figure 19. After the appropriate curve of Figure 19 has been selected the one-mile attenuation is easily found.

As an example of the above procedure see Figure 18. Crosses represent the measurements and the curve (*a*) is drawn through these measured values. When this curve is fitted on the curves of Figure 19 it is found that the best fit is on the 1350-kc curve between 3 miles and 10 miles. This then means the soil constant is 85×10^{-15} EMU. For the sake of simplification the points off the 1350-kc curve have been replotted as circles and curve (*b*) has been drawn in Figure 18 with the inverse one-mile signal shown at 240 mv/m.

The last few paragraphs have described the method of obtaining the maximum accuracy in determining the one-mile signal along a radial. As previously outlined it is necessary to measure these radials in quite a few directions. Each of these radials is then evaluated as just described. These one-mile inverse-signal intensities should then be plotted on polar graph paper as shown on Figure 20, curve (*A*). If this curve is substantially circular an average value may be easily determined. However, if it is like the curve (*A*), the rms value should be determined. This may be done by squaring and adding the signal intensity every 10 degrees. This sum is then divided by 36 and the square root of this ratio gives the rms value of signal intensity. This

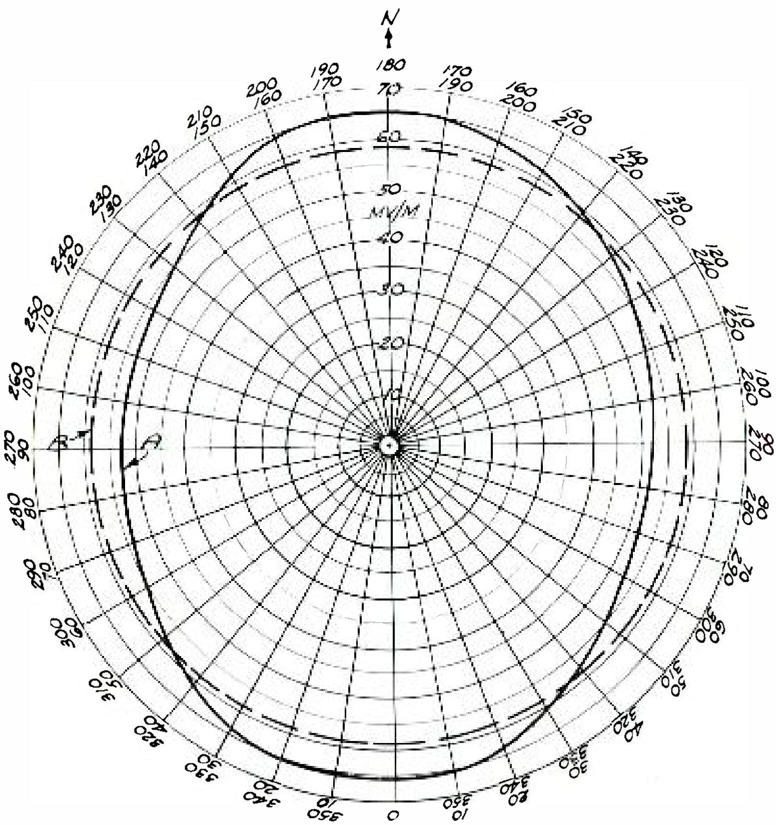

Fig. 20—One-mile signal-intensity pattern.

is shown in Figure 20 as curve (*B*). The efficiency may then be found as described in Part II by using the rms value.

If the ground pattern is not circular it may be because of one or more of the following reasons: (a) Irregularities in the ground system of sufficient magnitude to distort the field of the antenna; (b) a building in the immediate vicinity of an antenna, or (c) antenna supporting towers or other structures in the immediate vicinity of the antenna carrying appreciable induced r-f current. Condition (a) is practically limited to antennas on buildings where the ground system is far from being symmetrical. This is found, for instance, where the antenna is located on a corner of the building and ground system. This form of pattern distortion is seldom found to an appreciable extent when the antenna is over ordinary ground. Condition (b) may be found when a fairly large structure is near an antenna and may result in a non-circular pattern due to distortion of the field of the antenna or re-radiation if the building is of sufficient height to have high induced currents. Condition (c) is perhaps the most prevalent cause of a non-circular pattern. This may result effectively in two forms of losses, the first of which is due to the fact that the re-radiators are not generally found to have low loss, and secondly, due to the fact that better cancellation may exist on the ground than at higher angles[24]. This latter point is of minor importance if the pattern distortion is not serious.

Accuracy of Measuring Input and Radiated Power from an Antenna

In measuring the input power to an antenna the accuracy is dependent upon two factors as previously mentioned, namely, the accuracy of the resistance measurement and the accuracy of the current measurement. The accuracy of the resistance measurement is of course dependent upon the accuracy of the frequency of the test voltage.

If the frequency is assumed to be known to within ± 0.1 per cent the maximum errors that could occur under proper measuring conditions as outlined above for either the bridge or substitution methods, would be,

(1) Error in standard resistance	± 1	per cent
(2) Error due to frequency deviation	± 1	per cent
(3) Error in determining I^2	± 2	per cent
Total error in input power determination	± 4.04	per cent

This overall error, of course, does not allow for any observational errors. This assumes half-scale reading on a meter having 0.5 of one per cent accuracy on full-scale reading. It is true that the first two errors may be such that they cancel each other or even add and cancel the third error. However, with equipment ordinarily available for field use, the direction of these errors cannot be determined and the measurements should be considered subject to that maximum error. In the average case some cancellation of these errors would compensate for operational error. It may thus be said that the input power into

[24] W. S. Duttera, "Some Factors in the Design of Directive Broadcast Antenna Systems," RCA Review, July, 1937.

an antenna may be determined to within four or five per cent. This assumes of course that there is not appreciable coupling with an overhead transmission feed line which cannot be properly evaluated in the resistance measurement.

The accuracy of output power measurements, when made in accordance with the detailed procedure previously set forth are as follows:

(1) Error in the absolute measurement of field intensity	± 5 per cent
(2) Error in distance determination	± 2 per cent
(3) Error in determining inverse-field intensity	± 2 per cent
Maximum error in determining 1-mile signal	± 9.2 per cent
Maximum error in determining radiated power	± 19.3 per cent

Here again it is not usually possible to determine the direction of the individual errors and no allowance has been made for observational errors. Of course by extraordinary effort the second error may be reduced. It would appear that even under optimum conditions, the error in determining radiated power is subject to inaccuracies in the order of 20 per cent. This possible inaccuracy is quite high, but when it is considered that these are largely field measurements and not laboratory measurements, the results seem more reasonable.

Selection of Transmitter Site

It is sometimes necessary to select a transmitter site for one or more of the following reasons: (a) A new station may desire to obtain a transmitter site; (b) a station may desire to provide better service to the majority of its listeners by moving to a more suitable location; (c) a station may desire to install a directional antenna to limit radiation in certain directions and may find it necessary to move to accomplish this without reducing its local coverage; or (d) a station may desire to increase power and it may be necessary to relocate in order that excessive blanketing does not occur.

If for any of the above reasons or for any other reason it is desired to relocate a transmitter it is essential that the following factors be considered in selecting the site. These factors are roughly listed in the order of their importance. It should be noted that the evaluation of these factors is often quite complex and deserves the consideration of a consultant who is intimately familiar with all the aspects of this problem. The importance of the best selection of a site is self-evident when the magnitude of the investment is considered together with the possible cost and inconvenience of moving to a more suitable site at some future time.

(1) The site must be so situated that blanketing of an excessive percentage of receivers will not be produced. The blanketing signal may be reasonably considered to be 250 to 500 mv/m.

(2) The site should be so located with respect to the service area that a suitable directive antenna will furnish the desired protection, if this protection is necessary, without undue limitation of the service area.

(3) The site should provide a maximum signal for the greatest possible number of listeners. The city immediately adjacent to the station should receive not less than approximately 50 millivolts per meter to insure good service under adverse noise and atmospheric disturbances.

(4) The site should preferably be near established power and telephone facilities.

(5) The site should be located on terrain that permits good efficiency from the antenna system. This means, under ideal conditions, level marshy land. This condition is often hard to obtain and compromises are often necessary at the expense of somewhat lower one-mile signal intensity.

(6) The site should be near a main highway in order that it be accessible under adverse weather conditions.

(7) There should be available, at a reasonable cost, a sufficient area of land to accommodate the antenna and ground system.

(8) The antenna system should not be an undue hazard to established airlines.

After consideration and investigation of the above factors it is generally possible to locate several possible sites. In order to determine more exactly the expected performance from these sites it is generally necessary to set up a test transmitter and antenna. This should be done first at the most promising site. If this site is found by test to be below expectations the test should be made at the next desirable site and so on. The procedure of making a site test will be considered next.

Test Transmitter and Antenna

A site test first necessitates the setting up of a small transmitter and a small antenna to provide the test signal. The signal intensity from this transmitter is then measured over that area wherein the value of signal intensity from the proposed station is accurately desired. The site survey may be divided into three major parts. The first comprises the test transmitter, antenna ground system, and installation. The second is the measurement of field intensity from this transmitter and the third is the prediction of actual performance of the station from the test measurements.

The test transmitter may be a conventional low-power portable or semi-portable transmitter. This transmitter ordinarily should be

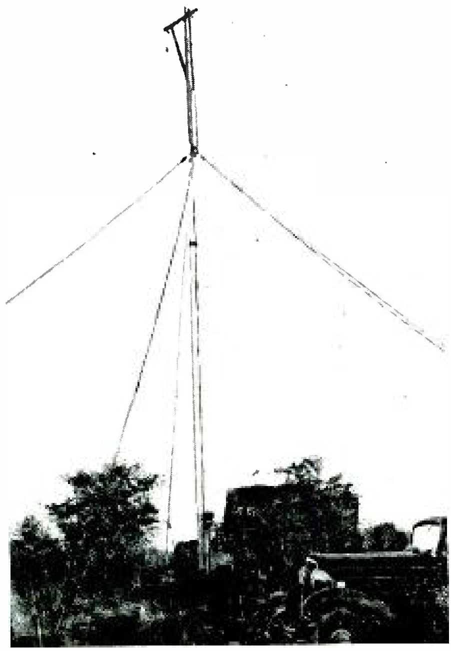

Fig. 21—A site test transmitter and antenna.

capable of delivering about 100 watts into the antenna with good frequency stability and little fluctuation of output power. The amount of power necessary is dependent upon the efficiency of the test antenna and the distance over which it is desired to make measurements. If measurements are to be made when there is high atmospheric interference an allowance in transmitter power must be made for this. That is also the case if the measurements are to be carried out to the distance at which the interference is comparable to or greater than the measured signal. Since these measurements are generally made at night, it is often found that they must be limited to about 30 or 40 miles due to fading. This limitation due to fading is, of course, most severe for the higher frequencies and the lower conductivity soils.

The transmitter should be capable of operation from the existing power supply at the proposed site. If there is no power supply it must be furnished from a gas-driven generator or some other power source. The transmitter should be capable of feeding an antenna of low resistance and high capacitive reactance. Since such a load circuit has a high impedance-to-resistance ratio, the transmitter, if a self-excited oscillator, must be designed with this fact in mind. The transmitter should have a suitable frequency monitor. For a survey, it is sufficient to use a wave meter of the most advanced type. In order to check the constancy of output power a suitable antenna meter should be provided. This meter reading should be recorded approximately every half hour and at all intermediate periods when there is a change in the antenna current, noting the time and the new antenna current value.

The antenna may be any conveniently erected vertical antenna. In order to facilitate loading and to reduce loading losses, it should be preferably a cage type. It is generally inconvenient to provide an antenna more than about fifty feet high, but a single wood pole of this height is generally available. The antenna may be suspended from an arm as shown in Figure 21. The transmitter in this picture is in a box at the left of the antenna.

The ground system, found most convenient, is assembled from six-foot galvanized No. 18 gauge wire fencing. This fencing is cut in three pieces each about 75 feet long. These pieces are placed on the surface of the ground. They form equal angles with each other and intersect at the base of the antenna. This then forms a ground system having a radius of 37.5 feet. If it is necessary, a more elaborate and efficient ground system may be employed.[25] These pieces are then securely soldered at the intersection and a connection is brought into the transmitter.

Field-Intensity Measurements from the Site Transmitter

The next step is the measurement of the signal intensity from this test transmitter. These measurements should be made in general accordance with the procedure outlined in the first section of this paper[21]. For these measurements it is very important that the one-mile field intensity be accurately ascertained as outlined above. Thus with accurate one-mile measurements and with measurements along

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[21] loc. cit.

[25] G. H. Brown, R. F. Lewis and J. Epstein, "Ground Systems As A Factor In Antenna Efficiency," *Proc. I.R.E.*, June, 1937.

radials in all important directions, it is possible to predict the actual field intensity obtainable from the proposed antenna system with a relatively high degree of accuracy.

It might be pointed out here that if the survey measurements are

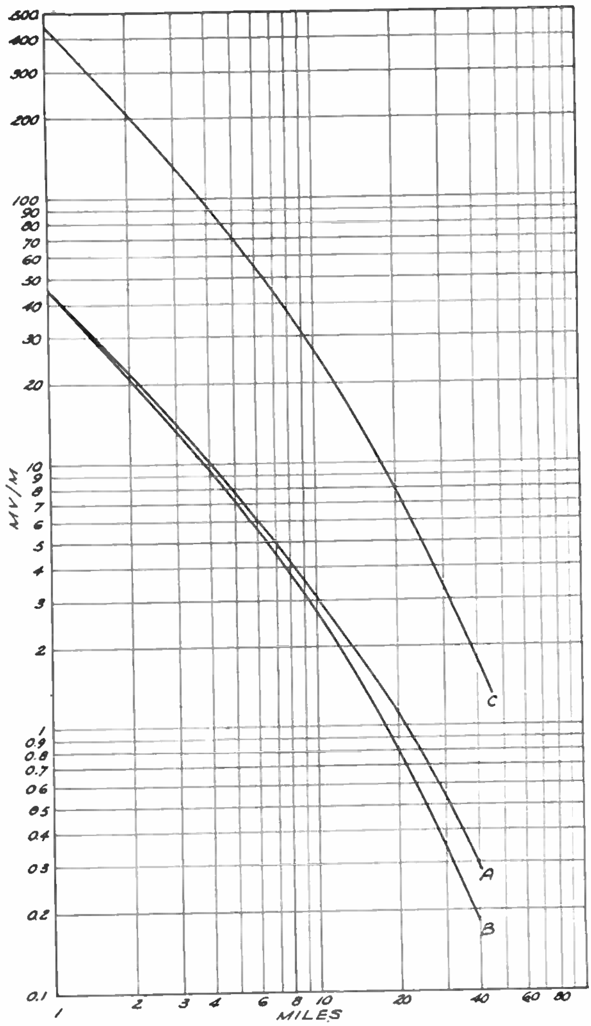

Fig. 22—Sample radial from a test transmitter.

made on the frequency of an existing transmitter, it is necessary to make these measurements after the sign-off of the regular transmitter. This of course means night measurements. It is sometimes possible to make measurements on a frequency 20 kc or so separated from the operating frequency. It is quite desirable to do this if it thereby permits day measurements. There are several good reasons for such a procedure: (1) Much better measuring locations may be chosen, free from power wires, etc., in the daytime and thus the average accuracy of most of the measurements is considerably improved; (2) a more accurate survey may be made in much less time; (3) the survey may be extended to a greater distance due to the absence of the fading limitation; (4) if there happens to be an open season on cattle rustlers or chicken thieves in that particular locality the engineer will complete the job with considerably more ease of mind and probably without the danger of collecting some lead pellets as souvenirs.

The measurements when made on a frequency other than the operating frequency must be properly corrected for the operating frequency. This procedure will be more fully described later.

Signal-Intensity Measurements from the Site Transmitters

These measurements may be divided into two groups. The first group of measurements is used to determine accurately the signal intensity at one mile. They should be made as the measurements described above to obtain the one-mile intensity. The second group of measurements should be made in radials as described in Part I[21]. These measurements should be carried out along the radial as far as it is desired to predict accurately the performance of the proposed antenna system. The distance to which these radials may be carried is limited, of course, by the power of the test transmitter, fading, etc., as mentioned above.

The radials may be plotted as shown in Figure 22, curve (*A*). The frequency in this case is, say, 1000 kc. This conductivity may be found by determining that curve of Figure 5 (Part I) which has the same slope at every distance between three and forty miles. This radial then is found to be over soil having a conductivity of 100. From Figure 19 the inverse-signal intensity is found to be 50 mv/m at one mile.

To find a corresponding radial over the same path for an operating frequency of say 1200 kc it is only necessary to adjust the ordinate for an inverse signal of 50 mv/m at one mile and read the signal intensity from the 1200-kc curve for the various distances. This is shown as curve (*B*) in Figure 22.

These radials should be made in quite a few directions from the transmitter as outlined above. The inverse-signal intensity at one mile is then assumed for the proposed antenna system and the ordinates of Figure 22 are multiplied by the ratio of this signal to the 50 mv/m

[21] loc. cit.

found above. If it is assumed that the proposed antenna operating on 1200 kc will produce a circular pattern having a one-mile inverse-signal intensity of 500 mv/m, the ordinates of curve (*B*) of Figure 22 are multiplied by ten. The predicted radial is then shown as curve (*C*). When these adjusted radials, in the various directions, are plotted the contours may be drawn and the coverage from the proposed site may be accurately determined.

In conclusion it should be pointed out that many special situations may arise, the solution of which often requires special attention. It is difficult to collect in a paper of limited length all of the various problems that may be encountered. It is likewise important to remember that certain of the empirical standards mentioned will be subject to change as the art develops.

ERRATA

RCA REVIEW, Vol. III, No. 3, January, 1939.

MEASUREMENT OF BROADCAST COVERAGE AND ANTENNA PERFORMANCE, PART II.

By W. A. Fitch and W. S. Duttera

Correction of equations 11 and 12:

$$Z_o = 120 \cosh^{-1} \left(\frac{S}{2r} \right) = 276 \log_{10} \left[\frac{S}{2r} + \sqrt{\left(\frac{S}{2r} \right)^2 - 1} \right] \quad (11)$$

$$Z_o = 69 \left[\log_{10} \left(\frac{4h^2}{1.41\, Rd} \right) - 4 \frac{\log^2_{10} \left(\frac{2h}{d} \right)}{\log_{10} \left(\frac{4h^2}{1.41\, Rd} \right)} \right] \quad (12)$$

WIDE-BAND VARIABLE-FREQUENCY TESTING TRANSMITTERS

By

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Summary—*A wide-band variable-frequency transmitter, designed primarily for testing purposes is described. The transmitter consists of a single-tube oscillator stage with a high degree of frequency control. Two models have been built which operate in the frequency range of 40 to 100 megacycles, with 500 to 1000 watts output. Means are also provided for telegraphically keying the transmitters. This type of transmitter supplies a need for a source of high-frequency power which is continuously variable over a wide range of frequency, but which will accurately maintain any frequency to which it is adjusted.*

INTRODUCTION

DURING the early part of the year 1937 the need was felt for more information on the propagation characteristics and multipath phenomena of radiated electric waves over wide-frequency bands in order that we might attack more intelligently the problem of television transmission. As a part of the program to supply this information the Transmitter Research and Development Laboratory at Rocky Point, Long Island constructed and installed in the top of the Empire State Building one each of two new types of variable frequency transmitters including their antennas. These transmitters were used for survey tests in collaboration with the Receiver Research and Development Laboratory at Riverhead, Long Island.[1] The transmitters have rather accurate frequency control and are designed to operate at approximately 150 and 90 megacycles with a power output of 500 to 1000 watts. Comparatively large power output was required in order to simplify and to increase the reliability and accuracy of the field measurements.

It might also be stated that in the past the testing of any apparatus such as antennas, to obtain characteristic data at different frequencies, required the construction of a number of models, or the adjustment of a more or less fixed-frequency test transmitter in several steps of frequency. These methods are both slow and expensive. Consequently, for a long time we have felt the need of means to reduce the time and expense necessary for taking these measurements by making available sources of high-frequency power which are continuously variable over a wide range of frequency, but which will accurately maintain any frequency to which they are adjusted. This need has been supplied to a great extent in our case by one of these new types of variable-frequency transmitters.

Description

Because of its simplicity of design, accuracy of frequency control, and because it fills a general need for testing purposes it is thought that a description of the 90-Mc survey transmitter would be of general interest.

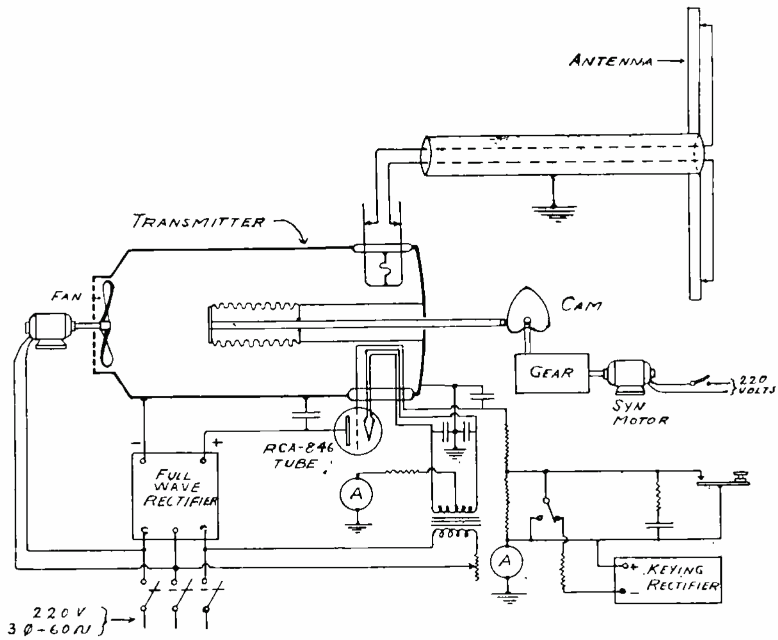

Fig. 1.—Schematic diagram of the variable frequency transmitter.

The transmitter consists of a single-tube oscillator stage with a high degree of frequency control. The principle by which this type of transmitter operates may best be understood by referring to Figure 1. The transmitter is a "grounded anode" type of oscillator in which the radio-frequency voltage of the oscillator tube grid is in phase with that of the filament, but the grid is driven to a greater amplitude. The oscillator driving power is applied through the tube filament. The only oscillating tank, or fly-wheel, circuit in the transmitter consists of a quarter-wave section of two concentric tube conductors which act as a frequency-control line. This line is constructed of copper tubing. One end of the two concentric conductors is shorted together and supported by a heavy copper disk. The grid and filament of the transmitter tube are coupled to the high-current, or short circuited, end of the frequency-control line through loop conductors which are located near the inner concentric conductor. The outer ends of these loops are grounded for radio-frequency currents to the outer concentric conductor through by-pass condensers before being connected to their respective power sources. In general the length of the frequency-control line determines the operating frequency. The outer line conductor

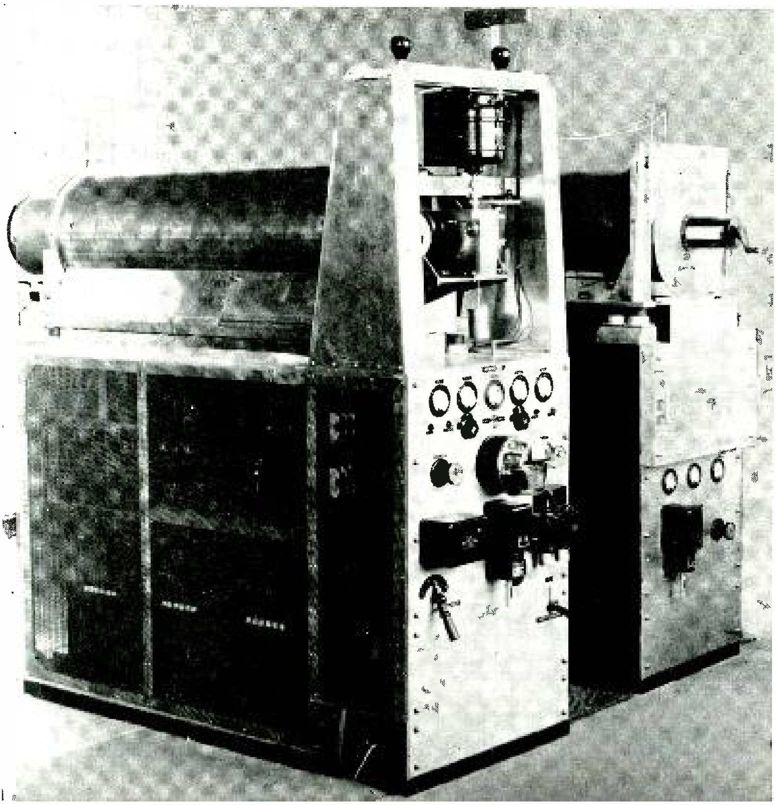

Fig. 2.—Front view of variable frequency transmitters.

is made considerably longer than the inner conductor to reduce influence of end effect when the frequency is varied through the band and to make the wave length correspond closely to the physical length of the inner conductor. It is also to be noted that the inner conductor is generally slightly shorter than a quarter-wave length due to its end capacity. In this case the line lengths are 93 to 96 per cent of quarter-wave length depending upon the transmitter frequency. The end capacity is a larger percentage factor affecting the line length, at the higher frequencies. The ratio of diameters of outer to inner conductor was made 3.6 because this results in optimum power factor for the line as an oscillating circuit.[2] Also, since the power factor is inversely proportional to the diameters of the conductors, relatively large diameters were used. The outer conductor is 16 inches diameter, which was the largest size seamless copper tubing readily obtainable. There have been published a number of papers pertaining to line control for transmitters and line calculations, to which the reader is referred for detail information.[3]

A general idea of the appearance of the survey transmitter may be obtained by referring to the left-hand unit in Figure 2. The right-hand unit in Figure 2 is a second transmitter of this general type which was built later. This illustration shows the transmitter frequency-control line mounted with rubber shock absorbers on the top of the main plate rectifier. On the front panel of the rectifier are mounted the meters, relays and various controls, while in front of the frequency-control line are mounted the frequency-sweep motor and cam arrangement. The shorted end of the line faces the reader. The transmitter oscillator tube is an air-cooled 846 which is mounted on the under side of the frequency-control line just behind the rectifier panel. The glass envelope of the tube projects up through a large opening in the outer concentric conductor of the frequency control line. The grid and filament of the tube are inductively coupled to the inner conductor of the line by means of loop-shaped conductors. In order to secure proper oscillations the grid loop must be more closely coupled to the frequency-control line than the filament is coupled. The amount of coupling required was determined by experiment. These loops are assembled together in a compact unit, and the outer ends of the loops are by-passed for radio-frequency currents to the outer concentric conductor of the frequency-control line with specially constructed mica condensers. The coupling loops also serve as grid and filament power leads. The anode of the 846 tube is fitted by means of a taper socket into a cylindrical shaped unit having a large number of air-cooling fins. This cooling unit is constructed entirely of copper for maximum heat-conducting efficiency. The clamp which supports the anode cooling unit, together with a sheet of dielectric material, fits against the outer concentric conductor of the frequency-control line in such a way as to form the anode radio-frequency by-pass condenser to ground. Means are provided for unclamping and lowering the entire tube mounting assembly consisting of the anode cooling unit, the tube, the coupling loops and the by-pass condensers. These parts can then be inspected and the tube can be changed through a door in the transmitter shielding. The other end of the frequency-control line contains a ventilating fan, mounted on rubber shock absorbers, for cooling the tube anode. This end of the line is also fitted with a screen cover to protect the fan and for electrical shielding. An air-flow interlock is mounted under the anode cooling fins to trip off the anode power supply should the air flow become insufficient.

A large portion of the inner conductor of the frequency-control line, at the open-circuit end, is constructed of copper and silver-plated metal bellows. A metal rod extends from the end bellows back through the inner conductor to a cam arrangement. This provides the means for expanding or contracting the bellows which changes the length of the inner conductor, thereby lowering or raising the transmitter frequency in response to the shape of the cam. For the survey tests, this transmitter was designed with a cam arrangement to give it a linear frequency sweep between 81 and 86 megacycles at either one or six sweep cycles per minute as desired.

The transmitter control circuits are designed for switching to either remote or local control for stopping, starting, and telegraph keying the transmitter. Keying of the transmitter is accomplished by applying a high negative grid bias to the oscillator tube from a small rectifier provided for the purpose. The normal working grid bias is obtained by passing the d-c grid current through a resistor. All access doors for tube replacement, etc., are equipped with interlocks to remove dangerous voltages should a door be opened with the power on. The main rectifier, which uses six 872A tubes in a three-phase full-wave circuit, is provided with a hand-operated tap switch for the plate transformers. This enables the rectifier to supply d-c anode current to the 846 transmitter tube in several steps of potential ranging from 2700 to 7700 volts. Two filament rheostats and a voltmeter are provided for proper adjustment of filament heating power. Time delay relays protect the filaments against improper starting. An a-c overload relay and status lights are provided for the main rectifier. An over-current relay, an ammeter and a voltmeter are provided in the anode circuit of the transmitter tube while a milliammeter is provided in the grid circuit. The whole transmitter unit is well shielded.

The transmitter output leads are located at the top and front of the unit where an insulated loop conductor extends down into the concentric frequency-control line. This loop can be moved up or down to adjust its distance from the inner conductor of the line for load coupling variation. A sliding ground connection is provided on the coupling loop for balancing the output voltages.

Figure 3 shows the antenna which was used with this transmitter at the top of the Empire State Building. The antenna was one of the folded doublet type. Handles were provided by means of which the antenna could be turned to a vertical or horizontal position in order to change its polarity during the survey tests.

The right-hand unit in Figure 2 illustrates the second variable frequency transmitter which has been built of this particular type. This transmitter was built primarily for laboratory use. Although it does not have as many refinements as the first transmitter, it has certain other advantages. This transmitter is not equipped with a plate power rectifier and the control circuits are not so elaborate, the transmitter control being entirely local. However, the outer conductor of

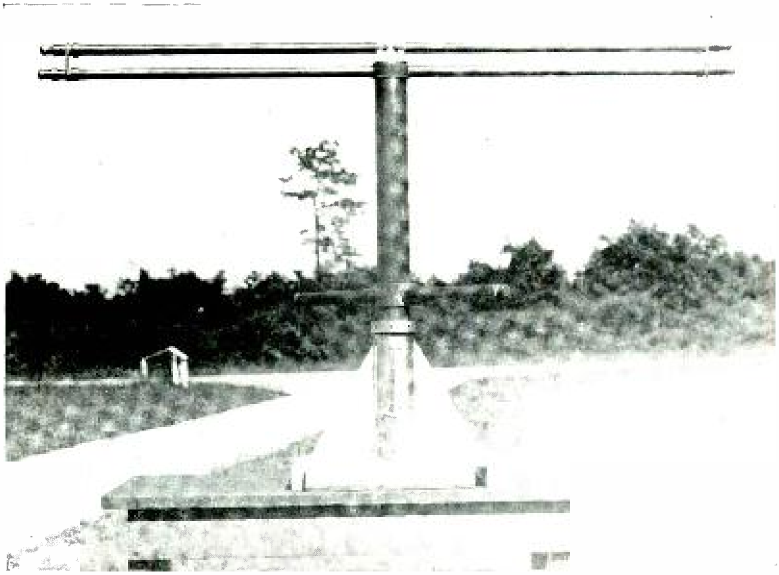

Fig. 3.—Folded doublet type of antenna.

the frequency-control line is longer which gives the transmitter a greater possible frequency range and the metal bellows section of the inner conductor is longer, which provides a greater range of continuous frequency adjustment. Instead of the cam arrangement, the second transmitter is provided with a hand-operated crank with a vernier scale, which turns a threaded rod for changing the length of the inner conductor of the frequency-control line. This permits very accurate frequency adjustment. There are also other improvements in the second transmitter such as larger radio-frequency by-pass condensers for the grid and filament coupling loops and the greater accessibility of the transmitter tubes by the easy removal of the upper section of the front panel. These various features of the second transmitter make it particularly useful for general laboratory work.

PERFORMANCE

The power output from each of these transmitters is 500 to 1000 watts depending upon the frequency and the transmitter adjustments. The transmitter plate efficiency varies from 40 to 50 per cent. This also depends upon the frequency and the transmitter adjustments. The first transmitter has a maximum variable frequency range of 8 per cent at about 85 megacycles. The second transmitter has a maximum frequency operating range of 40 to 100 megacycles by substituting different lengths of inner concentric line conductor. The continuously variable frequency range of the second transmitter is 7 per cent at 40 megacycles and 12 per cent at 100 megacycles. The frequency of these transmitters may be quickly and easily set to any value in the adjustable range with less than one-tenth of one per cent error from the calibrated value. For optimum results, the size of the grid- and filament-coupling loops should be changed for any considerable change in the length of inner conductor of the frequency-control line. In general, larger loops for coupling to the frequency-control line are required for the grid and filament of the transmitter tube at the lower frequencies. With proper design these transmitters are not troubled with parasitic oscillations. However, if the coupling loops have too much inductance and are too loosely coupled to the frequency-control line, parasitic oscillations may occur. These parasitic oscillations have been known to occur at a frequency which caused the frequency-control line to oscillate on a harmonic as well as the fundamental frequency. The remedy for this parasitic trouble is to use a coupling loop assembly having lower self-inductance and to couple it more closely to the frequency-control line.

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A WIDE-RANGE VIDEO AMPLIFIER FOR A CATHODE-RAY OSCILLOSCOPE

By

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Summary—*A wide-band amplifier for electrostatic deflection of a cathode-ray oscilloscope is described, and the principles and formulas necessary for its design are presented. The amplifier is flat from about 30 cps to 7 Mc and the phase shift approximately linear over this range. It is capable of providing full screen deflection for the nine-inch Type 914 tube when the gun voltage of the latter is reduced to 1400 volts, at which reduced voltage the focusing of the spot is still satisfactory.*

The amplifier appears able to handle the requirements of television for some time to come, as it has a range more than sufficient for testing present 441-line high-definition systems.

I. GENERAL CONSIDERATIONS

IN THE course of establishing suitable television terminal equipment for use in the television courses to be given at R.C.A. Institutes, it was deemed necessary to build a wide-range cathode-ray oscilloscope for use in testing and adjusting wave-shaping devices, measuring voltages, etc. It was felt that not only could an instrument be built which would exceed the range of those at present available, but also that the experience to be gained from such a project would be of inestimable value in the teaching of television design and practise. It is obviously desirable that instructors shall familiarize themselves with the fine points and details in the actual building of commercial equipment.

In order to have a fairly large picture of the wave, it was decided to use a nine-inch oscilloscope tube, and accordingly, the Radiotron No. 914 was chosen. This has electrostatic deflection in both directions, and can give a brilliant image at quite low gun voltages, while the fact that the deflection-plate connections are brought out at the neck of the tube instead of at the socket insures that the deflection-plate capacitances are at a minimum.

In building the wide-range amplifiers required for deflection, many problems were encountered which taxed the resources of the designer. In the first place, it was hoped to have an amplifier flat to ten megacycles, and yet have enough voltage output to give full screen deflection. In the design of ordinary video amplifiers, the main considerations that determine the design are gain and band width. However, for adequate electrostatic deflection, a further factor enters into consideration, and that is, the maximum voltage output possible from the last stage. For a given tube (hence tube capacitance), and given band width, the value of the load resistor (hence gain), and the maximum "B" supply voltage and "C" bias (hence grid swing for Class A operation), are fixed. As a consequence, the maximum peak-to-peak voltage output is fixed.

If this voltage is insufficient for deflection, then a larger last-stage tube must be chosen. Unfortunately, however, this larger tube may have sufficiently greater electrode capacitance to require a lower load resistor for the desired band width, so that the gain may be reduced

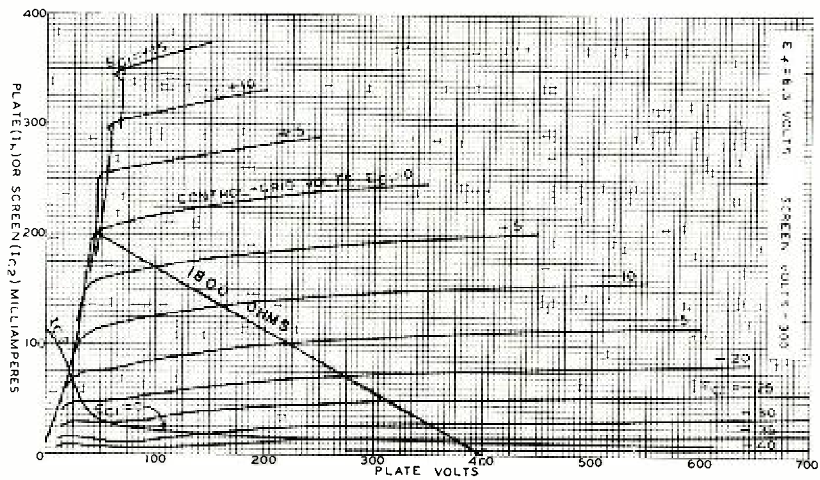

Fig. 1—RCA-807 tube average plate characteristics with E_{c1} as variable.

to a point where it nullifies the greater grid swing possible in this larger tube, and hence results in no greater output voltage. Furthermore, due to the fact that video amplifiers are of the modified resistance-coupled type, it is desirable to operate them Class A_1—i.e., with no positive grid swing. A further consideration is that even if the negative grid region has sufficient range for the voltage output required here, it should not be too non-linear.

In building this amplifier, several compromises had to be made in the design. The deflection sensitivity of the forward plates of the 914 Radiotron is 0.204 mm./volt for 2500 volts on the gun. The deflection voltage has to be balanced to ground so that the average potential of the deflection plates to ground is zero, under which condition minimum defocusing of the spot occurs. This means that the output stage has to be push-pull, but this is, if anything, an advantage, since then the voltage output per tube is half of the total required for full screen deflection. For the above deflection sensitivity, a total output of 1000 volts, or 500 volts (peak-to-peak) per tube, is required for an 8-inch deflection, which is the maximum diameter of the fluorescent screen of the 9-inch tube. This is more than any normal size tube can furnish; hence, it was decided to operate the 914 tube at a sufficiently reduced gun voltage so that maximum deflection could be obtained with about 350 volts (peak-to-peak) per tube, since the electron beam is not so "stiff" at this reduced voltage. It was felt that a pair of 807 tubes could furnish this value per tube over a band width somewhat less than 10 Mc. The final design resulted in a gun voltage of 1400 volts, a frequency response substantially flat to 7 Mc, and a deflection of 8 inches. This deflection is generally more than adequate for vertical deflection, since the vertical amplitude for a reasonable horizontal sweep will carry the beam into the rounded edges of the screen. For horizontal deflection it is sometimes advantageous to spread the sweep beyond the edges of the screen in order that a small portion of the wave may be expanded and studied more carefully. Hence, it was decided to employ the rear deflection plates for horizontal deflection, since their deflection sensitivity is 0.260 mm./volt. Their capacitance is 3 $\mu\mu f$ instead of 1.5 $\mu\mu f$, or not sufficiently greater to offset their greater sensitivity.

II. Design of Output Stage

In Figure 1 is shown the family of curves for an 807 tube. These are the same as for a 6L6 tube, but the latter has higher inter-electrode capacitances, is not quite as rugged, and does not lend itself quite as well to mechanical layout as the former tube. If we choose a "B" supply of 400 volts, and 300 volts on the screen, then a plate load resistor of 1800 ohms just about hits the knee of the 0-volt grid curve. The d-c component is 100 milliamperes, and the bias is -15 volts. The plate dissipation is 100 milliamperes at 220 volts or 22 watts, which is within the permissible rating for this tube. Experimental check of the screen-grid current showed that the dissipation at this electrode was not excessive.

The output voltage was about 350 volts (peak-to-peak), or 700 volts for two tubes in push-pull, and this, for the reduced gun voltage, was sufficient to give 8 inches of deflection. It will be observed that the d-c plate voltage is 220 volts, or less than the screen voltage, but this is permissible, if not conventional, as long as the knee of the curves is avoided, and the screen dissipation is not excessive.

The next question was how far could the frequency range be extended for the above plate load resistor of 1800 ohms. The total output capacitance, C_t, which includes plate-to-cathode, deflection plate-to-ground, and stray wiring, was found to be about 22.5 $\mu\mu f$ for each

807 tube. This value was too high to give a band width of anywhere near 10 Mc when used in conjunction with an 1800-ohm resistor in a "shunt peaking" circuit, so a more complicated circuit arrangement employing an m-derived section was used. This is shown in Figure 2. The values for the various parameters in terms of C_t and ω_o, the so-called cut-off frequency for an m-derived section, are

$$\left.\begin{aligned} L_1 &= \frac{1.6R_L}{\omega_o} \\ L_2 &= \frac{1.067R_L}{\omega_o} \\ R_L &= \frac{2}{\omega_o C_t} \\ C_2 &= 0.3C_t \end{aligned}\right\} \qquad (1)$$

Strictly speaking, the circuit employed is a half-section constant K, connected at its tee end to a series m-derived half-section, and the

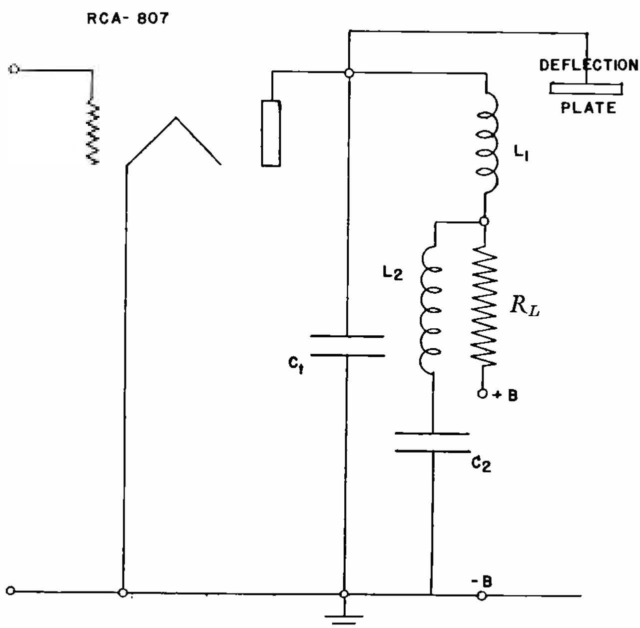

Fig. 2—Output circuit to CRO tube.

latter then terminated (at its π end) by a resistance equal to its nominal image impedance. Furthermore, input capacitance C_t is to be regarded as half, to act as part of the first-mentioned, constant-K

half-section, and the other half of C_t to act as a shunt capacitance to the half-section in order to level off the driving-point impedance and thus flatten the gain.

Referring to Eq. (1), since R_L and C_t are given, ω_o is determined and comes out to be 49.4×10^6 radians/sec., or $f_o = 7.89$ Mc. The highest frequency, f_h, to which the stage will be flat is about 89 per cent of this, or 7 Mc.

The next problem was to obtain non-inductive 20-watt resistors for R_L. After some testing, a ceramic placque-type, wire-wound resistor was found satisfactory. The nearest commercial value was 2,000 ohms, so each (for each 807) was shunted by a 20,000 ohm 1-watt carbon resistor. The coils L_1 and L_2 were of the single-layer solenoid type wound on a ½-inch bakelite form. The number of turns was adjusted so that L_1 resonated with C_t at 4.13 Mc, corresponding to a value of 65.9 microhenries, and L_2 resonated with C_t at 4.9 Mc, corresponding to a value of 47.4 microhenries. These adjustments were conveniently made by setting the signal source to the appropriate frequency mentioned above, making the desired coil the entire a-c plate load (by by-passing R_L), and "stripping down" the coil until maximum deflection was obtained on the oscilloscope screen. A wave form reasonably sinusoidal in shape is necessary from the signal source. The above-mentioned coils showed a considerably higher Q than the universal-wound type, and helped f_h to approach f_o as closely as it did.

It was found that for this type of circuit, the condenser C_2 is best made adjustable in order to obtain the proper, flat response, which tapers off beyond f_h. If C_2 is too great, the frequency response shows a dip in the center of its range, followed by a peak. As C_2 is decreased, the dip becomes less, as does also the peak, and the latter also moves out towards a higher frequency. If C_2 is just correct, the response is flat up to f_h, and then tapers off. If C_2 is too small, the response curve shows a peak in the middle of its range, and then drops off rapidly. This information enables the stage to be adjusted with relative ease. For this condenser, a disc about 1¼ inches in size was used as one plate, and the oscilloscope tube housing (iron, copper-clad) was used as the other. A screw soldered to the disc, and operating through a bakelite bracket, served to adjust this condenser. The adjustments for R_L, L_1, and L_2 are not critical when the above procedure is employed.

III. Design of Inverter Stage

There are two types of inverter stages for going from single-ended to balanced operation: (1) The type that gives the 180° phase shift by feeding a portion of the output of one stage back to a stage in push-pull with the one preceding the first-mentioned, and (2) the type that uses a tube having plate and cathode coupling to give the 180° phase shift needed for the grid excitation of the balanced stage. The former was not deemed advisable because at the higher frequencies there is an appreciable phase shift in each stage so that the phase inversion will be less than 180°. The latter method gives some de-amplification, but uses one less tube which could be used to restore any such loss in gain. Moreover, the inversion is practically 180° up to the highest frequency desired, and the inverse feed-back obtained tends to decrease non-linear distortion as well as the tube's input capacitance. These features permit a somewhat higher gain in the preceding stage.

Since the input voltage desired per 807 grid is 15 volts, a considerably higher input voltage is needed for the inverter stage. This in turn means a tube capable of taking a net grid-to-cathode swing of about the same value of 15 volts, and this in turn suggested a 6L6 tube. This also has a high transconductance and, therefore, gives an appreciable fractional gain even if a low load impedance is employed. Thus, with about 250 ohms in the cathode and plate circuits, an input voltage of 37 volts (peak) gives an output of 15 volts across the plate load and across the cathode load. With such a low load impedance, no peaking is required up to 7 Mc, and the phase shift on either side is also small, so that the difference in phase shift is negligible. It will be found that a 6L6 tube will just about deliver the output voltages required for the two 807 tubes without objectionable non-linearity.

IV. Design of Preceding Stages

Since the inverter requires such a high grid signal of 37 volts peak, and since the gain of any high transconductance stage for the desired pass band is between three and five, it is evident that the tube feeding the inverter must have an appreciable grid swing. Consequently a 6L6 tube was employed for this stage. Ordinary shunt peaking was employed. The total capacitance (output of this tube and effective input of the inverter, plus stray wiring) was 33.3 $\mu\mu f$. A value of 300 ohms was used for the plate load resistor, and the response was flat to 8 Mc, which is better than the output stage.

The two tubes preceding the above stage handle a low level of signal, hence 1851 tubes were employed. The first stage used shunt peaking, but the second stage employed an m-derived type of section. The reason for this was two-fold: (a) to obtain more gain, and (b) to enable a higher plate load resistance of 1000 ohms to be employed. It was found that the second stage would tend to overload before overloading the output stage if shunt peaking and consequently lower plate load resistors were employed. The first stage was flat to 7.5 Mc; the second, to 7 Mc.

V. Low-Frequency Response

The low-frequency response was tested by using a 30-cycle square-wave generator constructed at R.C.A. Institutes. It consists of a multi-vibrator locked at 30 cycles to the 60-cycle supply, followed by a two-stage "clipper". In this way unavoidable hum patterns could be rendered stationary and possibly less objectionable, and furthermore, the frequency stability of the square wave was much better than if an ordinary sine-wave, beat-frequency oscillator were fed directly through the clipper stages to give the desired square wave. However, the latter source was then used as a precautionary check to note if the transmission of square waves through the amplifier at various frequencies up to several hundred per second was satisfactory.

The considerations pertaining to the low-frequency response of an amplifier have been discussed in a previous article.* It was noted there that the overall response of an amplifier could be expressed as a series of products of time constants, in a form very similar to Foster's Reactance Theorem. Let the various circuit parameters be as denoted in solid lines in Figure 3, except that a second subscript denotes the particular stage of an n-stage amplifier in which the particular circuit parameters are located. Let $T_{LF} = R_{LF}C_F$, where R_{LF} is R_L and R_F in parallel, $T_g = R_gC_g$, and $T_F = R_FC_F$. Then the gain of n stages can be expressed as

$$\alpha = [G_{m1}G_{m2} \cdots G_{mn}]\ [R_{L1}R_{L2} \cdots R_{Ln}] \times$$

$$\left[\frac{\omega^n(\omega - j/T_{LF1})(\omega - j/T_{LF2})\cdots\cdots(\omega - j/T_{LFn})}{(\omega - j/T_{F1})(\omega - j/T_{g1})(\omega - j/T_{F2})(\omega - j/T_{g2})\cdots(\omega - j/T_{Fn})(\omega - j/T_{gn})}\right] \tag{2}$$

As pointed out*, if T_F in each stage is sufficiently large so that j/T_F is small compared to the lowest angular frequency ω desired, then the above complex quotient reduces to unity if the various T_{LF}'s and T_g's are equal in pairs, and the gain becomes equal to the product of the first two brackets of the right-hand side of the equation, i.e., constant amplitude response and zero phase shift. This is the relationship normally employed in video amplifiers.

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* "Some Notes on Video Amplifier Design," A. Preisman, RCA Review April, 1938.

However, T_F in each or some stages may not be sufficiently large, particularly when we wish to go down to as low a frequency as 30 cycles. This condition becomes especially troublesome when high-transconductance tubes are employed, and the "B" power-supply voltage is not sufficiently high. In this case the d-c component of the tube is generally large, and so a relatively low value of R_F must be employed in order not to obtain too low a plate voltage. Under those conditions T_F may be so low as to affect the gain. The main effect is that a square wave impressed upon such a stage comes out, at best, bowed instead of flat on the two half-cycles.

The parameters shown in broken lines in Figure 3 eliminate this

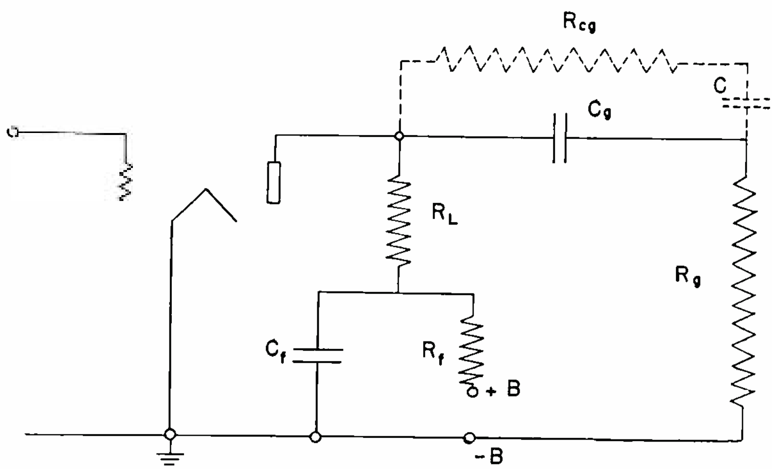


Fig. 3—Circuit for low frequency compensation.

difficulty. An additional resistance R_{cg} is employed as shown. This amplifier would be flat down to d.c., but in order to avoid this and the resultant instability an additional condenser C is added. If this be large enough so that its reactance at the lowest frequency is small compared to R_{cg}, then the behavior of the stage is affected very little by its presence. Although this "build-up" circuit would appear to add more wiring capacity to the stage and thus affect adversely the high-frequency response, actually, in practise, the total stray capacity may be changed little due to the fact that this circuit permits a lower value of C_g, so that the size of this latter unit may be sufficiently reduced to balance the presence of the other two parameters.

The proper values for these parameters are given by the following formula based on a value of R_L determined by the high-frequency response, and a maximum value of R_g determined by the permissible grid gas current of the following stage:

$$\frac{R_g}{R_L} = \frac{R_{cg}}{R_F} = \frac{C_F}{C_g} \tag{3}$$

The circuit is not critical and appreciable departures from these ratios can occur in practice especially if the blocking condenser is rather small. The actual procedure employed is to use rheostats for R_g and R_{cg}, and vary these until the square wave comes through the stage undistorted. Then small carbon resistors of the proper value are substituted in place of the rheostats.

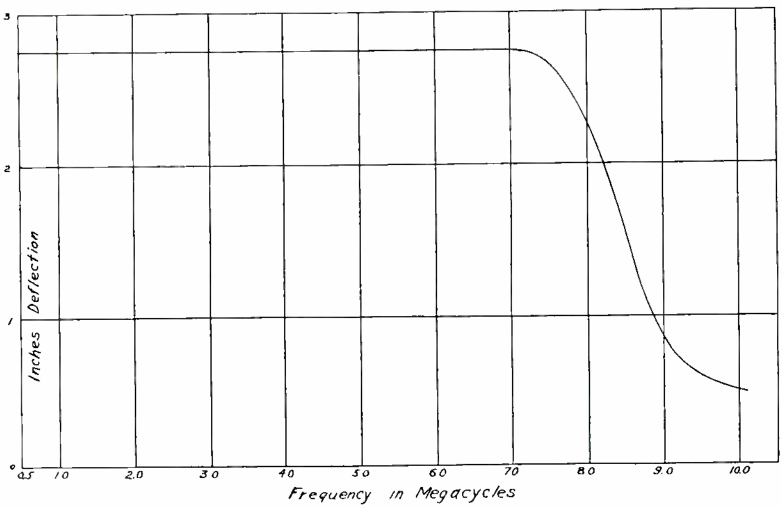


Fig. 4—Frequency response of cathode-ray oscilloscope.

VI. Grid-Bias Considerations

The low-frequency response of an amplifier can be markedly affected by the presence of a grid-bias resistor in the cathode circuit (self-bias), if this resistor is not adequately by-passed. Thus, let the bias resistor be R_c, and its by-pass condenser, C_c, and the two in parallel be denoted by Z_c. Similarly, let R_F and C_F in parallel be denoted by Z_F. Then the low-frequency gain of such a stage can be written as

$$\alpha = G_m R_L \left[\frac{1 - Z_F/R_L}{1 - G_m Z_c} \right] \tag{4}$$

If $Z_F/Z_c = G_mR_L$, then the gain becomes simply G_mR_L, or a real number; there is no net phase shift nor amplitude response variation at the plate terminal. This is due to the inverse-feedback characteristic of Z_c cancelling out the lagging phase shift and amplitude-response characteristics of Z_F. Since we have Z_F/Z_c equal to a real number, G_mR_L, the two impedances must have equal phase angles, hence, equal time constants, and the individual components must have impedance ratios of G_mR_L. Thus,

$$\frac{R_F}{R_c} = \frac{C_c}{C_F} = G_mR_L \tag{5}$$

if we do not desire any effect from Z_F. However, ordinarily we desire Z_F to cancel the effect of R_g and C_g in the following grid circuit, so that we must not allow Z_c to prevent this interaction. This in turn

(A). (B).

Fig. 5—(A) 30-cycle square wave—(B) 60-cycle square wave.

means a sufficiently large value for C_c to eliminate such feedback. For most of the stages the requisite value was impractically high. Hence, fixed bias was employed by using a small 150-volt regulated power unit, and reducing the voltage to the desired bias value for each stage by means of individual resistance voltage dividers. The individual voltage dividers, together with the regulated voltage source, prevented common coupling between the various grids.

For the first stage, however, self-bias was deemed more advisable since it permitted a much higher grid resistor to be employed, and, hence, a smaller coupling condenser. These reduce the loading on a source, such as the attenuator, both at low and high frequencies.

The common self-bias resistor in the final push-pull stage made low-frequency adjustments difficult in that the 807 tube fed from the plate side of the inverter required careful adjustment of its input circuit in order that a distorted square-wave voltage be not developed across the bias resistor. (The latter's low value of 75 ohms made by-passing impractical.) Such a distorted voltage would react on the other 807 grid to distort this side, too, even if the normal voltage input from the cathode side of the inverter was undistorted. By careful adjustment of the grid resistors and build-up circuit, and by observation of the results directly on the oscilloscope screen, a satisfactory 30-cycle square wave was obtained.

VII. Additional Precautions

In a system whose response goes up so far in the frequency spectrum, it is to be expected that commercial condensers should cease functioning as such, and become inductive, instead. For this reason

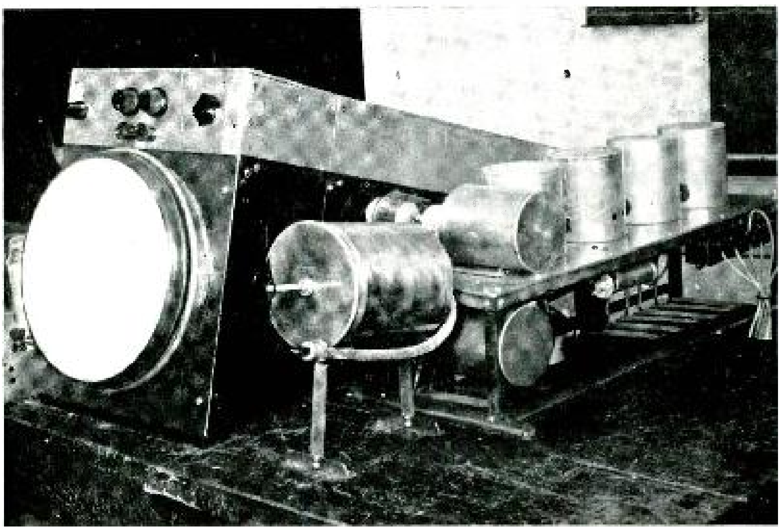
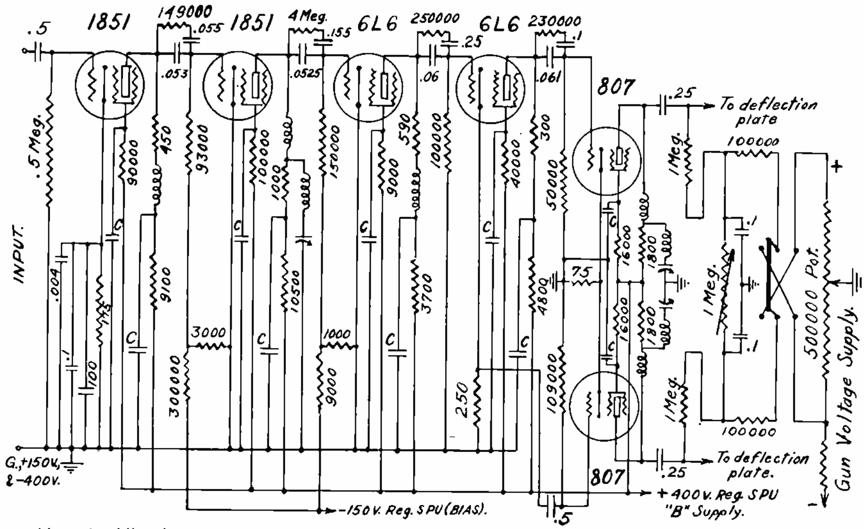


Fig. 6—The amplifier showing the oscilloscope tube and housing.

it was found necessary to shunt the 8-μf oil-filled by-pass condensers with 0.1 μf paper condensers and these in turn with 0.004 μf mica condensers. Each unit of the parallel grouping thus affords by-passing over a suitable overlapping portion of the spectrum, and so low is the Q of a condenser in the range where it acts as an inductance that no resistive de-coupling between the units was found necessary to prevent parallel resonance with consequent lack of by-passing. A further advantage is that only the small by-pass condensers need be located close to the tube in the shield can. On the other hand, no advantage was noted when mica condensers were shunted across the grid-coupling paper condensers, and so the former were omitted here.

Grounds for each stage presented some difficulty, since stray ground currents of very high frequency can cause instability in the amplifier. All grounded ends of the units in a stage were connected to one point in the stage-shield can to prevent ground current flow in the copper chassis. The only (unavoidable) ground current was that flowing into the input admittance of the following stage (grid capacity to ground) and this effect was not apparently appreciable.

In order further to de-couple the stages at the higher frequencies, a bifilar-wound inductance was inserted in each heater circuit close to the socket, and a by-pass condenser connected between one side of the heater circuit (at the socket) and ground. This minimizes coupling through the various cathode-to-heater capacitances and common heater-



Note 1.-All values in ohms or microfarads unless otherwise noted.
Note 2.- C represents an 8 mfd., 0.1 mfd., and 0.004 mfd. (mica) condensers in parallel.
Note 3.- Heater Circuit not shown.

Fig. 7—Schematic circuit diagram.

lead reactance. The author wishes to acknowledge the helpful suggestions by Mr. Otto Schade, of the Radiotron Division, RCA Manufacturing Company, in regard to the above precautionary measures.

VIII. Power Supplies

The plate supply for the entire five stages is from a regulated socket power unit. Six 2A3 tubes in parallel function as the series rheostat device. Due to the large load-current drain—about 350 milliamperes—it was found advisable to divide the load between two power supplies. At the suggestion of Mr. S. W. Seeley of the R.C.A. Patent License Division, an unregulated supply unit was connected in parallel with the regulated unit, and the input filter condenser of the former varied until it took slightly more than half the load. Subsequent adjustments of the d-c amplifier reduced the ripple component to a negligible amount as viewed on an auxiliary oscilloscope. One pair of such units are required for each amplifier.

IX. Conclusions

In Figure 4 is given the overall frequency response, as measured by screen deflection (output) versus constant sine-wave voltage input. As will be noted, the amplifier is flat to 7 Mc and then drops off rather sharply, although, if sufficient signal is available, it can be used on sinusoidal inputs to 10 Mc. In Figure 5 is shown a 30-cycle square wave and a 60-cycle square wave. The slight curvature represents an unavoidable hum voltage present in the square-wave source. If the latter is not locked to the a-c power supply this hum pattern will be found to travel across the wave. Finally, in Figure 6 is shown a photograph of the amplifier together with the oscilloscope tube and housing, and in Figure 7, a schematic circuit diagram.

The writer wishes to express his appreciation of the services of Mr. A. H. Smith, Radio Technician of R.C.A. Institutes, for his excellent work in building and testing this unit.

MEASUREMENTS OF ADMITTANCES AT ULTRA-HIGH FREQUENCIES

By

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Summary—*A substitution method is described for determining admittances at ultra-high frequencies. Typical measurements of resistors and insulators are given.*

The method employs a short transmission line excited by an ultra-high-frequency oscillator. The receiving end of the line is short-circuited and the sending end is shunted by a variable capacitor and the unknown admittance. A vacuum-tube voltmeter indicates the resonant voltage at a point on the line. The unknown admittance is removed, resonance restored by the variable capacitor, and the same voltage obtained by sliding a known resistor along the line. The unknown admittance is then determined by the frequency, capacitance change, and the position of the known resistor. With proper line constants, the equivalent sending-end resistance of the resistor is equal to the product of its resistance and the square of the ratio of the total line length to the distance of the resistor from the receiving end.

INTRODUCTION

A METHOD of admittance measurement is described* which has been found generally useful at very high frequencies. The method for the determination of the effective resistance of fixed resistors will first be given and then modifications will be described which adapt it to the measurement of the power factor and dielectric constant of insulators and the measurement of the input and output impedance of vacuum tubes. The results of measurements are given for a number of resistors and insulators.

I. THE EFFECTIVE RESISTANCE OF FIXED RESISTORS

A. General—For measurement purposes at high frequencies, it is desirable in many cases to use fixed resistors of medium or high values. Frequently it is assumed that these resistors have the same effective values of resistance at the frequency of use as they have in direct-current use. In general, this assumption is not justified by the facts; the error depends on the type of resistor, its size, and the frequency at which it is used. The error can be fifty or sixty per cent in some cases.

* The method of measurement described in this paper was presented at the U.R.S.I. meeting in Washington, D. C., on April 30, 1937.

The present work is concerned with the investigation of the behavior of several types of resistors at frequencies of 30 to 250 megacycles per second, and of nominal resistance value up to 200,000 ohms. Some of these types appear to be better suited than others for the purpose of measurements at very high frequencies and it is hoped that by utilizing the results to make corrections to the direct-current values, other experimenters can have available resistors whose values are known in these ranges to within a few per cent.

Contrary to what might be expected, the effective resistance of a fixed resistor at high frequencies is generally less than its direct-current value. This increase in conductivity at high frequencies is called the "Boella" effect.[1,2] An attempt has been made to explain this behavior on the basis that the resistor acts as a transmission line having uniformly distributed resistance in series and capacitance in shunt.[3,4] This theory has not led to numerical results agreeing with experiment due probably to lack of knowledge of the precise characteristics of the line which corresponds to the resistor. The present measurements show that there can be also a counter-effect, similar to skin effect, due probably to a change in distribution of the current flow in the resistor, which tends to lower the conductivity at the higher frequencies.†

B. Method of Measurement—The method of measurement employs a resonant uniform line of low resistance, less than a quarter wavelength long, short-circuited at one end, and loaded by capacitance at the open end. If an emf is induced in such a line near the shorted end, the form of the voltage distribution along the line will be sinusoidal, the voltage at any point being given by $V = V_m \sin \beta x$, where V_m and β are constants and x is the distance from the point to the shorted end. For a short line, $\sin \beta x = \beta x$ approximately so that the voltage distribution is nearly linear. If this induced emf is constant and a resistor R_1 of high value is shunted across the line at a point near the high potential end, the voltage at all points along the line will be reduced in the same proportion and the form of the voltage distribution will be unaltered. If this resistor is removed and a second resistor R_2 of lower value is shunted across the line, a position

† A recent paper by L. Hartshorn[5] shows a very good agreement between the theory and measurements. In general, the resistors which were measured were of a solid type; the name of the maker was not specified. In a few cases, the description of the resistors indicates that they are similar to the I.R.C. type resistor which we have measured. One of these, a 50,000-ohm resistor, gives fair agreement with the theory, while another of 1-megohm value departs rather widely from the theory. The latter resistor shows an apparent skin effect such as we have observed. In general, Hartshorn finds a greater decrease from the d-c values than we have observed for resistors of this type.

for this second resistor can be found, between the position of the first resistor and the shorted end, where the same reduction in voltage will be produced at all points on the line and, hence, the voltage distribution will be identical with that obtaining for the first resistor. This condition is indicated by the same deflection of a vacuum-tube voltmeter located at a fixed point on the line. For this to be the case, the power dissipated in the resistor in each instance must be the same and, therefore, the ratio of the effective resistances of the two resistors is equal to the ratio of the square of the respective voltages at the two positions. Hence, if the form of the voltage distribution is known, the effective resistances of resistors can be compared. As mentioned above, the form of the voltage distribution is known theoretically in the case of a uniform line. For a short line the distribution is nearly linear and the voltages are proportional to the respective distances x_1 and x_2 from the shorted end. It follows that

$$\frac{R_1}{R_2} = \left(\frac{x_1}{x_2} \right)^2$$

For the general case, the voltages are proportional to $\sin\beta x$, where β is 2π divided by the wavelength and $\dfrac{R_1}{R_2} = \left(\dfrac{\sin \beta x_1}{\sin \beta x_2} \right)^2$. However, due to the fact that the short at the end of the line does not have zero reactance, there is some uncertainty as to the point from which to measure the line length. Furthermore, due to the possibility of some non-uniformity of the line, it is more precise to determine experimentally the voltage distribution along the line. A method of accomplishing this, which also gives additional information of value with respect to the resistors themselves, is described later. The mathematical theory of the line used as an auto-transformer is given in the Appendix.

C. The Apparatus—Figure 1 is a photograph of the apparatus as used for resistor measurements. The box at the center contains three complete oscillator circuits, which cover the frequency range from 30 to 300 megacycles. Each circuit employs an RCA-955 acorn-type triode. A Colpitt's circuit is used with a variable capacitor having a split stator and grounded rotor. The plate voltage is fed to each tube through individual resistors and any one of the oscillators is put into operation by closing its heater circuit. These circuits are controlled by the toggle switches seen on the end of the box.

The line consists of a copper rod, a quarter of an inch in diameter and about a meter long, mounted a quarter of an inch above the brass bed plate. The characteristic impedance of the line is about 100 ohms. The line is grounded to the base at the extreme left end and is terminated by an adjustable circular-plate capacitor at the right end. Additional grounding blocks can be applied at a number of positions along the line to shorten it for the higher frequencies. In the photograph, the line is shown adjusted for operation at 250 megacycles, its length being the distance from the circular-plate capacitor to a shorting block close to the right end of the oscillator box. Another shorting block near the left end of the oscillator box reduces coupling effects due to the unused section. Coupling between the oscillator and the line is effected by a single-turn loop inside the oscillator box, one terminal of which is grounded and the other terminal brought outside and connected to a clip which slides on the copper rod of the line near the

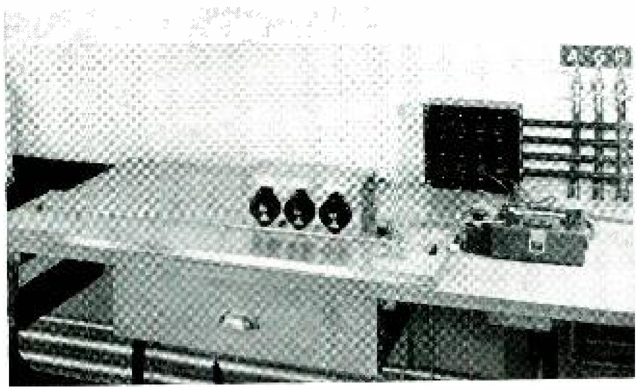

Fig. 1—Set-up of the apparatus used for the resistor measurements.

shorted end. Variation in the coupling is obtained by adjustment of this clip. The vacuum-tube voltmeter which is used as an indicator is seen on the right of the oscillator box. It employs an RCA-955 tube connected as a triode plate detector with the plate current balanced out of the plate-circuit milliammeter by current taken from the heater supply. On the 0.02-milliampere range of this meter, full-scale deflection is obtained with about 0.3-volt a.c. on the grid of the tube. While the input loading due to electron transit time is worse with this circuit than with the diode connection, the sensitivity is much higher and the effect of input loading can be reduced by connecting the vacuum-tube voltmeter at an intermediate point on the line as shown. The connection from the grid of the tube to the line is preferably made through a small capacitor and a very short lead.

The resistors under test are provided at one end with a heavy brass block which rests on the bed plate, while the other terminal rides on the copper rod. A paper scale is mounted on the brass bed plate to read the position of the resistor; the reflected image of the resistor in the bed plate is used to avoid parallax in the readings. The zero of the scale is at the permanently grounded end of the line.

D. Calibration of the Line—To calibrate the line it is assumed that the effective resistances of resistors of low value are the same as their direct-current resistances. The procedure is explained with reference to Figure 2 which gives the results of a calibration at 30 megacycles using the entire length of line. In this calibration one-watt resistors of the I.R.C. type were employed.

A resistor having a nominal value of 250 ohms is first placed on the line at the 10-centimeter point of the scale. The line is adjusted to resonance and the vacuum-tube voltmeter is read. This resistor is then removed and the position of a 500-ohm resistor is found at which the deflection of the vacuum-tube voltmeter is the same. If necessary, the line is readjusted to resonance. The corresponding positions of a series of higher-valued resistors are similarly determined with an occasional recheck of the deflection for the 250-ohm resistor at the 10-centimeter setting. Then, assuming the voltage at the 10-centimeter point to be unity, the voltage at the positions found for the other resistors is calculated from the square root of the ratio of the measured d-c values of these resistors to the d-c value of the 250-ohm resistor. These values are plotted in Figure 2 using solid circles. The observations fall on a curve which is very nearly a straight line, corresponding to the voltage distribution on a transmission line which is short relative to a quarter wavelength.

The calibration is then repeated starting, in this case, with a 500-ohm resistor at the 10-centimeter position and finding the positions of higher-valued resistors giving the same voltmeter deflection. The resulting voltages are plotted in Figure 2 using dots in circles. Similarly, starting with a resistor of 1000-ohms nominal value, the points are plotted with dots in squares. All of the points which correspond to resistors of nominal value of 10,000 ohms or more are marked with the nominal resistance value.

It is evident that all of the resistors up to 15,000 ohms fall, within experimental error, upon a common curve while the higher-valued resistors fall above the common curve. Since it is not probable that the per cent change in the lower-valued resistors with frequency would be the same for all values, it appears that these resistors do not depart substantially from their d-c values at 30 megacycles. On the other hand, the points calculated for the higher-valued resistors on the basis of the d-c values, fall above the curve showing that the effective resistances of these resistors are lower than their d-c resistances. Assuming that the calibration curve is correct as determined by resistors up to 15,000 ohms, and that the resistors themselves up to this value are rated by the d-c values, the effective values of still higher resistors are determined by comparison with the known resistors on the calibrated portion of the curve, preferably by taking the mean of several comparisons. In turn, these calibrated resistors can be placed at the extreme high-voltage end of the line and compared with lower-valued

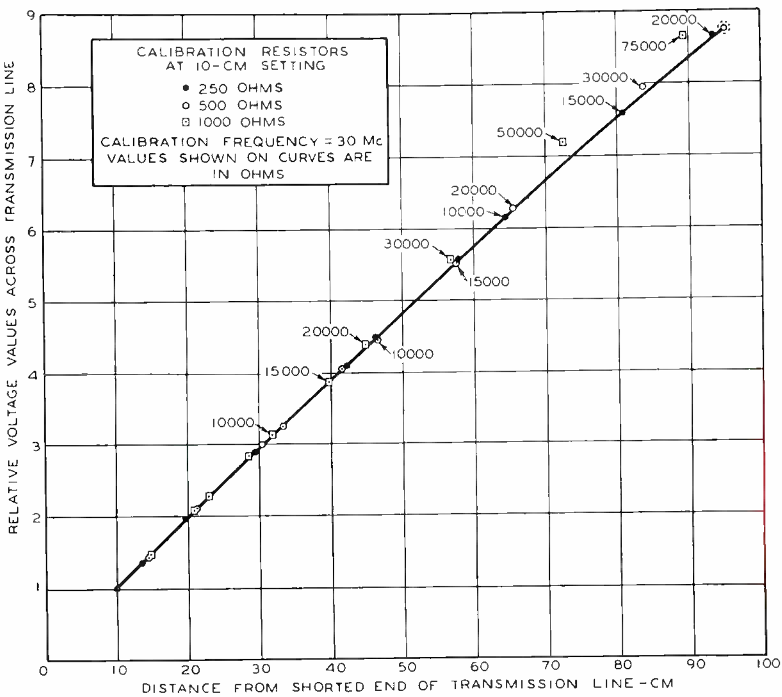

Fig. 2—Calibration curve of transmission line at 30 Mc.

resistors falling on the calibrated portion. This procedure gives the voltage at the high end of the line. Such a point is denoted in Figure 2 by the dot surrounded by two circles. For greatest accuracy it is desirable to use high-valued resistors at the high-voltage end of the line. To avoid errors due to disturbance of the voltage distribution on the line, it is desirable to use low-valued resistors only at the lower-voltage portion of the line. Having obtained a complete calibration curve and possessing a standardized set of resistors, it is then possible to measure any other resistors. It will be noted that the method does not depend upon the calibration of the voltmeter and that during a measurement the power taken from the oscillator is constant.

At higher frequencies the method of calibrating the line is the same; however, the resistors will show substantial deviation from the d-c values at lower nominal values and the stepping-up process is consequently longer.

E. Experimental Results—In Table I are given the per cent corrections to the d-c values to give the effective values for 1-watt I.R.C. ceramic resistors of nominal value from 8000 to 200,000 ohms and for frequencies from 30 to 250 megacycles. These resistors employ a metallized glass rod. In addition to the "Boella" effect, there is apparently an effect of redistribution of current which opposes the "Boella" effect and which results in less change in the effective resistance at 250 megacycles than at 100 megacycles for the higher-valued resistors. A measurement of the effective resistances of five different 200,000-ohm resistors at 100 megacycles gave corrections which, for these resistors, varied by only two per cent of the d-c values.

TABLE I

Per Cent Change in Effective Resistance from Direct-Current Values
I.R.C. Ceramic-Type Resistors

Nominal Value Ohms	One Watt Size 30 Mc	50 Mc	100 Mc	200 Mc	250 Mc
8,000	0	0	— 6	— 6	—5
10,000	0	0	— 4	— 2	0
15,000	0	— 2	— 7	— 8	—4
20,000	0	. .	— 9	—10	—7
30,000	— 4	— 6	—12	—11	—6
50,000	— 7	—11	—16	—10	—6
75,000	— 8	—12	—16	—13	—2
100,000	— 9	—12	—18	—11	—3
200,000	—17	—17	—23	—17	—4

In Table II are given the per cent correction values for ½-watt and ⅓-watt I.R.C. ceramic resistors measured at 100 and 250 megacycles. These resistors also use a metallized glass rod. In the case of the 30,000-, 50,000- and 100,000-ohm resistors, two resistors were measured in each instance, as indicated by the numbers 1 and 2. Here, some variation is shown in the behavior of resistors of the same nominal value. The corrections for the ½-watt size are relatively small.

Measurements were also made at 100 and 250 megacycles on a number of ½-watt Bakelite-molded and solid carbon ¼-watt resistors. The results are given in Table III. In the case of the Bakelite-molded resistors, the 8000- and 20,000-ohm values show a markedly smaller correction value than the others. These particular samples were of

TABLE II

Per Cent Change in Effective Resistance from Direct-Current Values
I.R.C. Ceramic-Type Resistors

Nominal Value Ohms	½ Watt Size 100 Mc	½ Watt Size 250 Mc	⅓ Watt Size 100 Mc	⅓ Watt Size 250 Mc
10,000	..	..	..	0
15,000	..	..	..	+1
20,000	—2	+4	—8	—5
30,000 (1)	—7	0	—7	—6
30,000 (2)	—6	—1	—10	—10
50,000 (1)	—5	+6	—17	—14
50,000 (2)	—10	—2	—9	—4
100,000 (1)	—9	—1	—25	—13
100,000 (2)	—7	+8	—23	—9

an old type (distinguished by a dot in the color code marking). The new type (employing a circular band in the color code marking) has terminal lugs which, for the purpose of heat dissipation, extend into the metallized glass tube for a considerable distance. These lugs because of capacitance effects through the glass tend to short out part of the resistance at very high frequencies. The solid-carbon type resistors show a large reduction in resistance at high frequencies.

TABLE III

Per Cent Change in Effective Resistance from Direct-Current Values

Nominal Value Ohms	I.R.C. ½-Watt Size Bakelite 100 Mc	I.R.C. ½-Watt Size Bakelite 250 Mc	Nominal Value Ohms	Solid-Carbon ¼-Watt Size 100 Mc	Solid-Carbon ¼-Watt Size 250 Mc
8,000	—4	+3	5,000	—17	—31
10,000	—41	—46	10,000	—23	—36
15,000	—41	—42	20,000	—46	—60
20,000	—3	+2	20,000	—40	—53
30,000	—50	—50	30,000	—46	—55
50,000	—49	—54	30,000	—47	—56

II. THE POWER FACTOR AND DIELECTRIC CONSTANT OF INSULATORS

A. General—There have been a great many measurements of the power factor and dielectric constant of solid dielectrics at room temperature and at frequencies up to one megacycle. In a few instances the measurements have been carried to very high frequencies. In particular, Rohde,[6,7] and Rohde and Schwarz[8] have made numerous measurements at frequencies up to 500 megacycles. Their results are summarized in the second volume of Hollman's book, "Ultra-Short Waves".[9] In the case of solid dielectrics, no substantial difference is found in the behavior of insulators at these high frequencies from the behavior at lower frequencies, as was expected. In the present work, a large number of solid insulators were measured at about 60 and 120 megacycles. The results indicated substantially no dependence of power factor and dielectric constant upon frequency. However, a low power factor is more essential at high frequencies than at low, for the reason that the power loss in a given piece of insulator at constant voltage and power factor is directly proportional to the frequency.

With respect to temperature variation, however, the power factor increases with increasing temperature; in fact, more and more rapidly as the temperature rises. This phenomenon can be unstable or explosive in character. If enough power is dissipated in an insulator to heat it substantially, the power factor increases due to the heating which in turn causes a further increase in the dissipation of power and further heating resulting in a very great power loss and possible destruction of the insulator. This effect is important at high frequencies in high-power installations and particularly in transmitting-tube insulators where the insulator is heated initially by the filament and plate power of the tube. Thus, the requirements for an insulator for such operation are very severe; an insulator in which the losses are entirely negligible at low temperatures can be very poor at high temperatures.

There does not appear to be much in the literature with respect to the properties of insulators at high temperatures except a paper by Strainer and Zinke[10] which reports investigations on a number of German ceramics and glasses. All of the measurements in the present paper were made a. approximately 60 megacycles since the method appeared to be at least as simple and accurate at this frequency as any method using lower frequencies. Temperatures up to about 400 degrees C were employed. No attempt was made to obtain precision measurements, since only the usual accuracy sufficient for engineering purposes was required. For economy in time in the measurement of a large number of specimens, and because we were actually interested only in test conditions corresponding to the conditions of use, electrodes were not deposited upon the specimens. Some care was taken to prepare the specimens with parallel plane surfaces and fairly thick plates of the material were generally employed to reduce the error due to possible air gaps.

B. Method of Measurement and the Apparatus—The apparatus employed is largely the same as described previously. The screw of the circular-plate capacitor on the end of the line is provided with a dial to permit reading to a fraction of a turn. This capacitor is also calibrated for direct capacitance between the plates for different distances of separation in terms of turns of the screw. The diameter of the circular plates is 3 centimeters. A small, vernier, circular-plate capacitor is provided in parallel with the main capacitor for fine tuning adjustments. The line is calibrated for voltage to the bed plate at points along the line relative to the voltage across the circular-plate capacitor at the end of the line. Such a calibration for voltage ratio is made for each length of line at the average frequency of use. The procedure of measurement is the following. A zero reading of the dial is made with the circular plates in light contact as in the use of an ordinary micrometer. Then, counting the number of complete revolutions of the screw, the insulator under measurement is inserted between the plates and again with light contact, the screw dial is read. These readings together with the capacitor calibration give the capacitance C_1 with air as a dielectric. It would be an advantage to add a scale or counter to read revolutions of the screw. The screw is then tightened against the dielectric so as to minimize the air gap, the oscillator is tuned approximately to resonance, fine adjustment is made with the vernier capacitor and the deflection of the vacuum-tube voltmeter is read. The insulator is removed and the separation of the plates of the circular capacitor is reduced until resonance is again obtained with no change in the oscillator or vernier. The separation now gives, from the calibration curve, the capacitance C_2 of the capacitor with the dielectric interposed, and the ratio C_2/C_1 gives the dielectric constant with considerable accuracy. The frequency is read by means of an absorption-type frequency meter which is loosely coupled to the line. Then a resistor is placed on the line and the position found where the deflection of the vacuum-tube voltmeter is reduced to the same value as the earlier reading with the dielectric in place. From the line calibration, the voltage ratio for this point is obtained. Dividing the effective resistance of the resistor by the square of the voltage ratio gives the effective shunt resistance r across the terminating condenser due to losses in the dielectric. The power factor δ is given very accurately by

$$\delta = \frac{1}{\omega C_2 r} \text{ where } \omega = 2\pi \text{ times frequency}$$

except in the unusual case of a very high power factor, in which case the exact expression is

$$\delta = \sin \arctan \frac{1}{\omega C_2 r}$$

Since a number of minor changes were made in the apparatus for the purpose of dielectric measurements, the line was recalibrated for voltage distribution at each of the frequencies of measurement. With a set of calibrated resistors, this is done simply. A resistor of say 50,000 ohms is shunted directly across the circular-plate capacitor and the vacuum-tube voltmeter reading at resonance is obtained. This resistor is removed (completely) and the positions on the line of other lower-valued resistors are found, which give the same vacuum-tube voltmeter reading. The square root of the ratios of these latter resistors to the first is then plotted against the scale readings to give a curve of voltage ratio which has the advantage of being nearly a straight line.

It can be assumed with a fair degree of accuracy that the capacitance of the circular-plate capacitor is inversely proportional to the separation of the plates, measured in terms of turns of the screw. In this case the dielectric constant is given to an accuracy of better than ten per cent by the ratio of the turns separation with the dielectric in place and with the dielectric removed. An alternate method is to calibrate the circular-plate capacitor against a cylindrical vernier capacitor. The vernier capacitor is connected between the high-potential plate of the circular-plate capacitor and ground with very short leads. With the grounded plate of the circular-plate capacitor removed, the first operation is to obtain resonance. Replacing the grounded plate with a definite wide separation, we determine the change in the vernier capacitor necessary to restore resonance. Then the grounded plate is made to approach the high-potential plate by definite rotations of the screw and the corresponding capacitance changes of the vernier capacitor are noted. To reduce errors due to inductances of leads and the vernier capacitor itself, the measurement should be made at relatively low frequencies. The shorting block at the end of the line can be replaced by an insulating block and a coil connected across this block from line to ground to tune the system to low frequencies.

The apparatus as modified for dielectric measurements at high temperatures is shown in Figure 3. An oven of asbestos board surrounds the high-potential end of the line at the right. A cut-down Bunsen burner is put underneath the oven to heat it, the flame impinging on a perforated iron baffle plate which is above the bottom of the oven. The top plate of the oven is also perforated with a large number of small holes so as not to smother the flame of the burner. The temperature is controlled by the size of the flame. A thermocouple measures the temperature of the air in the oven. This thermocouple is removed during the actual measurements since capacitive coupling to it affects the measurements. The line is brought into the oven through a mica window and is supported just outside the oven by a clamp of ceramic having very low loss. To reduce loss of heat from the specimen under measurement, the line is made of thin-walled brass tubing which is copper plated. Also, the screw of the circular-plate capacitor is made of brass tubing. The bearing of the screw is split and the tightness is adjustable. Hard solder is used to connect the circular plate of the capacitor to the end of the line.

C. Experimental Results—In Table IV are given the results of measurements on a number of different types of insulators at room temperature and at frequencies of about 60 and 120 megacycles per second. The differences in the results at these two frequencies are

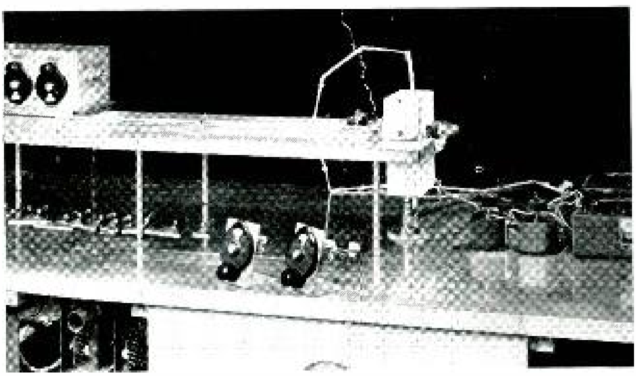

Fig. 3—Set-up of the apparatus used for the insulator measurements.

TABLE IV

Material	Dielectric Constant 60 Mc	Dielectric Constant 120 Mc	Power Factor $\times 10^4$ 60 Mc	Power Factor $\times 10^4$ 120 Mc
Fused Quartz (clear)	3.8	3.8	3	5
Fused Quartz (milky)	3.5	3.5	3	5
Alsimag No. 211	4.4	4.4	4	5
Polystyrene	2.6	2.5	5	7
Alsimag No. 196	4.9	5.0	10	13
Glass (Borosilicate No. 707*)	3.7	3.7	12	12
Italian Lavite	4.7	4.8	15	17
Glass (Nonex No. 772*)	4.2	4.2	28	25
Glass (Pyrex No. 774*)	4.1	4.2	54	51
Hard Rubber (Low loss)	2.9	3.0	57	50
Hard Rubber (Ordinary)	3.1	3.1	83	84
Glass (Soda Lime No. 008*)	6.1	6.2	106	103

* Corning Glass Works Code Number.

probably due to experimental error. The last two columns represent power factor in hundredths of a per cent.

It is a fact that at ordinary temperatures and at lower radio frequencies, the losses in many insulators are so low as to be completely negligible. The choice of insulator, therefore, should depend upon such factors as mechanical properties and cost. For example, suppose the capacitance through the insulator is 10 micro-microfarads, the frequency 1000 kilocycles, and the power factor as high as one per cent. The shunt resistance due to dielectric loss is more than one megohm, which is many times the resonant impedance of the usual circuit. However, at 100 megacycles the effective shunt resistance is only 10,000 ohms.

In Table V is shown the effect of temperature on the power factor of various insulators. The measurements were made at approximately 60 megacycles. The materials are arranged in the order of increasing power factor at 400° C. Measurements were made upon ruby mica, but were not reliable because of our inability to obtain specimens of sufficient thickness. The results indicate very low losses at 400° C.

TABLE V

Material	Power Factor $\times 10^4$ 25° C	300° C	400° C	Dielectric Constant
Ceramic A	5	2	3	3.7
Fused Quartz (clear)	3	4	6	3.8
Ceramic B	6	8	8	5.2
Alumina	13	12	17	4.2
Alsimag 211	4	15	26	4.4
Italian Lavite	16	33	60	4.7
Alsimag 196	10	40	79	4.9
Glass (Borosilicate No. 707*)	12	23	103	3.7
Glass (No. 172*)	53	..	246	5.8
Glass (Nonex No. 772*)	28	121	286	4.2
Glass (No. 776*)	22	..	304	4.5
Glass (No. 704*)	24	..	332	5.0
Glass (Pyrex No. 774*)	54	347	679	4.1
Glass (Uranium No. 332*)	36	..	694	4.8
Glass (Soda Lime No. 008*)	106	1250	..	6.1

* Ceramics A and B are developmental materials originated by L. R. Shardlow of the Research and Engineering Department, RCA Manufacturing Company, Inc., Harrison, N. J.
* Corning Glass Works code number.

III. The Input and Output Impedance of Vacuum Tubes

The same method of measurement is used for the determination of the input and output impedance of vacuum tubes, either positive or negative and including the effects due to electron transit time.

In the case of the input impedance when the loading is positive, the grid of the tube is connected by means of a short lead to the high-potential end of the line through a by-pass capacitor, the grid bias being supplied through a high resistance. The other desired operating voltages are applied to the tube electrodes, care being taken to by-pass to ground through very low reactances. Resonance is obtained and the vacuum-tube voltmeter is read. The grid connection is then opened or the tube biased to cut-off as desired, resonance restored by the terminal capacitor, and a standard resistor applied to the line at a point which gives the same deflection of the vacuum-tube voltmeter. The voltage calibration of the line then gives the resistive part of the impedance of the tube, as described above, and the change in the terminal capacitor gives the capacitive component.

In the case of negative loading, the deflection of the vacuum-tube voltmeter is obtained in the absence of both the tube loading and standard resistor. Then, both of these are applied and the resistor adjusted to give the previous reading. The positive loading due to the resistor is then numerically equal to the negative loading due to the tube.

APPENDIX

Theory of the Transmission Line Used for the Measurement of Admittance

The method of measurement described in this paper is essentially of the substitution type. The unknown admittance is connected across a parallel resonant circuit, the capacitance of which is adjusted until the circuit is in resonance with the frequency of the driving oscillator. The resonance reading is indicated by a vacuum-tube voltmeter. A resistance is then substituted for the unknown admittance. The resistance is adjusted and the capacitance simultaneously varied to obtain resonance until the same reading is obtained on the vacuum-tube voltmeter as before. The unknown admittance is then given by the value of the substitution resistance, the frequency, and the change in capacitance required to re-establish resonance.

The substitution resistance in the actual set-up consists of a fixed resistor connected across a low-loss transmission line which constitutes the inductance of the parallel resonant circuit. The transmission line is short-circuited at one end and attached to a variable capacitor at the other, or sending, end. The fixed resistor is arranged so that it can be readily connected across the line at any point along the line length.

When the fixed resistor is connected across the line at its short-circuited end, the admittance of the line at its sending-end terminals is equal to that obtained in its unloaded state, i.e., with the fixed resistor removed. On the other hand, when the fixed resistor is connected across the line at its open end, the sending-end admittance is equal to the sum of the admittance of the resistor and the sending-end admittance of the line in its unloaded state. When the fixed resistor is connected across the line at some intermediate point, the value of the sending-end admittance lies between these two limiting end values. The transmission line may thus be regarded as an auto-transformer which steps down the admittance of the fixed resistor. The effective admittance is a function of the resistor admittance, the position of the resistor on the line, and the parameters of the line. To determine the quantitative relationship between these factors, we proceed as follows:

Consider a transmission line short-circuited at its distant end, and bridged by an admittance Y at a point x centimeters from the distant end. The line length is l centimeters.

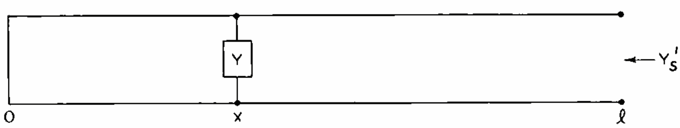

Fig. 4—Schematic representation of transmission line with bridging admittance.

The admittance at x of the portion of the line to the left of the bridging admittance is

$$Y_{sx} = Y_o \coth Px \tag{1}$$

where

Y_o = characteristic admittance of the line
P = propagation constant of the line

The vector sum of this admittance and the bridging admittance may be regarded as representing a receiving-end admittance for the rest of the line, extending from x to l. The value of this receiving-end admittance is

$$Y_r = Y + Y_o \coth Px \tag{2}$$

The sending-end admittance, at l, is

$$Y_s' = Y_o \cdot \frac{Y_r + Y_o \tanh P\,(l - x)}{Y_o + Y_r \tanh P\,(l - x)} \tag{3}$$

and this may be written, because of (2), as

$$Y_s' = Y_o \frac{Y \tanh Px + Y_o [1 + \tanh Px \tanh P(l-x)]}{Y \tanh Px \tanh P(l-x) + Y_o [\tanh Px + \tanh P(l-x)]} \tag{3a}$$

When the bridging admittance is removed, the sending-end admittance is simply

$$Y_s = Y_o \coth Pl \tag{4}$$

With the unknown admittance, Y_u, connected across the sending end of the line, the bridging admittance removed, and the tuning capacitance adjusted to obtain resonance, the total admittance at the sending-end terminals of the line is

$$Y_s + Y_u + ib_c = g_s + g_u \tag{5}$$

since resonance requires that the total susceptance be zero, i.e., $(b_c + b_s + b_u) = 0$. In the above

$Y_s = g_s + i\,b_s =$ sending-end admittance of the line when the bridging admittance is removed

$Y_u = g_u + i\,b_u =$ unknown admittance

$b_c =$ susceptance of the tuning capacitance required for resonance, when the unknown is connected

Now when the unknown admittance is removed and the bridging admittance is moved along the line until the same reading is obtained at resonance as before, the total admittance at the sending-end terminals of the line is

$$Y_s' + ib_c' = g_s + g_u \tag{6}$$

Resonance now requires that $(b_c' + b_s') = 0$. The new quantities are

$Y_s' = g_s' + ib_s' =$ sending-end admittance of the line when the bridging admittance is connected

$b_c' =$ susceptance of the tuning capacitance required for resonance when the unknown admittance is removed

Equating (5) and (6) gives us an expression for the unknown admittance in terms of the total substitution admittance:

$$Y_u = (Y_s' - Y_s) + i\,(b_c' - b_c) = \Delta Y_s + i\,\Delta b_c \tag{7}$$

The components of the unknown admittance are accordingly

$$g_u = g_s' - g_s = \Delta g_s \tag{7a}$$

$$b_u = (b_s' - b_s) + (b_c' - b_c) = \triangle b_s + \triangle b_c \tag{7b}$$

If the bridging admittance is merely a conductance, the first term of the right-hand member of (7b) vanishes, and we have

$$b_u = \triangle b_c \tag{7c}$$

This quantity is given directly by the change in tuning capacitance required to re-establish resonance, and by the operating frequency.

The quantity $\triangle Y_s$, which appears in (7), is the one which we are after. It may be resolved by subtracting (4) from (3a). Thus

$$\triangle Y_s = Y_s' - Y_s =$$

$$Y_o \frac{Y \tanh Px + Y_o [1 + \tanh Px \tanh P(l-x)]}{Y \tanh Px \tanh P(l-x) + Y_o [\tanh Px + \tanh P(l-x)]} - \frac{1}{\tanh Pl}$$

By means of some straightforward manipulation, this can be written as

$$\triangle Y_s = Y \left(\frac{\sinh Px}{\sinh Pl} \right)^2 \frac{1}{1 + \left(\dfrac{Y}{Y_o} \right) \dfrac{\sinh Px}{\sinh Pl} \sinh P(l-x)} \tag{8}$$

As a partial check on our results we observe that when $x = 0$, $\triangle Y_s = 0$; and when $x = l$, $\triangle Y_s = Y$. This is reassuring, for it is what our preliminary reasoning required. Now in (8), $\triangle Y_s$, Y, Y_o, and P are, in general all complex quantities of the form $(\triangle g_s + i\triangle b_s)$, $(g + ib)$, $(g_o + ib_o)$ and $(\alpha + i\beta)$, respectively. We notice, however, that if

$$\left(\frac{\sinh Px}{\sinh Pl} \right)^2 \cong \left(\frac{\sin \beta x}{\sin \beta l} \right)^2$$

and

$$\left| \frac{Y}{Y_o} \frac{\sinh Px}{\sinh Pl} \sinh P(l-x) \right| << 1$$

then (8) may be written as

$$\triangle Y_s = Y \left(\frac{\sin \beta x}{\sin \beta l} \right)^2 \tag{8a}$$

If, in addition, $\sin \beta l \cong \beta l$, then a further simplification results:

$$\triangle Y_s = Y \left(\frac{x}{l} \right)^2 \tag{8b}$$

The approximations involved in going from (8) to (8a) are easily permissible provided a reasonably low-loss transmission line is used, and provided the impedance of the bridging resistor is high compared with the characteristic impedance of the line.

In the case of the comparison of two resistors, R_1 and R_2, which are located at two intermediate points on the line, x_1 and x_2, such that the deflection of the vacuum-tube voltmeter is the same, each resistor can be considered the equivalent of the same admittance $\triangle Y_s$ at the sending end.

Hence, from (8a)

$$Y_s = \frac{1}{R_1} \left(\frac{\sin \beta x_1}{\sin \beta l} \right)^2 = \frac{1}{R_2} \left(\frac{\sin \beta x_2}{\sin \beta l} \right)^2$$

or

$$\frac{R_1}{R_2} = \left(\frac{\sin \beta x_1}{\sin \beta x_2} \right)^2 \tag{9}$$

and in the case of a short line,

$$\frac{R_1}{R_2} = \left(\frac{x_1}{x_2} \right)^2 \tag{10}$$

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ERRATA

Mr. R. P. G. Denman of London, England, has kindly called attention to typographical errors appearing in RCA REVIEW, the corrections of which are given below:

RCA REVIEW, Volume II, No. 2, October 1937.

ANALYSIS AND DESIGN OF VIDEO AMPLIFIERS

By S. W. Seeley and C. N. Kimball

The symbol "S" in the first equation under the heading of Appendix II (bottom of Page 182), should be replaced by the symbol "j", indicating an imaginary quantity. The corrected equation reads as follows:

$$Z_L = \frac{\dfrac{R}{j\omega C}}{R + \dfrac{1}{j\omega C}} = \frac{R(1 - j\omega CR)}{1 + R^2C^2\omega^2} = \frac{R}{\sqrt{R^2C^2\omega^2 + 1}}$$

RCA REVIEW, Volume III, No. 3, January 1939.

ANALYSIS AND DESIGN OF VIDEO AMPLIFIERS PART II.

By S. W. Seeley and C. N. Kimball

The equation (at the bottom of Page 304) for slope in the top of the wave should read:

$$\frac{E_c}{E} = \frac{t}{CR} = \frac{1}{120 \times 0.25 \times 0.5} = 6.7 \text{ per cent}$$

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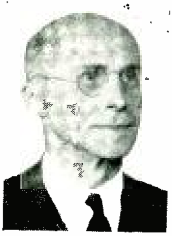
IOURY G. MALOFF, a former lieutenant of the Russian Navy, arrived in this country in 1923 and joined the General Electric Company. In the latter part of 1926 he severed his connection with that company to engage in independent consulting work, and in 1929 became Chief Engineer of the Valley Appliance Corporation at Rochester, N. Y. Later in that year he became Chief Engineer of the Colonial Radio Corporation. In 1931 he joined the RCA Research Staff at Camden where he is now Engineer in Charge of the Television Tubes and Tube Applications Section of RCA Manufacturing Company. Mr. Maloff is a Fellow, Institute of Radio Engineers.



JOHN M. MILLER received the M.A. degree from Yale University in 1907, and the Ph.D. degree from the same university in 1915. From 1907 to 1918 he was employed as Physicist in the Bureau of Standards in Washington. From 1918 to 1925 he served as Radio Aid and Physicist with the Naval Research Laboratory. From 1925 to 1936 he was engaged as radio engineer with the Atwater Kent Manufacturing Company. Since 1936 he has been a member of the Research and Engineering Department of the RCA Manufacturing Company at Harrison, N. J., and since 1937 has been assistant head of the Tube Research Section. Dr. Miller is a member of Sigma Xi and a Fellow of I.R.E.

Albert Preisman received his A.B. and E.E. degrees from Columbia University in 1922 and 1924, respectively, and is now taking graduate courses at Columbia University and the Brooklyn Polytechnic Institute for a Doctor's degree. After working with the Wagner Electric Corporation of St. Louis and the New York Edison Company, he joined the RCA Photophone organization in January, 1929. In 1932, he transferred to RCA Institutes, where he is head of the Audio-Video Frequency Engineering Department.

Bernard Salzberg received his degree of E.E. in 1929 and his master's degree in 1933 from the Polytechnic Institute of Brooklyn. He worked for the Brooklyn Edison Company from 1923 to 1925 and in the Summers of 1926 and 1927. During the Summer of 1928 he worked for the Fada Radio Company. From 1929 to 1931 he was transmitter operating engineer and engineer on research and development in the field of ultra-high frequencies for R.C.A. Communications at Rocky Point, Long Island. Since 1931 he has been working on vacuum tube development problems in the Research and Engineering Department of the RCA Manufacturing Company at Harrison, N. J. Mr. Salzberg is a Member of the I.R.E.

Stuart W. Seeley received his B.Sc. Degree in Electrical Engineering from Michigan State College in 1925. He was an amateur experimenter and commercial radio operator from 1915 to 1924. Following this he joined the experimental research department of the General Electric Company, and a year later became Chief Radio Engineer for the Sparks Withington Company. Since 1935 he has been an engineer in the RCA License Laboratory.

George L. Usselman received his B.S. degree in Electrical Engineering from the Kansas State College in 1916. After his graduation he spent three years in the General Electric Company's test department. In 1920, shortly after the Radio Corporation of America was formed, Mr. Usselman joined its operating division. In 1925 he transferred to the Engineering Department and, engaged in transmitter design and research problems. Mr. Usselman is an associate member of I.R.E. and of A.I.E.E.

TECHNICAL ARTICLES BY RCA ENGINEERS

Published First Quarter, 1939*

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* Does not include January issue of RCA REVIEW.