VOLUME 23 NOVEMBER, 1935 NUMBER 11

PROCEEDINGS
of
The Institute of Radio Engineers

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Institute of Radio Engineers Forthcoming Meetings

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November 18, 19, 20, 1935

CINCINNATI SECTION
November 12, 1935

DETROIT SECTION
November 15, 1935

LOS ANGELES SECTION
November 19, 1935

NEW YORK MEETING
November 6, 1935
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WASHINGTON SECTION
November 11, 1935

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 23 November, 1935 Number 11

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The Institute of Radio Engineers

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AIMS AND OBJECTS. The Institute functions solely to advance the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Among the methods of accomplishing this is the publication of papers, discussions, and communications of interest to the membership.

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INSTITUTE NEWS AND RADIO NOTES

Rochester Fall Meeting

The Rochester Fall Meeting for this year will be held at the Sagamore Hotel on November 18, 19, and 20. Interesting technical programs have been arranged and an exhibition of radio equipment of interest to engineers is scheduled. The program follows:

MONDAY, NOVEMBER 18

9:00 A.M.	Registration Opening of Exhibits
10:00 A.M.	**Technical Session** "Superheterodyne Oscillator Design Consideration," by W. A. Harris, RCA Manufacturing Company, Radiotron Division. "Electrical Quality of Radio Components," by C. J. Franks, Boonton Radio Corporation.
12:30 P.M.	Group Luncheon
2:00 P.M.	**Technical Session** "New Problems in Metal Tubes," by Roger M. Wise, Hygrade Sylvania Corporation. "Latest Developments in Electron Optics" (with demonstration), by W. H. Kohl, Rogers Radio Tubes Company, Ltd.
4:00 P.M.	Inspection of Exhibits Meeting of RMA Committee on Television Meeting of RMA Committee on Sound Equipment
6:30 P.M.	Group Dinner
8:00 P.M.	Joint Technical Session with Radio Club of America "Electron Multipliers and New Electron Technique" (with demonstration), by V. K. Zworykin, RCA Manufacturing Company, Victor Division.

TUESDAY, NOVEMBER 19

9:00 A.M.	Registration Opening of Exhibits
9:30 A.M.	Joint Technical Session with RMA Engineering Division "A Tragedy in Specifications," by L. C. F. Horle, Consulting Engineer. "Management's Stake in Standards," by P. G. Agnew, American Standards Association.
12:30 P.M.	Group Luncheon
2:00 P.M.	**Technical Session** "The Status of the Radio Spectrum," by C. B. Jolliffe, Federal Communications Commission.
3:00 P.M.	Inspection of Exhibits Meeting of RMA General Standards Committee
6:30 P.M.	Stag Banquet (Informal) Toastmaster, A. F. Van Dyck "European Experiences in Radio," by L. M. Clement, RCA Manufacturing Company, Victor Division "Speech with Sound Effects," by David Grimes, Philco Radio and Television Corporation.

WEDNESDAY, NOVEMBER 20

9:00 A.M. Opening of Exhibits

9:30 A.M. **Technical Session**

"Instantaneous Tracing of Tube Characteristics" (with demonstration), by Otto Schade, RCA Manufacturing Company, Radiotron Division.

"Quantitative Influence of Tube and Circuit Properties on Random Electron Noise," by S. W. Seeley and W. A. Barden, RCA License Laboratory.

12:30 P.M. Group Luncheon

2:00 P.M. **Technical Session**

"Design of Doublet Antenna Systems," by H. A. Wheeler, Hazeltine Corporation.

"Iron-Core Antenna Coil Design," by George H. Timmings, Meissner Manufacturing Company.

4:00 P.M. Exhibits Close

Meeting of RMA Committee on Broadcast Receivers

Meeting of RMA Committee on Vacuum Tubes

Meeting of RMA Committee on Facsimile

Advance reservations for hotel accommodations should be addressed to the Hotel Sagamore as early as possible as these facilities are limited.

Committee Work

TECHNICAL COMMITTEE ON ELECTRONICS—I.R.E.

A meeting of the Technical Committee on Electronics operating under the Institute's Standards Committee was held in the Institute office on October 4 and those present were B. E. Shackelford, chairman; F. R. Lack, R. W. Larsen, G. F. Metcalf, O. W. Pike, B. J. Thompson, Dayton Ulrey, P. T. Weeks, and H. P. Westman, secretary.

The meeting was devoted to reviewing the work of the various subcommittees for the purpose of correlating their activities and preparing a schedule of future meetings to complete the work of these groups.

SUBCOMMITTEE ON SMALL HIGH VACUUM TUBES—I.R.E.

The Subcommittee on Small High Vacuum Tubes which operates under the Technical Committee on Electronics met on October 4. P. T. Weeks, chairman; G. F. Metcalf, H. A. Pidgeon, E. W. Schafer, and H. P. Westman, secretary, were in attendance.

The committee reviewed that portion of the 1933 Standards Report devoted to the testing of vacuum tubes and prepared a considerable amount of revision material.

Subcommittee on Definitions—I.R.E.

The Subcommittee on Definitions operating under the Technical Committee on Radio Receivers met in the Institute office on September 19 and those present were F. X. Rettenmeyer, chairman; V. M. Graham, A. V. Loughren, H. Shieve, H. A. Wheeler, and H. P. Westman, secretary.

As this was the first meeting of the committee, it outlined the scope of its activities and the definitions appearing in the 1933 report on receivers were reviewed.

Institute Meetings

Atlanta Section

The May meeting of the Atlanta Section was held on the 2nd in the wirephoto room of the Associated Press in Atlanta. It was presided over by I. H. Gerks and there were fifteen members and guests present.

A "Demonstration of Wirephoto" was presented by E. E. Patton, supervising traffic engineer of the Associated Press. He presented first a short talk outlining the theory and development of the wirephoto system and then explained the operation of the transmitting and receiving equipment. The equipment which is in operation at that office was then inspected and a general discussion in which most of those present participated was held.

Buffalo-Niagara Section

The Buffalo-Niagara Section met on September 18 at the University of Buffalo. L. E. Hayslett, chairman, presided and the attendance was fifty-three.

A paper on "Mechanical Converters" was presented by William W. Garstang, chief engineer of Electronic Laboratories of Indianapolis, Ind. He presented first a brief history of the vibrator type converters. Various circuits associated with these devices were then described and it was his opinion that power packs using the vibrator with rectifier tubes are more practical than the synchronous type with their numerous contact points. Various methods used to eliminate hum and radio-frequency disturbances were explained. Apparatus was available to demonstrate the performance of many different types of commercial vibrators. A cathode-ray oscillograph was used and showed the effect on the wave form of changing circuit constants. Many improvements of the vibrator element have increased its abilities and useful life. A hard tungsten point mounted on a reed with a wiping action makes the best contact. Ingenious schemes to eliminate the mechanical vibrator noise have been developed. The present-day vibrator occupies very little space. In the general discussion which followed, G. C. Crom stated that he believed the self-rectifying synchronous vibrator to be more practical than those used in conjunction with rectifier tubes.

Cincinnati Section

A meeting of the Cincinnati Section was held at the University of Cincinnati on September 17 and was attended by thirty-four. A. F. Knoblaugh, chairman, presided.

A paper on "Automatic Compensation for Bias and Plate Voltage Regulation in Class B Amplifiers" was presented by R. J. Rockwell, engineer in charge of audio developments of the Crosley Radio Corporation. He discussed generally the difficulties in obtaining good regulation of plate and bias voltages for class B stages and outlined the effects of poor regulation. He then described the system which compensates for the regulation and makes unnecessary the use of supply circuits having exceptionally good regulation. He presented data on a typical class B modulator whose distortion at rated power output was nine per cent with rectifier bias, five per cent with battery bias, and only two per cent with the compensator system.

The section voted to coöperate in a proposed Cincinnati Technical Societies Council, the first task of which will be to plan for the publication of a technical calendar to be issued monthly and contain announcements of all technical societies meetings for the month.

Detroit Section

The September 20 meeting of the Detroit Section was held in the Detroit News Conference Room. A. B. Buchanan, chairman, presided and the attendance was forty. Twelve were present at the informal dinner which preceded the meeting.

A symposium of general electronic subjects was presented by six speakers. Walter Hoffman, chief engineer of WWJ, described the new plant and studios being erected by WWJ, the Detroit News. He described the studio arrangement and the transmitter building plan. He then described the three-eighths-wave square tower which is supported on a single insulator and expands similar to an inverted pyramid to the sixty-foot level from which it is of uniform cross section to the top.

Stanley Almas of KLA Radio Corporation discussed the use of parabolic reflectors for sound pickup. He then outlined new methods for controlling feedback in public address work.

Floyd Rousch described the concentric cable transmission line used by WJR in conjunction with their new fifty-kilowatt transmitter. He was followed by Richard Schaeffer who explained the use of electronic tubes in industrial control with particular application to automobile factories. F. C. Denstaedt of the Detroit Police Department then spoke briefly on the possibilities of gas engine driven generators for mobile police and public address use.

The series of papers was ended with one by A. B. Buchanan of the Detroit Edison Company who spoke on "The Patrol of High Tension Lines with Radio Noise Finders for the Prevention of Line Failures." Several types of audio-frequency amplifiers and radio receivers used were described and the methods of locating the source of line noises discussed.

San Francisco Section

The San Francisco Section held a meeting on September 4 at the Bellevue Hotel. R. D. Kirkland, chairman, presided and thirty-eight were present. Fifteen attended the informal dinner which preceded the meeting.

C. W. Hansell, transmitter development engineer of RCA Communications, presented a paper on "Resonant Line Frequency Control and Its Applications." In it he covered the development, design, construction, and commercial application of these stabilizers, illustrating many points with lantern slides and a model. At the conclusion of his paper he demonstrated the remarkable stability of these oscillators with two oscillators operating on a wavelength of one meter using independent power supplies. A general discussion followed and many of those present examined in detail the equipment used in the demonstration.

Washington Section

E. K. Jett, chairman, presided at the September 12 meeting of the Washington Section which was held in the Potomac Electric Power Company Auditorium. Seventy-one attended the meeting and thirty-one were present at the informal dinner which preceded it.

A paper on "The Broadcast Antenna" was presented by T. A. M. Craven. The speaker emphasized the practical applications in broadcasting of the theoretical considerations relative to antennas in general. He stressed the necessity and importance of the coefficient of reflection of the earth in the immediate vicinity of the antenna and stated that he considered it of equal importance in certain circumstances to the height of the antenna and to the ground system. Some practical difficulties in adjusting and designing directional antennas were described and the effect on the electrical characteristics of an antenna due to the presence of another radiating structure in the inductive field was emphasized.

Some practical difficulties encountered in directional antennas were described. It was pointed out that the inability to control the sky waves was not due to sound theoretical considerations but probably to difficulties in adjusting an antenna properly to secure the predicted radiation pattern. His conclusion was that practical experience confirmed the theoretical considerations already outlined by engineers in the PROCEEDINGS and that departures therefrom are due principally to mechanical difficulties as well as neglect of the importance of the coefficient of reflection of the ground near the antenna.

Personal Mention

C. H. Bachman has joined the staff of Hygrade Sylvania Corporation at Emporium, Pa.

E. T. Cahalan has entered the Production Engineering Department of F. W. Sickles Company at Springfield, Mass.

H. N. Coulter, Lieutenant, U.S.N., has been transferred from the Naval Air Station at Sunnyvale to the U.S.S. *Arizona*, basing at San Francisco.

C. C. Eckert is now in the Engineering Department of Aircraft Radio Laboratory at Wright Field, Dayton, Ohio.

B. V. K. French has left RCA License Laboratory to become chief engineer for the Case Electric Corporation at Marion, Ind.

J. M. Glessner has joined the engineering staff of Heintz and Kaufman of South San Francisco, Calif.

Virgil M. Graham has left Stromberg-Carlson Telephone Manufacturing Company to join the engineering staff of Hygrade Sylvania Corporation at Emporium, Pa.

Previously with Electron Research Laboratories, J. M. Hollywood has joined the engineering staff of Ken-Rad Tube Corporation, Owensboro, Ky.

Previously with Stupakoff Laboratories, A. P. Huchberger has become president of International Engineers of New York City.

Formerly with Humble Oil and Refining Company, J. F. Imle has joined the Lago Petroleum Corporation at Maracaibo, Venezuela.

Donald McNicol, past president of the Institute has presented a nucleus of a communication museum to be housed in the National Research Building at Ottawa, Ontario, Canada. The collection includes numerous rare items many of which were used during the opening years of the Twentieth Century. In arranging for its housing at Ottawa, Colonel Steel of the Canadian Broadcast Commission coöperated with Mr. McNicol.

Proceedings of the Institute of Radio Engineers
Volume 23, Number 11 *November, 1935*

TECHNICAL PAPERS

AN UNATTENDED ULTRA-SHORT-WAVE RADIOTELEPHONE SYSTEM*

By
N. F. Schlaack and F. A. Polkinghorn
(Bell Telephone Laboratories, Inc., New York City)

Summary—*Some of the factors involved in the application of an ultra-high-frequency radio link to wire telephone plant are discussed. A description is given of an ultra-high-frequency radio circuit which has been set up and operated between Green Harbor and Provincetown, Mass. This circuit is used as a part of a regular toll telephone circuit between Boston and Provincetown. It has been found that the equipment can remain in operation over considerable periods without attention or adjustment.*

FOR several years attention has been directed by Bell Telephone Laboratories toward determining the characteristics of ultra-high frequencies and their possible application to the telephone plant. The results of some of the more fundamental studies have been published.[1] They led to the belief that ultra-high-frequency radio might find a useful field as an adjunct to the wire telephone plant in crossing natural barriers where other means might prove difficult or expensive.

It was found that ultra-high frequencies have a comparatively short useful range since they are not reflected from the Kennelly-Heaviside layer and are attenuated according to the inverse square of the distance. From an engineering standpoint the principal advantages of the use of ultra-high frequencies lie in the fact that atmospheric disturbances are much less than at lower frequencies and antennas of a much higher gain can be built in a given space. As far as transmission is concerned, the relation between transmitted and received energy in a short-range ultra-high-frequency radio system using half-wave antennas is practically independent of frequency. The ease and

* Decimal classification: R423.5×R450. Original manuscript received by the Institute, May 3, 1935. Presented before Tenth Annual Convention, Detroit, Mich., July 1, 1935.

[1] C. R. Englund, A. B. Crawford, and W. W. Mumford, "Some results of a study of ultra-short-wave transmission phenomena," Proc. I.R.E. vol. 21, pp. 464–492; March, (1933); *Bell Sys. Tech. Jour.*, April, (1933).

J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," Proc. I.R.E., vol. 21, pp. 427–463; March, (1933); *Bell Sys. Tech. Jour.*, April, (1933).

C. B. Feldman, "Optical behavior of the ground for short radio waves," Proc. I.R.E., vol. 21, pp. 764–801 June, (1933).

efficiency of the production of energy, however, decreases quite rapidly as the frequency increases, particularly when crystal controlled transmitters are considered.

In order to determine some of the possibilities and capabilities of an ultra-high-frequency radiotelephone system a channel was set up between the experimental stations of Bell Telephone Laboratories located at Deal and Holmdel, N. J. This channel was operated on an experimental basis for nearly two years and much useful engineering and operating information was obtained.

In order to make the new facility practicable for use in as many as possible of the situations for which it is technically adapted, it is necessary to keep the total operating costs low. With ordinary radio equipment arrangements the costs of continuous attendance at both terminals of the circuit would considerably exceed all of the other operating charges combined. By designing the equipment to include certain features over and above the basic requirements, it is possible to reduce to a minimum the attendance necessary to assure continuous operation. By using equipment capable of continuous operation out of doors, considerable additional economies are effected.

The operation of a radio system as a part of a toll telephone circuit involves considerations not found in other types of operation.

With most radio systems now in use, including broadcasting, it is desirable to regulate the volume at the input to the radio transmitter in order to obtain the greatest modulated ouput. This is necessary primarily to override radio noise. When the volume is regulated in a system which may be connected to two-wire circuits, it is usually necessary to add voice-operated switching devices to prevent echoes and singing around the circuit.[2] Ultra-high-frequency radio circuits, however, are normally quite stable and comparatively free from noise. It is possible, therefore, to omit volume regulation and voice-operated devices at a considerable saving in cost. It is necessary under this condition to provide a radio transmitter of somewhat higher power capacity than would be required if volume regulation were used but the cost of this additional power is small compared to the cost of the features required to provide for "regulated volume operation."

In order to form a two-wire two-way circuit which can be switched to other circuits in the telephone plant the two unilateral radio circuits operating in opposite directions are brought together in hybrid coils like the oppositely bound paths of an ordinary four-wire circuit. The usual compromise networks are provided to balance the connect-

[2] S. B. Wright and D. Mitchell, "Two-way radiotelephone circuits," Proc. I.R.E., vol. 20, pp. 1117–1130; July, (1932); *Bell Sys. Tech. Jour.*, July, (1932).

ing circuits. The problem of preventing the possibility of singing is similar to that encountered in setting up wire circuits, namely, that losses around any path exceed the gains by an adequate margin under all conditions.

When a radio transmitter and receiver are located near each other, as it is desirable to do for the application under consideration, some induced voltage from the local transmitter is impressed on the local receiver. In a single circuit system this would appear as an echo and, in general, need not be reduced to an extremely low value. The maximum value of echo or side tone to be tolerated arising from this source depends primarily upon the delay and the losses encountered in the connecting circuits. The factors which control the amount of echo are, of course, the physical spacing between transmitter and receiver and the frequency separation between the received and transmitted signal.

If unattended operation is to be obtained it is desirable that starting and stopping both the transmitter and the receiver be separately controlled from the telephone office. It is also desirable that some sort of local testing arrangement be provided to allow the test board operator to determine whether the transmitter and receiver are operating properly.

In order to obtain representative information on the feasibility of operating an ultra-high-frequency telephone circuit on an unattended basis in the telephone plant and to secure a better idea of the mechanical and electrical requirements, equipment was constructed by Bell Telephone Laboratories for such an installation. Conventional types of tubes, designed for use at frequencies below thirty megacycles were employed, and it was found, on this account, that the practical frequency limit was about 65 megacycles.

With the coöperation of the New England Telephone and Telegraph Company this equipment was used to establish an experimental ultra-short-wave circuit between Green Harbor and Provincetown, Mass., as indicated on Fig. 1. The physical conditions are favorable for an ultra-short wave link between these two points which are about twenty-four miles apart. Sand dunes near Provincetown, rising about eighty feet in height, make it possible to secure an optical path across the bay. The radio circuit is extended by wire from Green Harbor to Boston to form a direct Boston-Provincetown toll circuit. It is used as one of a group of terminal circuits and is operated at the normal over-all net loss for this type of circuit, namely, nine decibels.

At Boston and at Provincetown the circuit appears at a jack in the switchboard beside the jacks of wire toll circuits. As far as the operator is concerned, switching and ringing operations are performed in the same manner as for other similar grade toll circuits and there is nothing to designate that this toll circuit has a radio link. The insertion of a cord into the jack starts the radio transmitter at that end of the circuit. The receivers at both ends are kept in constant operation while the circuit is available for traffic. Ringing is accomplished by sending a 1000-cycle tone interrupted at twenty cycles over the circuit. Since the radio transmitter requires less than one second to start, the

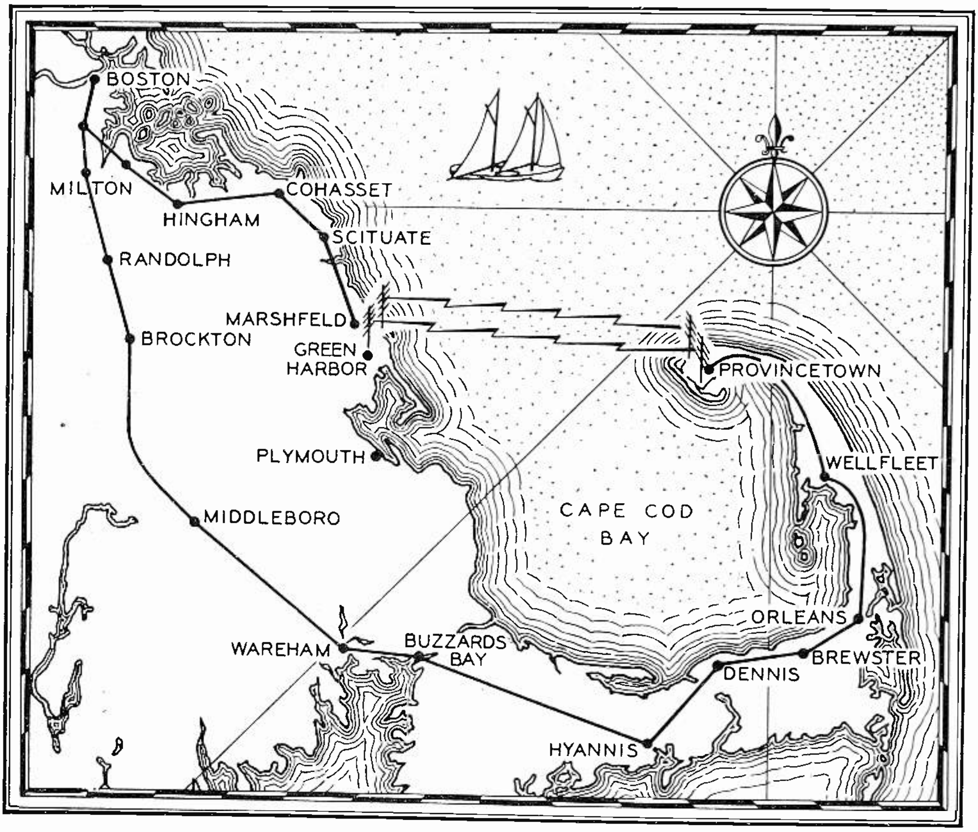

Fig. 1

operator may ring practically immediately after inserting the cord. Privacy equipment similar to that used on the transatlantic short-wave radio channels, is installed at the terminal offices.

The receivers are started and stopped by the operation of a key at the local test board. The power supply is arranged so that when the receiver is in operation, power is also supplied to some of the filaments of the transmitter. Provision is made for testing the over-all operation of the local transmitter and receiver by the manipulation of a key at the test board at each end. Throwing a key causes a tone to be generated at the transmitter by making the first audio stage oscillate.

This modulates the transmitter. An auxiliary oscillator in the receiver is also started. This causes the signal from the transmitter to fall in the band of the local receiver. If both transmitter and receiver are operating properly, the tone will be heard by the test board operator.

Transmission from Green Harbor to Provincetown is accomplished on a frequency of 65 megacycles and in the reverse direction on 63 megacycles. This does not represent the minimum possible frequency spacing for this equipment, but was a convenient one for the experiment.

The transmitters are crystal controlled and are capable of delivering fifteen watts of carrier power which can be completely modulated. It was estimated that this would give a reasonably satisfactory circuit. Crystals of the "BT" type[3] having a low temperature-frequency

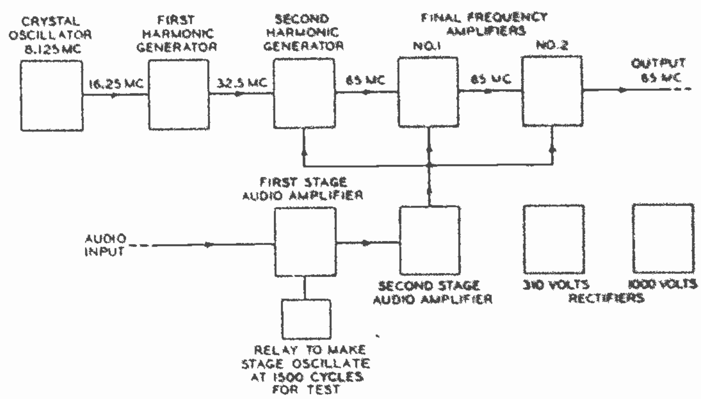

Fig. 2—Block schematic of ultra-short-wave transmitter.

coefficient were used in airplane type temperature controlled ovens in order to maintain the signal within the band of the receiver.

A block schematic of the Green Harbor transmitter is shown in Fig. 2. The Provincetown transmitter is of similar construction. The quartz crystal oscillator, in which the second harmonic is selected, is followed by two harmonic generators, a push-pull modulating amplifier, and a push-pull power amplifier. Modulation is accomplished by supplying audio-frequency power to the plate and screen of the modulating amplifier and to the screens of the second harmonic generator and the power amplifier. The audio-frequency power for modulation is obtained from a two-stage transformer coupled amplifier. Small receiving type pentodes (Western Electric No. 294A) are used in each socket except for the power amplifier stage and the last audio stage which use respectively two 75-watt tetrodes (Western Electric No. 282B) and two low power triodes (Western Electric No. 275A) in

[3] F. R. Lack, G. W. Willard, and I. F. Fair, "Some improvements in quartz crystal circuit elements," *Bell Sys. Tech. Jour.*, July, (1934).

push-pull. The audio gain provided in the transmitter makes it possible to obtain complete modulation with an audio input of seventeen decibels below a milliwatt in six-hundred ohms. Two rectifiers employing hot-cathode mercury-vapor tubes (Western Electric No. 301A)

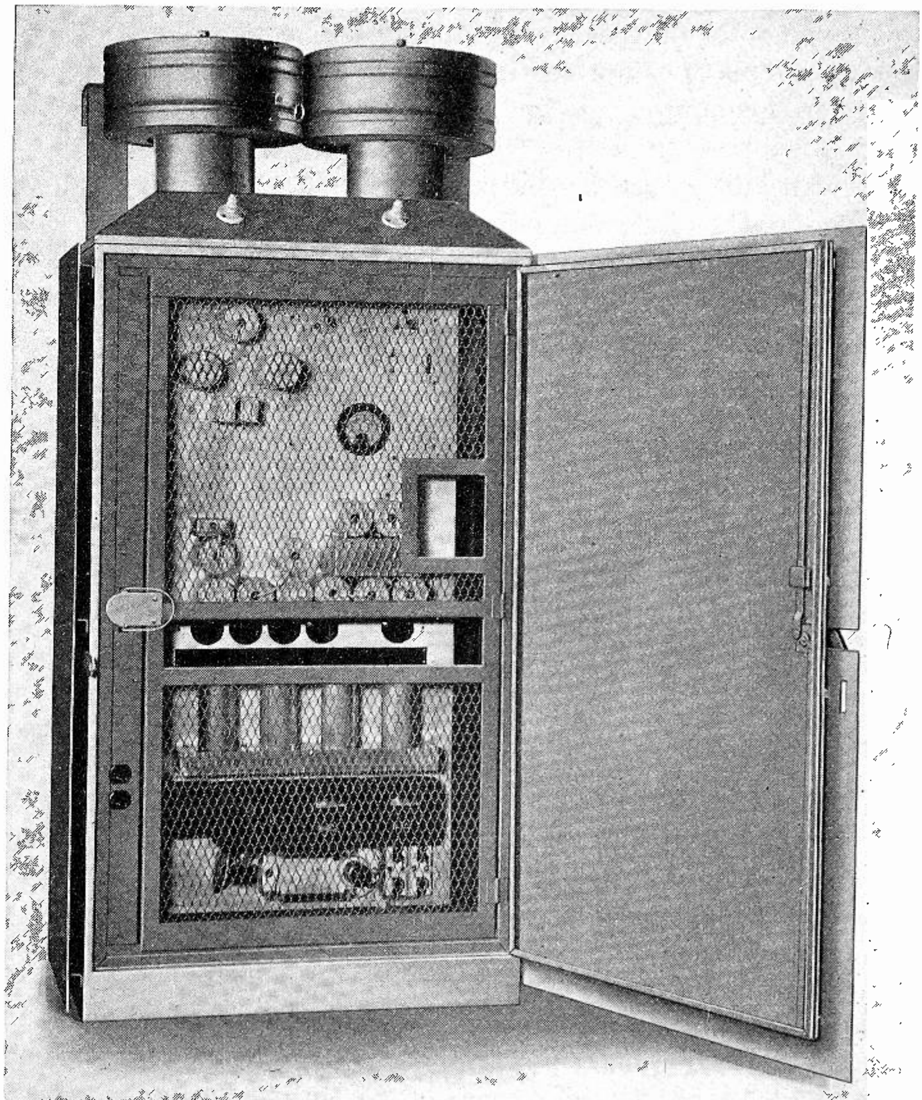

Fig. 3—Ultra-short-wave transmitter mounted in metal container suitable for pole mounting.

supply plate and screen potential for all tubes. Grid-bias potentials are obtained from cathode resistors, and grid leaks. The grid and plate circuits of each stage are shielded from each other to prevent extraneous coupling and interstage feedback. The transmitter operates entirely on standard 110-volt 60-cycle supply.

The receivers also operate from a 110-volt 60-cycle circuit. They are of the double detection type. A block schematic is shown in Fig. 4. To make unattended operation possible and at the same time permit high selectivity at these frequencies, a crystal oscillator is used as the source of beating frequency. The second harmonic of the crystal is selected in the oscillator circuit and a single stage harmonic generator produces sufficient voltage of the eighth harmonic for satisfactory operation of the detector. The intermediate-frequency amplifier consists of three stages of amplification at 1600 kilocycles and has a band width of approximately 50 kilocycles. A small amount of automatic volume control is provided to compensate for slight variations in re-

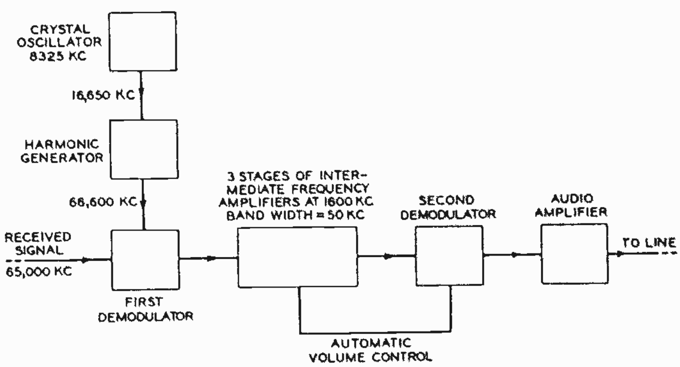

Fig. 4—Block schematic of ultra-short-wave receiver.

ceived voltage caused by variation in humidity and other factors. The receivers are capable of delivering 0.3 watt of undistorted power to a six hundred-ohm impedance. This is well in excess of the power required during normal operation.

The transmitting and receiving antennas are identical and each is mounted on a single wooden pole extending about ninety feet above the ground. The two poles at each end of the circuit are spaced fifty feet apart on a line at right angles to the direction of transmission. This spacing was determined to some extent by the side tone requirements and the frequency spacing. The spacing chosen is only a practical solution for the condition involved and should not be taken as a minimum value. Horizontal exciter and reflector elements are supported on standard crossarms. Four pairs of half-wave exciter elements, each comprising two half-wave conductors, are spaced one-half wavelength apart in a vertical plane on one side of the pole. Four pairs of half-wave reflector elements are similarly arranged on the opposite side of the pole. The spacing between exciters and reflectors is one-quarter wavelength. These antennas when mounted as shown in Fig. 6, give a gain, measured at the other end of circuit, of twelve decibels over a simple half-wave element with the same power input, at the same mean height. This type of antenna was used as it gave the highest gain and directivity which could be conveniently mounted on a single pole. High directivity was desirable not only as a means of increasing the re-

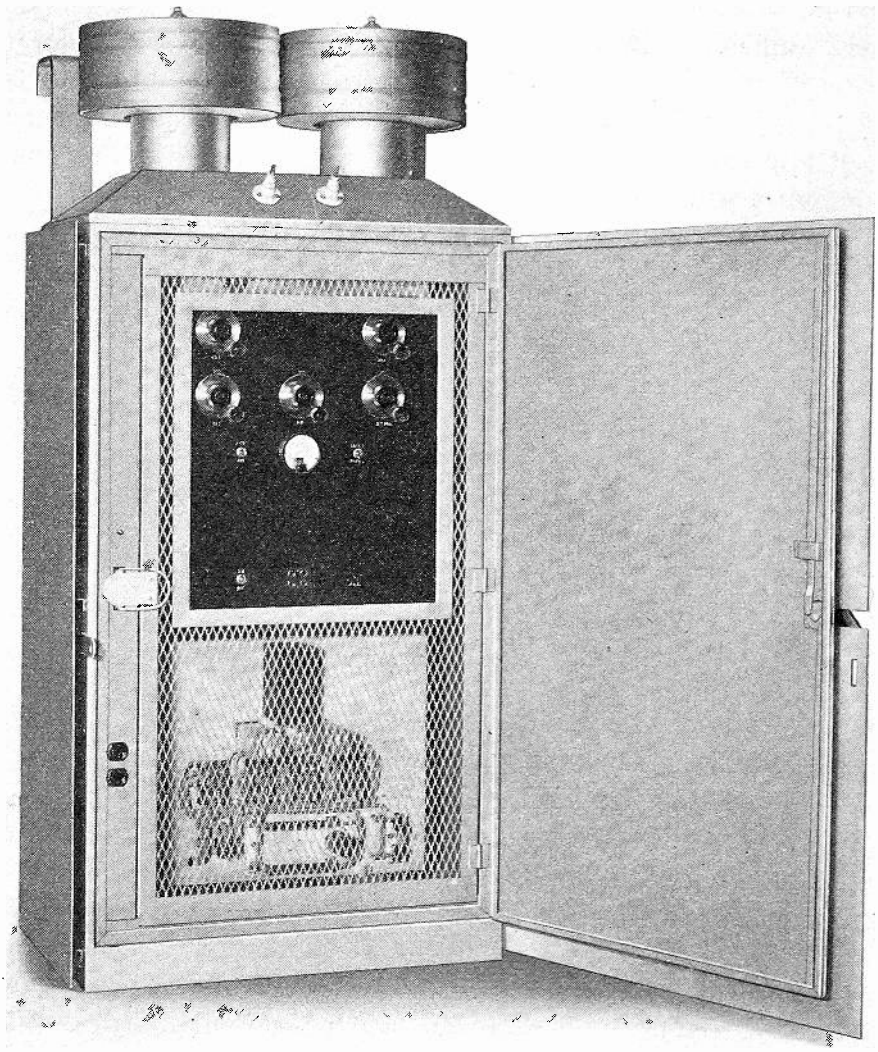

Fig. 5—Ultra-short-wave receiver mounted in metal container suitable for pole mounting.

ceived signal but also to exclude automobile ignition and other noises originating near the receiving stations.

The transmitters and receivers are mounted on the poles with their respective antennas. They are connected to the antennas by means of open-wire transmission lines. A special metal container which can be used for either a transmitter or a receiver was designed in order that the equipment might be mounted in this manner. Attention was paid in the construction of the metal containers to the protection of maintenance personnel. Access to the inside of the transmitter cannot be obtained until all high voltages have been removed. The high voltage

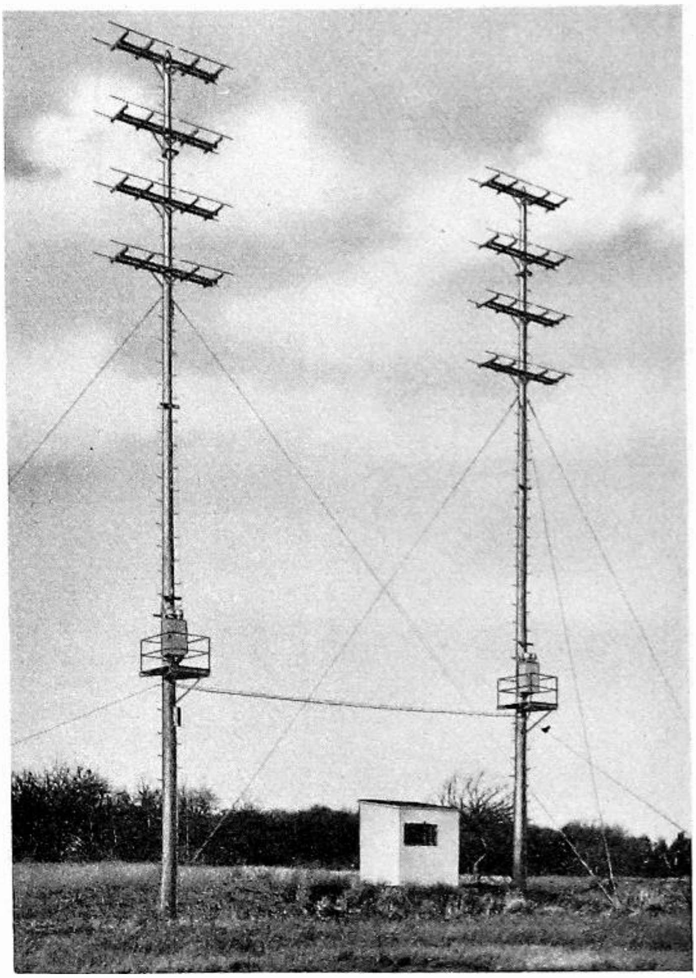

Fig. 6—General view of antennas and pole-mounted radio equipment at Green Harbor terminal.

cannot be reapplied until the metal screen door, see Fig. 7, on the front of the transmitter has been closed. This is accomplished by mounting a four-terminal plug on the screen door which breaks the power supply circuits to the transmitter.

In order to allow the equipment to be inspected and serviced with ease, the entire transmitter or receiver is mounted on a metal panel arranged so that it may be swung out as shown in Fig. 7. Two outlets for 110 volts alternating current have been installed in the left-hand door jamb to provide a convenient place for connecting soldering irons, etc.

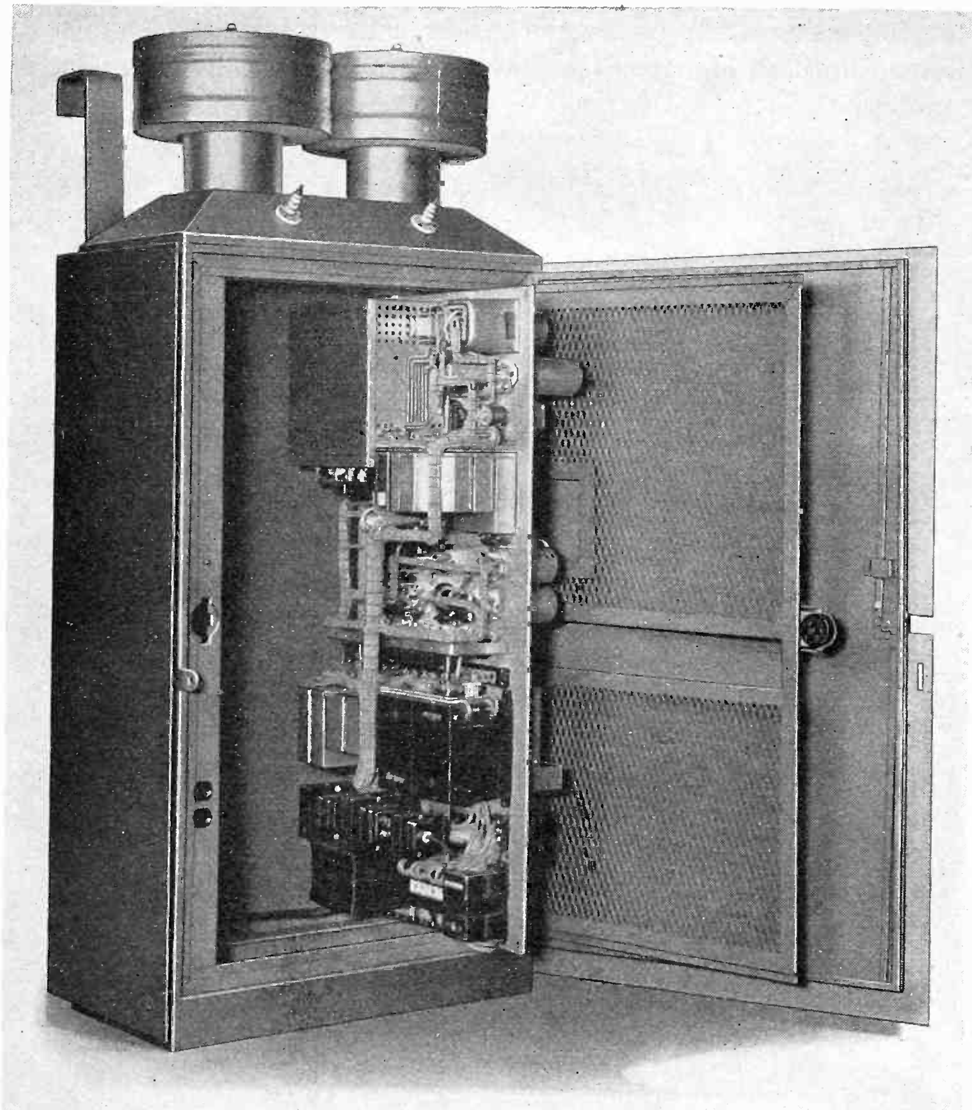

Fig. 7—Open view of ultra-short-wave transmitter.

The equipment was installed at Green Harbor and Provincetown, Mass., during the month of June, 1934. Signal-to-noise ratios of about 55 decibels, were realized.[4] The power loss over the radio path, meas-

[4] The signal referred to is that amplitude of 1000-cycle tone which produces complete modulation of the transmitter. Both the signal and the noise are measured through a weighting network which offers no attenuation to 1000 cycles but attenuates other frequencies in accordance with their relative interfering effects. If the noise is assumed to be uniformly distributed over the voice band from 250 to 2750 cycles, the total computed noise energy passed by the network would be eight decibels less than if no network were used.

ured from transmission line to transmission line, is approximately 100 decibels. This is substantial agreement with the level terrain formula derived by C. R. Burrows, A. Decino, and L. E. Hunt from transmission tests over flat land.[5]

The circuit has been available for traffic almost continuously since installation, except for periods when tests were being conducted or minor modifications were being made. Daily observations of the circuit loss have not shown variations greater than ± 4 decibels from the normal value. Noise from local thunderstorms has never prevented the circuit from being utilized in the normal manner. The several months of traffic operation to which the circuit has been subjected have disclosed no important technical difficulties with this type of system. It has been found that the radio apparatus can remain in operation over periods of several weeks without attention or adjustment.

[5] Chas. R. Burrows, Alfred Decino, and Loyd E. Hunt, "Ultra-short-wave propagation over land," Proc. I.R.E., to be published, December, 1935.

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RADIO-FREQUENCY DISTRIBUTING SYSTEMS*

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During the past few years considerable thought has been given to brodcast receiving antenna systems which combine construction simplicity with improved performance. It has been current practice in the past to erect a separate antenna for each radio receiver. This has resulted in considerable annoyances and extremely poor performance in crowded urban and suburban districts due to the proximity and consequent coupling of antennas on apartment house roofs. It is not unusual in such instances for the power output of a given receiver to vary as much as ten decibels due to the tuning of a receiver on an adjacent antenna.

Recent developments in radio-frequency distribution systems permit of the simultaneous operation of several thousand broadcast radio receivers from a single antenna without interaction between receivers. Moreover these systems permit of the use of an antenna located several hundred feet from the connected receivers. Consequently an antenna may be erected in a location relatively free from interference and the broadcast signals may be delivered to the connected receivers through the interference infested areas without undue signal loss and with almost complete freedom from local interference. The location of the antenna in a position relatively remote from interference generating and interference conducting devices is a matter of the upmost importance in the reception of noise free broadcast programs.

An adequately shielded receiver is in itself almost immune from the effects of locally generated interference. Moreover, the electric field set up by interference generating or distributing devices is largely confined to a small area in the immediate vicinity of these devices. This is due to the fact that metal lath, building frames, etc., in the immediate vicinity tend to localize the interference field and to the fact that the electric field near a radiator (i.e., within a fraction of a wavelength) is inversely proportional to the cube of the distance from the radiator.[1] The complete elimination of radio interference is im-

* Decimal classification: R320. Original manuscript received by the Institute, October 5, 1934.

[1] G. H. Browning, "Reducing man-made static," *Electronics*, p. 366, December, (1932).
J. A. Ratcliffe, L. G. Vedy, and A. F. Wilkins, "The spreading of electromagnetic waves from a Hertzian dipole," *Jour. I.E.E.* (London), vol. 70, no. 425, p. 522; May, (1932).

practicable, if not impossible, if the use of electricity for light, power, and communication is to be retained. Radio-frequency distribution systems would therefore appear to offer the best as well as the most economical solution of satisfactory broadcast reception in urban and suburban areas. Indeed certain municipalities[2] have passed ordinances aimed at the reduction of local interference which require the complainant to provide a suitable antenna system before his complaint will be recognized.

A radio-frequency distribution system consists essentially of an antenna, a distribution medium, and suitable coupling devices for each connected radio receiver. If the system loss from antenna to connected receiver is of appreciable magnitude, amplifiers to compensate such losses may be required. The most practical distribution medium is a radio-frequency transmission line properly terminated in its characteristic impedance to prevent reflections. Practical transmission lines have a characteristic impedance of 50 to 100 ohms except when loaded to a higher impedance with properly spaced inductances.[3] A standard antenna has a capacity reactance of about 1000 ohms at 1000 kilocycles. Consequently unloaded lines require a step-down transformer between the antenna and transmission line if large reflection losses are to be avoided. From the well-known formulas.

$$f_c = \frac{1}{\pi\sqrt{LC}}$$

and,

$$Z = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$

it appears that a transmission line having a distributed capacity of 20

Ellis Van Alta and E. L. White, "Radio interference from line insulators," *Trans. A.I.E.E.*, vol. 49, p. 1; January, (1930).

C. Manneback, "Radiation from transmission lines," *Trans. A.I.E.E.*, vol. 42, p. 289; February, (1923).

[2] "Interference responsibility shared by radio user," *Elec. World*, p. 738; April 23, (1932).

[3] M. I. Pupin, "Propagation of long electrical waves," *Trans. A.I.E.E.*, vol. 16, p. 93, (1899).

M. I. Pupin, "Wave transmission over non-uniform cables and long distance air lines," *Trans. A.I.E.E.*, vol. 17, p. 445, (1900).

G. A. Campbell, "On loaded lines in telephone transmission," *Phil. Mag.*, p. 313, March, (1903).

B. Gherardi, "Commercial loading of telephone circuits in the Bell System," *Trans. A.I.E.E.*, vol. 30, p. 1743, (1911).

T. Shaw and W. Fondiller, "Development and application of loading for telephone circuits," *Bell Sys. Tech. Jour.*, p. 221, April, (1926).

E. H. Colpitts and O. B. Blackwell, "Carrier current telephony and telegraphy," *Trans. A.I.E.E.*, vol. 40, p. 205, (1921).

micromicrofarads per foot may be loaded to an impedance of 1000 ohms by 200-microhenry coils spaced at 10-foot intervals. The cut-off frequency of such a line would be about 1600 kilocycles. If an unbalanced line of this type terminated in its characteristic impedance of 1000 ohms were connected directly to a 200-micromicrofarad antenna some 70 per cent of the induced antenna voltage at 1000 kilocycles would appear across the input terminals of the line.[4] The same voltage would appear across the input terminals of a step-down transformer feeding an unloaded low impedance line if the transformer ratio were such as to step up the line impedance to 1000 ohms. While higher step-up ratios are possible, an impedance ratio of 50 or 100 ohms to 1000 ohms appears to be a good compromise between performance and cost for such coils. In general, repeating coils intended to couple the transmission line to the antenna have close coupling (of the order of 90 per cent or more) between primary and secondary. However, loose coupling may be used to advantage in some cases such as between an antenna and a loaded transmission line. For example, if a balanced line is employed a coil will be required between the antenna ground system and the line to preserve the line balance. If the line is loaded to an impedance less than 1000 ohms the coil may be desirable to reduce reflection losses. The leakage reactance of the transformer may, if properly proportioned, be used as the first set of loading coils in the line. Also the leakage reactance might conceivably be used to some advantage in loading the antenna in special installations in which antenna capacity is known.

There are two practicable types of transmission line[5] for this type of service which effectively avoid interference pickup. The first is the balanced and transposed line, the chief virtue of which lies in the fact that longitudinal currents of identical phase and magnitude will be induced in each leg of the line by local interference fields and will be balanced out if the line balance is sufficient. The second is the coaxial conductor line, which, since it has no external field, can be made practically immune from interference pickup. This line consists of an

[4] F. M. Colebrook, "An experimental and analytical investigation of earthed receiving aerials, *Jour. I.E.E.* (London), vol. 71, no. 427, p. 235, (1932).
E. B. Moullin, "Radio-Frequency Measurements," J. B. Lippincott Co., p. 413.

[5] S. Pero Mead, "Wave propagation over parallel tubular conductors," *Bell Sys. Tech. Jour.*, p. 327, April, (1925).
E. J. Sterba and C. B. Feldman, "Transmission lines for short-wave radio systems," *Bell Sys. Tech. Jour.*, p. 411, July, (1932).
W. L. Everett and J. F. Byrne, "Single wire transmission lines for short-wave antennas," Bulletin No. 52, Engineering Experiment Station, Ohio State University.
T. McLean, "Transmission line feed for short-wave antennas, *QST*, p. 25 October, (1932).

inner conductor which lies along the axis of a continuous cylindrical shield. The usefulness of the former depends almost entirely upon line balance, whereas, the usefulness of the latter depends principally on the effectiveness of the outer (shield) conductor.

It is very important that a transmission line have no large impedance irregularities and that it be properly terminated at its characteristic impedance. Otherwise reflections or standing waves may result which cause excessive attenuation at certain frequencies or at certain points along the line. It frequently happens that such attenuation irregularities are not apparent when the voltage is measured across the termination resistance only. Since receivers are usually fed from intermediate points on the line as well as from the termination end, it is important that no standing waves of appreciable magnitude exist. The correct termination resistance is of course the geometric mean of the impedances measured at the sending end of the line with the receiving end open-circuited and short-circuited. In loaded lines an irregularity may be introduced by a loading coil of the wrong inductance such as might occur from shorted turns or by the wrong capacity between loading coils due to improper coil spacing or high capacity from a coil to ground. Consequently, it is essential that some care be taken in the design of loading coils and their housings to prevent the possibility of high capacity from coil to ground or the shorting of part of the coil through careless installation.

The loss of loaded lines is generally determined by the resistance of the loading coils particularly at the high-frequency end of the spectrum. Since the resistance of the coil increases with frequency, the line loss increases in direct proportion as given by the following formula[6] for attenuation:

$$A = \frac{R}{2}\sqrt{\frac{C}{L}}$$

which holds for lines in which ωL is large compared to R and the shunt conductance is negligibly small. The losses in loaded lines employing loading coils of moderate size and cost is of the order of 0.1 to 0.25 decibel per section.

The balance of ordinary twisted pair may be materially altered during installation, with the result that such a line is more susceptible to local interference. For this reason it is more or less common practice to shield all transmission lines employed in radio-frequency distribution systems. The sheath in addition to its shielding action also serves

[6] K. S. Johnson, "Transmission Circuits for Telephonic Communication," D. Van Nostrand Co., p. 147.

to maintain the original configuration of the lines and to insure a uniform capacity from each conductor to ground. An ideal shield has infinite conductivity and is electrically continuous. A lead sheath of the thickness usually used is therefore a better shield than a thin copper braid. However, a heavy copper braid or spiral tape of untinned copper can be made more effective than a lead sheath. Tinned copper braid due to its higher resistance is generally less effective than untinned copper braid as a shield. In coaxial conductor lines an outer sheath of low conductivity may be of advantage since signal currents are confined to the inner surface of the outer conductor and the interference is confined to the outer surface of the sheath. If the sheath is well grounded along its length, the interference current along the sheath may be considerably reduced, thereby reducing pickup.

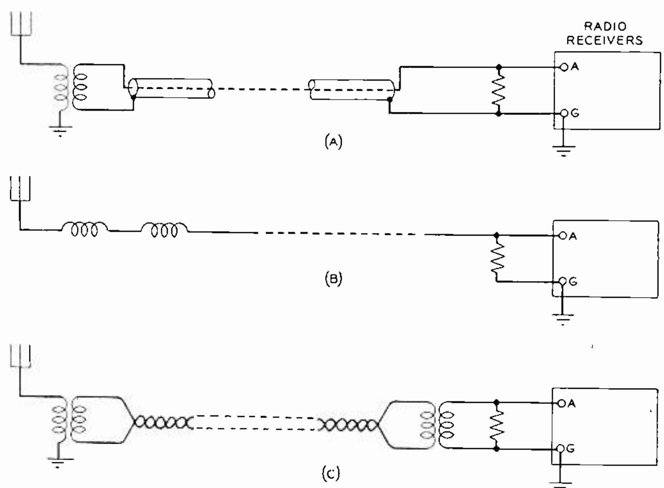

Fig. 1—Simplified passive radio-frequency distribution system.

It is obvious that several types of systems offer possibilities. Thus a simple system consisting of an antenna, repeating coil, and unbalanced transmission line might be used to feed a single receiver connected across the termination impedance as shown on Fig. 1 (A). Or a loaded unbalanced transmission line connected directly to the antenna and feeding a single receiver as in Fig. 1 (B) might be employed. Or an unloaded balanced transmission line with repeating coils at each end might be employed as in Fig. 1 (C). Obviously the system of Fig. 1 (B) might be modified to employ a balanced loaded line fed from an antenna and counterpoise, a doublet or other balanced structure, and coupled to the receiver and terminating resistance through a repeating coil.

If more than a single receiver is to be supplied from the same transmission line, coupling elements (either passive or active) must be employed to prevent receiver interaction. Since radio receivers differ widely with respect to input circuits it is essential to consider the state of affairs existing when the connected receiver load consists of equipments whose input impedance varies considerably within the tuning interval. The input impedance of certain types of receivers may vary from a few ohms to a maximum of about 1000 ohms during tuning. Moreover it may be that all receivers are at times tuned to the same frequency. If satisfactory operation is to result it is essential that the interaction between receivers be negligibly small. Two problems are involved; namely, change in signal level due to tuning, and the effect on the tuning of the first tuned circuit of a previously adjusted receiver when the tuning of any or all of the remaining receivers is changed. Obviously, a complete solution of either of these problems will involve the constants of the coupling device, the constants of the transmission line, and the constants of the input circuits of the receiver. A detailed analysis of the action of coupling networks is beyond the scope of this paper. However, a few pertinent facts are evident from a cursory inspection and these will be discussed. Those receivers employing low inductance primaries (of the order of 10 to 20 microhenrys) loosely coupled to a tuned secondary are probably of most practical interest, since they present one of the most serious problems, and are still relatively numerous. Greatest interaction between receivers will obviously occur when all receivers are connected across the line at one point, say across the termination.

For the conditions outlined above it is obvious that when a large series resistance is employed to couple the receiver to the line, the voltage transferred to the first tuned circuit, for a constant line voltage, will be nearly proportional to frequency. When a small fixed condenser is substituted for the resistance the voltage transfer will vary approximately as the square of the frequency. With a series inductance, on the other hand, the voltage transfer will be essentially uniform with frequency. Numerous combinations of resistance, inductance, and capacity can be used to effect a uniform voltage transfer; however, such networks appear to have no particular advantage over the simple series inductance for the case assumed, which is probably representative of the greatest number of low input impedance receivers. The coupling impedance must, of course, be high compared to the line termination (say twenty to thirty times the termination resistance). The required magnitude of such impedances is directly proportional to the number of connected receivers. The detuning effect, which may be defined as the difference in frequency of maximum secondary current when the receiver is connected through the coupling elements to the line, and resonance frequency, when the receiver is disconnected from the line, may be shown to be directly proportional to the square of the mutual impedance between primary and secondary of the input circuit and inversely proportional to the series coupling impedance. The voltage step-up from line to first tuned circuit is of course directly proportional to the mutual and inversely proportional to the coupling impedance. Obviously the detuning effect is the limiting factor in the determination of the coupling inductance. The loss in step-up of a given receiver due to the coupling inductance with ten connected receivers of the low impedance type is of the order of ten decibels for a detuning effect of about one kilocycle under conditions likely to be obtained in practice. Since the loss from antenna to receiver will be proportional to the number of connected receivers, systems employing such networks are limited to about ten or fifteen connected receivers.

A loosely coupled transformer may also be employed as a coupling element. The same general reasoning as that applied to the case above applies here. Fig. 2 illustrates the relative magnitudes of the leakage compared to the primary and secondary self-inductances.[7] Obviously

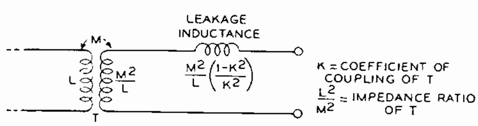

Fig. 2—Equivalent circuit constants of coupling transformer.

a voltage step-up cannot reduce the effective coupling loss since for a given amount of detuning the leakage must be increased in proportion to the step-up. A practical case might involve a coefficient of coupling of 25 to 30 per cent with a leakage reactance of the order of one millihenry and mutual impedance equal to the secondary self-impedance, to work from a 100-ohm line. Since the number of connected receivers in systems employing passive coupling elements of the above type are limited to about ten to fifteen, the increase in line loss due to the connected receivers is negligible. Moreover it is evident that certain types of modern receivers having a high input impedance loosely coupled to the tuned circuits might be connected directly to the low impedance transmission line without the use of coupling elements. The chief disadvantage of a practical system employing passive coupling elements lies in the low loss (about 20 decibels) between any two connected receivers, and in the small number of receivers that can be served by a single antenna. Thus an oscillating receiver or certain

[7] W. J. Creamer, "Autotransformer circuit analysis," *Radio Engineering*, vol. XIV, no. 7, p. 18; July, (1934).

types of double detection receivers may feed sufficient energy back into the transmission line to spoil reception on other receivers.

This state of affairs can of course be remedied by the use of higher loss coupling elements. Since the average receiver must receive signals over a range of intensity of about 60 decibels, there should be an equivalent amount of isolation between any two connected receivers. Probably the most practical method of accomplishing this result is through the use of resistance pads. If L-type pads of high loss are connected across the line each with a loss of 25 to 30 decibels, it is obvious that the necessary isolation may be obtained. It will be necessary of course to compensate for such losses by means of amplifiers. If a large number of such pads are connected across the line, it is obvious that

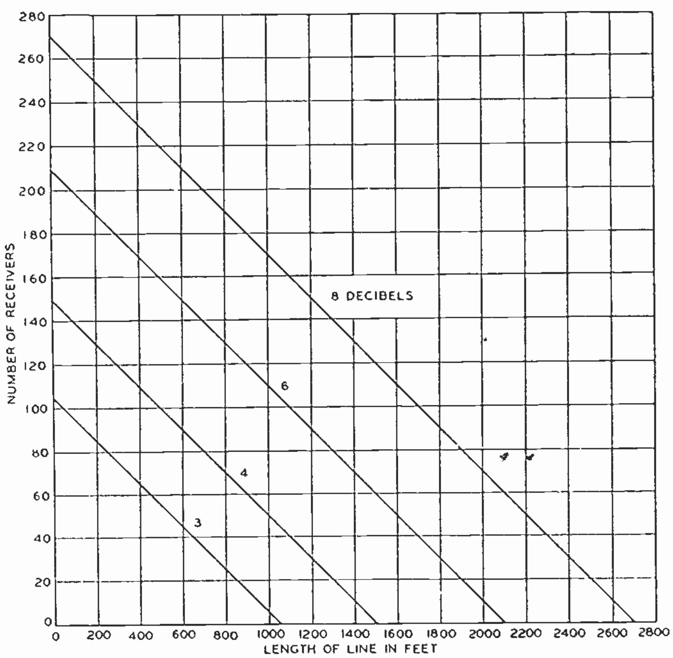

Fig. 3—Plot of $N=300A-10X$ for Western Electric No. 700 cable.

the line loss between the first and the last connected receiver will be increased. As a practical case assume that 10,000-ohm pads are to be uniformly distributed along an unbalanced 65-ohm line having a loss of 0.3 decibel per hundred feet. The attenuation constant for such a line is given by the expression[6]

$$A = \left(\frac{R}{2} + \frac{LG}{2C}\right)\sqrt{\frac{C}{L}}.$$

This assumes that the series inductive reactance per unit length of line is large compared to the series resistance per unit length and that the reciprocal of the shunt capacity reactance is large compared to the shunt conductance. This assumption is justified for most practical transmission lines at broadcast frequencies. Assume a total permissible line attenuation corresponding to attenuation constant A_1. If the length of line is X feet, then it can be shown by substitution in the above equation that the number of pads N is related to the length of line by the following expression:

$$N = 300A_1 - 10X.$$

Fig. 3 is a plot of this expression for several values of line loss. Thus if a permissible total line loss of about 4 decibels is assumed, seventy-five receivers may be supplied from a 750-foot line. Obviously such a difference in signal is inappreciable and such an arrangement should be satisfactory if a broad band amplifier of some 30-decibel gain is employed at the antenna end of the transmission line. If the pads are arranged as shown in Fig. 4 the loss between any two connected receivers will be about 65 decibels.

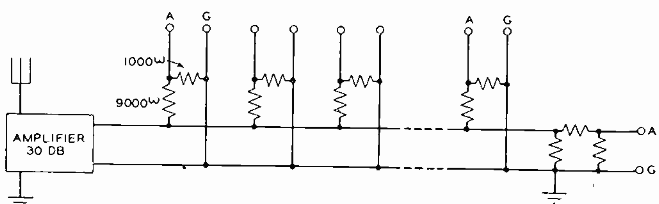

Fig. 4—Resistance pads as coupling elements.

Untuned amplifiers may also be employed as coupling elements as shown in Fig. 5. If neutralized they constitute excellent coupling elements since they may be made to operate as substantially unilateral devices. The loss from plate to grid of a neutralized amplifier is about 70 decibels at broadcast frequencies whereas unneutralized three-element tubes show a loss from plate to grid of about 25 decibels. Moreover the gain of the amplifier may be made to compensate for system losses. In general, systems employing coupling amplifiers will accommodate several hundred connected receivers, the practical limit being determined by line losses and effective amplifier gain.

The principal disadvantage of amplifiers as coupling elements lies in the fact that modulation between strong signals or between a strong carrier and the filament supply frequency, due to the nonlinear operational characteristic of the tube and circuit, may occur. A filament type tube utilizing alternating current directly on the filament and employing the customary center tap or its equivalent will modulate a signal with the second and higher order harmonics of the filament supply frequency. The degree of modulation or the magnitude of the interference side bands are in general functions of the signal amplitude and of the square of the filament supply voltage. This is easily shown by considering the noise due to:

1. Change of potential of plate and grid with respect to a given point on the filament with time.
2. The nonlinear operational characteristic of the tube and associated circuits.

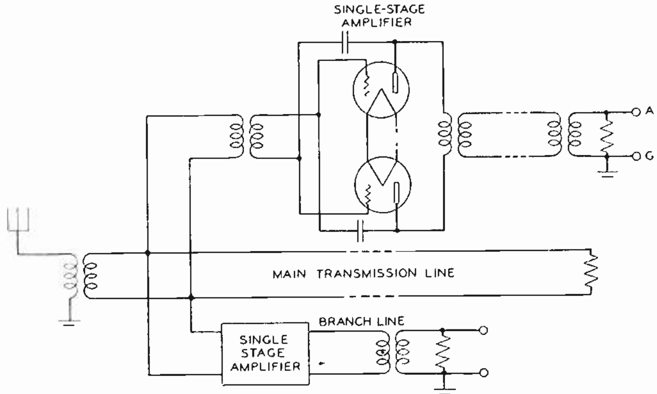

Fig. 5—System using active coupling elements.

For simplicity consider a straight homogeneous filament all points of which are symmetrical with respect to grid and plate. Let the filament supply center tap be common to plate and grid circuits (see Fig. 6). Let $(e = C \sin \omega t)$ represent a signal voltage applied to the grid and let

$\left(E_\gamma = \frac{E_B}{\mu} + E_c\right)$ represents the effective grid potential and

$(e_2 = A \sin pt)$ represents the filament supply voltage.

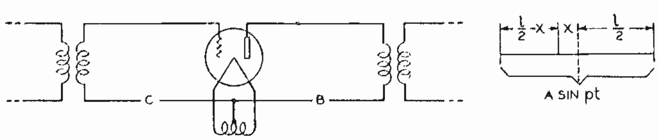

Fig. 6—Simplified tube circuit with alternating-current filament supply.

If the tube characteristic is given by

$$i = \beta_1 e + \beta_2 e^2 + \beta_3 e^3 + \cdots \beta_n e^n$$

where e represents the total voltage referred to the grid circuit of the tube, it remains to evaluate

$$i = F(e^n).$$

The voltage at distance X from the center of the filament will be

$$\frac{X}{l} A \sin pt$$

where l represents the length of the filament. The differential space current for length dx of the filament will be

$$di = K\left[(E_\gamma + C \sin \omega t) - \frac{x}{l} A \sin pt\right]^n dx.$$

Integration shows that the coefficients of one of the two interference sidebands which after detection yields the second harmonic of the filament supply frequency [i.e., sin $(\omega \pm 2p)t$] is given by

$$\alpha\left[\frac{n(n-1)(n-2)}{96} A^2 C E_\gamma^{(n-3)}\right.$$

$$+ A^2 C\left(\frac{C^2}{768} + \frac{A^2}{7680}\right) n(n-1)(n-2)(n-3)(n-4) E_\gamma^{(n-5)}$$

$$\left.+ \frac{n(n-1)\cdots(n-6)}{61440} A^4 C^2 E_\gamma^{(n-7)} + \cdots\right].$$

In addition to these two interference side bands there are a number of other modulation products which contribute to the resulting interference. Since modulation is due to nonlinearity of the tube characteristic, it follows that it will be much more severe on tube overload and since for third and higher order terms the coefficient of sin $(\omega \pm 2p)t$ is proportional to A^2C to a first approximation it is obvious that very large carriers will be accompanied by an appreciable interference side band which, when demodulated by the receiver, will result in the second harmonic of the filament supply frequency. Since the interference side bands do not appear in the common leg of the output circuit of push-pull amplifiers it cannot be balanced out but is transmitted along with the signal. At full load the effective modulation of a carrier by the harmonic of the filament supply may be of the order of 5 per cent for average filament type tubes whereas a value of 0.1 per cent should be maintained for satisfactory operation. There are several obvious methods of reducing such interference such as low voltage, high current filaments, but most of these measures bring into play factors other than those considered above with a resulting interference level which is still annoying.[8] The obvious remedy lies in the use of quiet indirect heater tubes.

By a similar process of reasoning it can be shown that cross modulation (i.e., third order) resulting in the transfer of modulation from a strong signal to a weak carrier can be expressed as[9]

$$\frac{3\beta_3}{\beta_1} P^2 M$$

which represents the degree of modulation of the weak carrier by the signal of the strong carrier. In this expression β_1 and β_3 are first and third order coefficients of the power series representing the tube characteristics, P the amplitude of the strong carrier, and M the percentage modulation of the strong carrier. In a good amplifier tube the values of these coefficients are of the order of 10^{-3} and 10^{-4}, respectively. Similar reasoning indicates the necessity of reducing the second and other higher order coefficients to a minimum if intermodulation between signals in the coupling amplifiers is to be avoided.

The use of balanced or push-pull coupling amplifiers will reduce even order modulation products. The use of adjustable band elimination filters in the antenna circuit to reduce the signals of near-by local stations to values well below the overload point of the tubes will effectively reduce odd order effects since the amplitude of the interfering voltage appears as a squared term. It is obvious therefore that while amplifiers may be excellent coupling devices considerable care must be exercised in their design and manufacture if satisfactory results are to be obtained. Quiet heater type tubes, with a high degree of linearity are a necessity. Multigrid tubes are generally unsuited to this type of service. In general a low-μ tube capable of a large grid swing is most suitable. It has been found entirely practicable to produce amplifiers in which the hum modulation is less than 0.1 per cent at full load of several volts applied to the grid in which the even and odd order modulation products are down some 60 decibels and 30 decibels, respectively, compared to the fundamental.

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[8] K. H. Kingdon and H. N. Nott-Smith, "The operation of radio receiver vacuum tubes on alternating current," *Gen. Elec. Rev.*, pp. 139 and 228; March and April, (1929).

W. J. Kimmell, "Cause and prevention of hum in receiving tubes employing alternating current direct on the filament," Proc. I.R.E., vol. 16, p. 1089; August, (1928).

E. Lofgren, "Elimination of hum in mains operated radio receivers by means of compensation methods," *The L. M. Ericsson Review*, No. 4–6, p. 171, (1932).

J. O. McNally, "Analysis and reduction of output disturbances resulting from alternating-current operation of the heaters of indirectly heated cathode triodes," Proc. I.R.E., vol. 20, p. 1263; August, (1932).

[9] E. Peterson and F. B. Llewellyn, "The operation of modulators from a physical standpoint," Proc. I.R.E., vol. 18, p. 38; January, (1930).

F. A. Polkinghorn, "Short-wave transoceanic telephone receiving equipment," *Proc. Radio Club Amer.*, December, (1932).

From the foregoing discussion it is evident that there are a number of possible types of radio-frequency distribution systems. The choice of any given arrangement depends primarily on the service requirements which it is intended to meet. Obviously workable systems containing only passive elements can satisfactorily serve only a relatively small number of connected receivers without excessive signal loss if a reasonable degree of isolation of the connected receivers is to result.

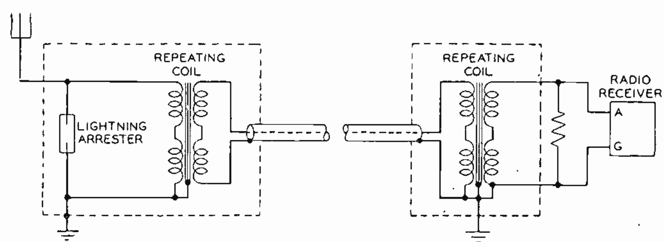

Fig. 7—Simple radio-frequency distribution system serving one receiver.

However such systems are relatively inexpensive and require practically no maintenance. Large systems require amplifiers to compensate for coupling element losses. Such systems when properly designed will not only serve more connected receivers but will permit of a greater degree of receiver isolation. Moreover in installations intended to serve one hundred or more receivers they may represent a smaller total investment than a number of passive element systems each serv-

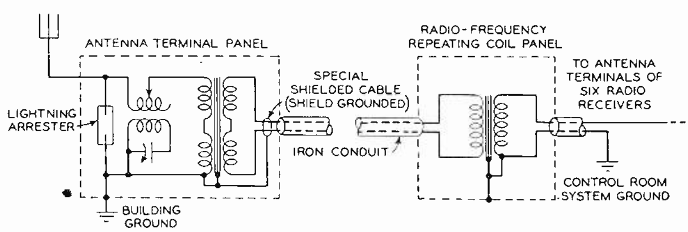

Fig. 8—Passive radio-frequency distribution system installed in Waldorf-Astoria Hotel.

ing a part of the connected receiver load. Maintenance on such systems can usually be limited to tube replacements. It is evident, however, that both types of systems are required to cover the field adequately.

One commercially available type of radio-frequency distribution system consists of an unloaded transmission line, which may be of either the balanced or unbalanced type, coupled through repeating coils to the antenna and to the radio receiver. This system will feed from one to fifteen radio receivers connected in multiple, but it may be used to supply one radio receiver of any type as shown in Fig. 7.

Coupling devices are unnecessary when high input impedance receivers are used due to the fact that these receivers are designed for this type of service. Fig. 8 illustrates the system employed to feed

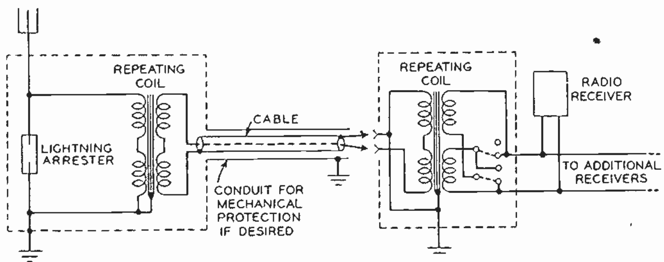

Fig. 9—Western Electric No. 1A radio-frequency distribution system.

six Western Electric receivers in the Waldorf Astoria Hotel in New York City. This particular installation employs a balanced and shielded transmission line some 600 feet in length terminated at both the antenna and receiver ends in balanced and shielded transformers. The design of the input circuits of the receivers and the repeating coil are such that the receivers constitute a proper termination for the line. The loss in each repeating coil is about 0.25 decibel and that of the

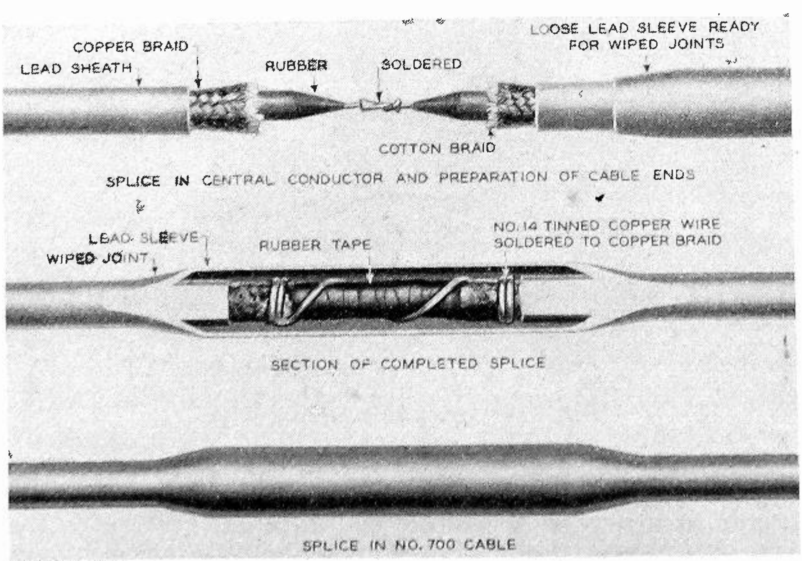

Fig. 10—Splice-in head cover coaxial transmission line.

transmission line about 0.3 decibel per hundred feet at 1000 kilocycles. Unbalanced lines, such as the coaxial conductor type, may be employed with this system. The losses in coaxial conductor lines increase as the square root of the frequency if conductance losses are negligible.[5] If rubber insulation is employed it is to be expected that the loss would increase at a rate greater than the square root of frequency and less than the first power of frequency. This has been verified experimentally. The impedance of the coaxial conductor line which is illustrated in Fig. 10 is 65 ohms. This cable consists of a solid inner conductor of No. 18 untinned copper separated from a one-eighth-inch rubber wall by a cotton wrap. A second cotton wrap separates the rubber wall from a heavy untinned copper braid which is in turn covered with a heavy lead sheath for moisture proofing. The cable is intended for underground use or for use in damp locations. Fig. 11 illustrates the

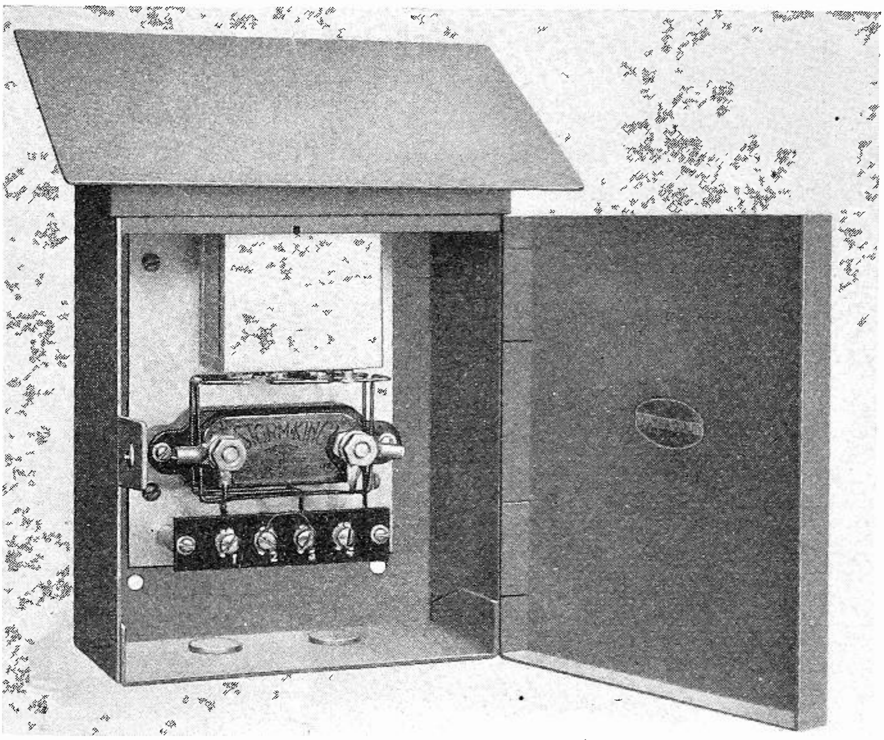

Fig. 11—Antenna coupling panel.

antenna coupling panel which contains a repeating coil and lightning arrester and is intended to couple the antenna system to the transmission line. It is intended for mounting in a metal box and should be located as close to the antenna as practicable to prevent interference pickup by the exposed lead-in.

A similar system using amplifiers as coupling elements and intended to supply ten broadcast radio receivers of any type is shown on Fig. 12. This system employs an antenna coupling panel to couple the antenna system to a balanced transposed transmission line. Single stage balanced amplifiers are employed as coupling elements. Each amplifier ee a single receiver through a length of balanced transposed transmission line terminated in a repeating coil.

The antenna coupling panel is similar to the panel described above except that provision is made for mounting four sections of high-pass, low-pass, or band-elimination filter. It is intended for mounting in a nel box and is located as close as practicable to the antenna.

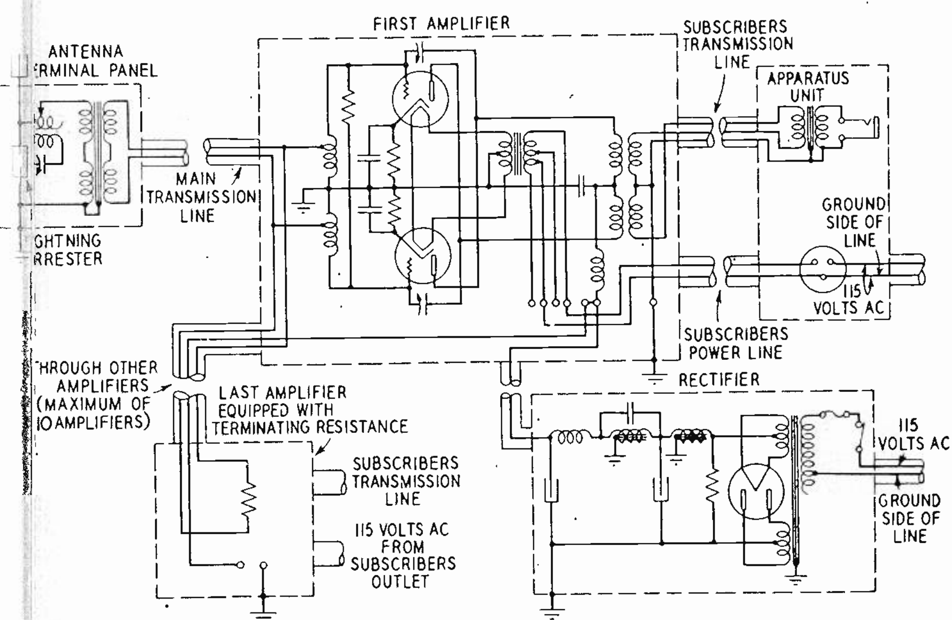

Fig. 12—Western Electric No. 2A radio-frequency distribution system.

The balanced transmission line used with this system consists of No. 20 twisted telephone bridle wire. The impedance of this type of line is 100 ohms and the loss about 0.6 decibel per hundred feet. This wire is intended to be pulled into rigid conduit or wire mold. Flexible conduit is very unsatisfactory for this type of service since it may act as continuous loading due to the formation of oxide between the surfaces in contact and in some cases, due to its inductive character, it may induce interference in the transmission line from local interference sources.

The coupling units consist of single stage neutralized push-pull amplifiers (see Fig. 13), each supplying one connected receiver. Since the amplifiers are fed by low impedance lines, balanced step-up transformers are used to effect a voltage step-up of some 15 decibels into the tubes. Three-element tubes are employed to obtain low second and third order coefficients and to permit a large grid swing without appreciable modulation. The filaments of the tubes are connected in series so that heater failure of either tube renders the amplifier inoperative. Otherwise poor performance might result due to failure of one heater.

A single power supply unit is required for ten amplifiers. The regulation of this unit is sufficient to insure proper operating conditions whether one or ten amplifiers are energized. The transmission line between the coupling amplifier and its connected receiver may be 500

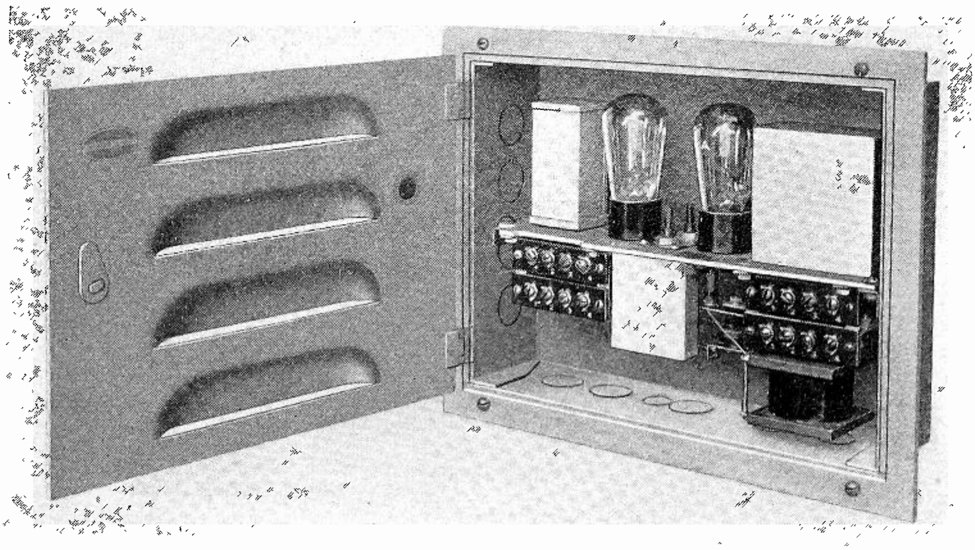

Fig. 13—Coupling amplifier.

feet in length. This permits considerable flexibility in that the coupling amplifiers may be located in a group at some central location if desired. It is evident that the recommended length (500 feet) of transmission line may be exceeded by any reasonable amount, the additional loss of 0.6 decibel per hundred feet being the only penalty.

A similar system employed in the Waldorf Astoria Hotel which is intended to supply a maximum of 180 radio receivers from a single antenna is shown in Fig. 14. The essential differences between this system and the one just described lie in the use of a loaded transmission line and amplifiers which operate directly from the loaded line without the use of input transformers. An antenna coupling panel is used to couple the antenna to the transmission line. If there is an appreciable length of line (say 50 feet) between this panel and the first coupling amplifier, this portion of the line is unloaded. Balanced, transposed lines are used throughout. The first coupling amplifier contains a step-up transformer which steps up the line impedance to 1000 ohms. The remainder of the transmission line is loaded unless there is a run of 50 feet or so containing no amplifiers, in which case it may be economical to step the line impedance down for this run. Loading coils are assembled in astatic pairs and each pair is matched to 0.1 per cent. Space is provided in the amplifier for mounting the loading coil as-

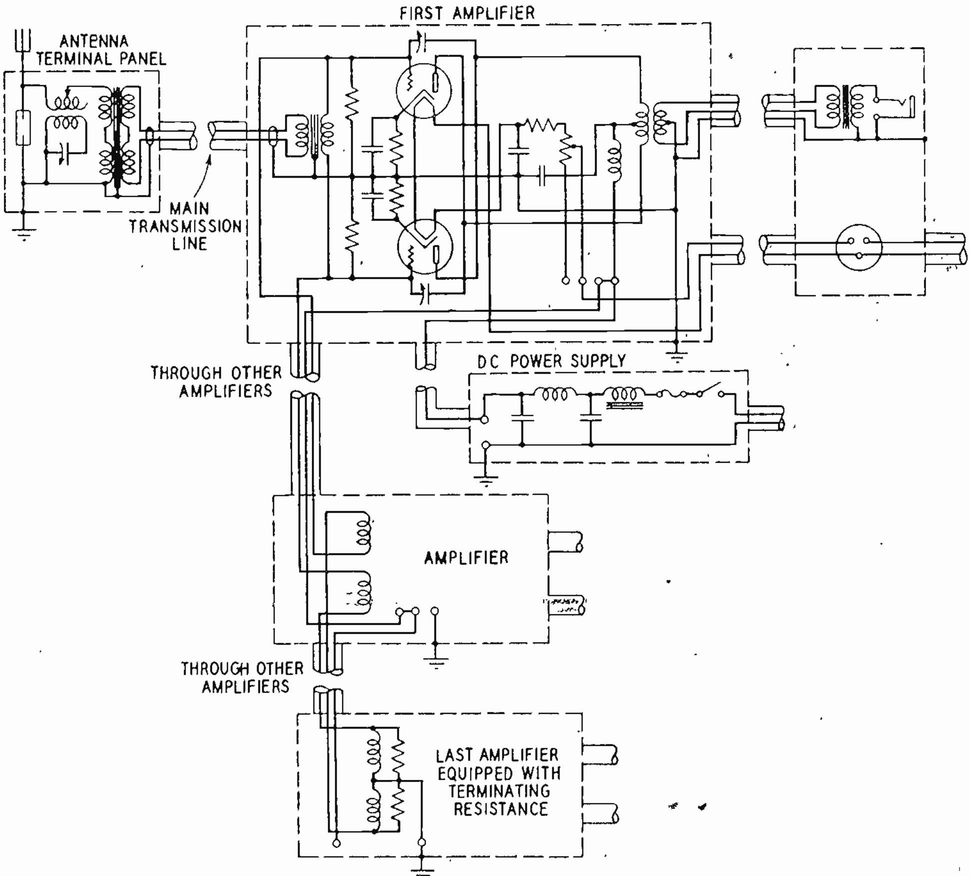

Fig. 14—System installed in Waldorf-Astoria Hotel.

sembly as shown in Fig. 15. It is frequently impractical to install loading coil boxes in accessible locations at appropriate intervals along the main transmission line, and since amplifiers will not always be located exactly at loading points, provision is made in the coupling amplifiers for the installation of shunt condensers across the line to correct for improper inductance spacing. Indeed it is desirable to omit loading coil boxes entirely, a feature which may be realized with this system. This permits loading coils to be installed in the most convenient amplifier. The input impedance of the neutralized coupling amplifiers is, of course, entirely capacitative. The spacing of the loading coils therefore depends on the number of amplifiers connected across the line since the total capacity per section must be a constant (i.e., within about 25 per cent). While amplifiers do not in themselves constitute a loss they effect the allowable length of line in so far as they affect the number of loaded sections. The last amplifier houses the termination resistance assembly. This assembly consists of an accurately center tapped resistance shunted by two low resistance choke coils. The coils serve as a direct-current short circuit to prevent tube shorts, etc., rendering the system inoperative. As a result plate-to-grid short circuits, etc., affect only the amplifier in trouble.

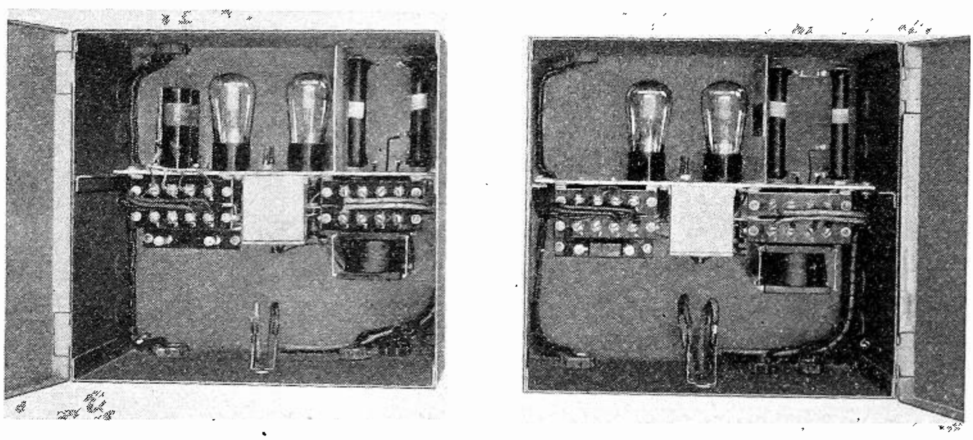

Fig. 15—Coupling amplifier with and without loading coils.

The coupling amplifiers are balanced and neutralized. Since they operate from a high impedance line, input transformers are not required. The tube heaters are connected in series so that a heater failure renders the amplifier inoperative. Otherwise poor performance might result due to the failure of one heater. Provision is made in the receiver outlet for energizing the coupling amplifier heater with the radio receiver switch if desired. This is accomplished by the use of a three-conductor cord for the receiver and a three-pole power receptacle.

The necessity of providing filters is illustrated by the system installed in the Waldorf Astoria Hotel. In this installation the antennas are located between the towers at a height of about 650 feet from the street. The antenna of television station W2XAB which operated on 2800 kilocycles was located about 400 feet air line from these receiving antennas. As a result a signal of about seven volts was induced in the Waldorf receiving antenna from W2XAB. Although the loss to 2800 kilocycles in this particular system is considerable, there occurred an appreciable amount of cross modulation in some of the connected receivers between W2XAB and local broadcast stations. This condition was effectively eliminated by the use of three sections of low-pass filter in the antenna coupling panel with a loss of some 60 decibels to

2800 kilocycles. This same condition might have occurred due to a near-by broadcast transmitter in which case adjustable band-elimination filters would have been required.

A common amplifier type of radio-frequency distribution system is shown diagrammatically in Fig. 16. An antenna coupling panel is used near the antenna to couple the antenna to a coaxial conductor transmission line. This line, which may be 750 feet in length, is connected through a filter panel to a maximum of four power amplifiers. Each of these amplifiers supplies a maximum of ten coaxial conductor branch lines to each of which a maximum of seventy-five radio receivers may be connected. The maximum recommended length of the branch lines is 750 feet.

The filter panel includes five adjustable band-elimination filters, although high-pass or low-pass filters may be mounted interchangeably if desired. The purpose of this panel is to prevent an excessively strong signal entering the system which might cause modulation in either the radio receivers or the amplifiers. Each filter section has a loss of about 20 decibels at the frequency to which it is adjusted, so that an over-all loss of 100 decibels is obtainable at any frequency in the band. The loss outside the suppressed band is negligible. The input impedance of the power amplifier is about 300 ohms. Consequently four of these amplifiers connected in multiple constitute the correct termination for either the filter panel or the coaxial conductor transmission line. These amplifiers are arranged either for rack or cabinet mounting and consist of four stages of neutralized push-pull aperiodic amplification, the tubes of the last stage having a rating of about 2.5 watts each for class A operation. The over-all gain is about 40 decibels. Second order modulation products at full load are down about 80 decibels in a properly adjusted amplifier. Third order modulation products are down some 65 decibels.

Radio receivers are connected to the branch lines through resistance pads having a loss of about 25 decibels. The loss between any two connected receivers is about 65 decibels. One of the coupling units, which comprises a three-resistance pad, is used only at the end of the line, one of the resistors serving as a line termination. The other coupling unit, which comprises a two-resistance L-type pad, is used to couple receivers to the line at intermediate points. This type of coupling element presents a fixed load to the line and the resulting line loss is practically independent of the number of connected receivers, depending instead on the number of outlets. These units are intended for mounting in a single gang switch box or in combination with the power outlet in a double gang box. A total maximum branch line loss of 4 decibels has been chosen as suitable for this type of service.

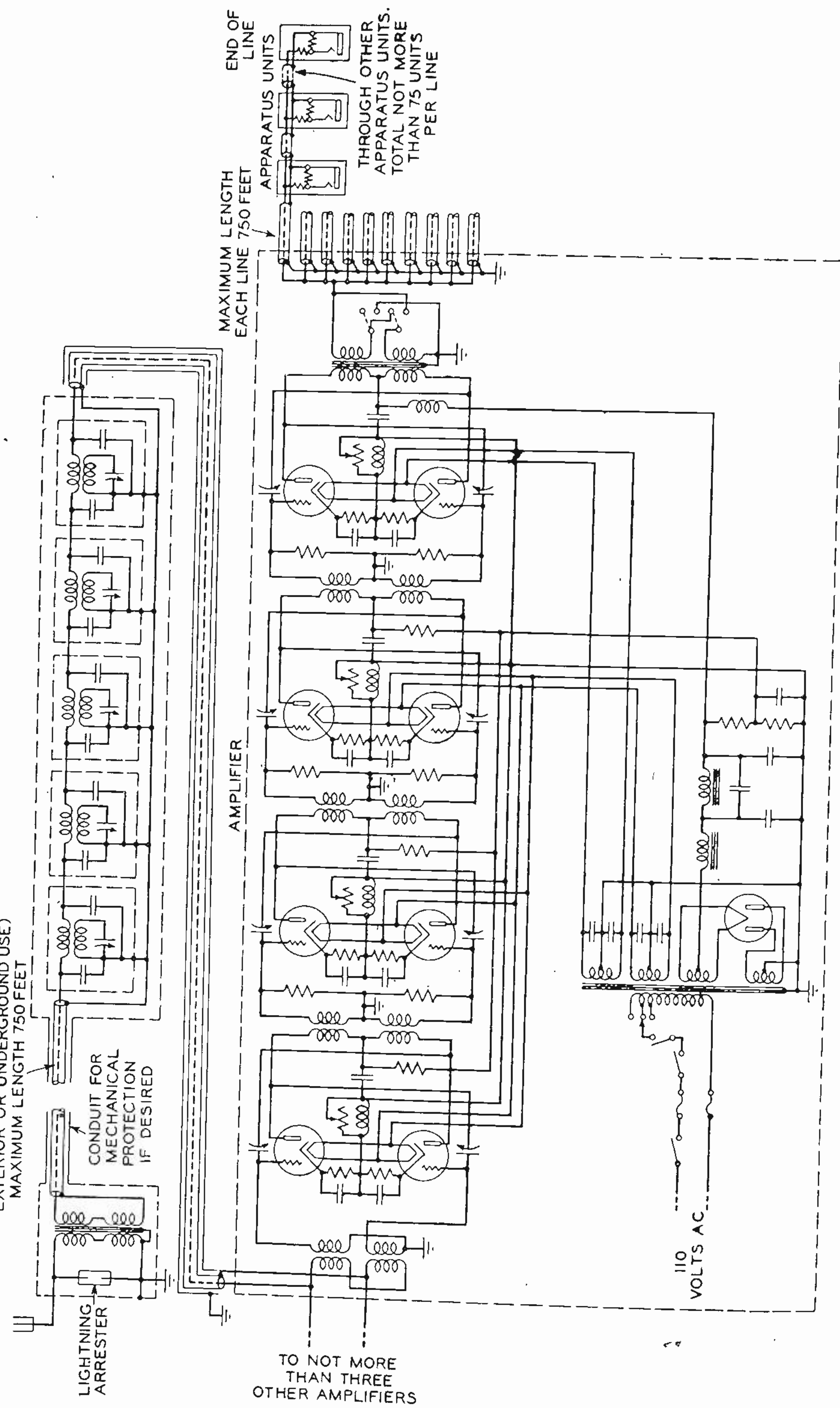

Fig. 16—Western Electric No. 3A radio-frequency distribution system.

For large installations this system is economical and easily installed. A normal installation will serve three thousand connected receivers with a maximum possible difference in delivered signal level of about 4 decibels. The degree of isolation between any two connected receivers is better than can be obtained by unneutralized three-element tube coupling amplifiers and is practically the equivalent of neutralized coupling amplifiers. Moreover the filter panel provides means of reducing the signals of near-by radio transmitters sufficiently to prevent cross modulation in connected receivers having insufficient high-frequency selectivity. It is obviously entirely practicable to extend the number of receivers to be served by a single antenna from three thousand to a much larger number by using amplifiers as boosters at any receiver location or at the end of any of the branch lines.

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DEVELOPMENT OF CATHODE-RAY TUBES FOR OSCILLOGRAPHIC PURPOSES*

By

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Summary—*Some typical electrical characteristics of a cathode-ray tube electron gun are shown and the function of the various gun elements described. Light output, luminescent screen efficiency, space distribution of radiation, as well as decay and spectral distribution characteristics of willemite screens, are shown and the relations of the various factors discussed. The starting and dynamic characteristics of the tube are discussed. Magnetic and electrostatic deflection are discussed with regard to sensitivity, frequency range of application, and impedance of deflection plates. In conclusion, a few general precautions in the operation of cathode-ray tubes are given.*

THE growing realization by engineers of the adaptability of cathode-ray equipment to the quick and reliable solution of many high-frequency problems presents a challenge to the designer of cathode-ray tubes. Needless to say, the designer has accepted the challenge by developing new and improved types of cathode-ray tubes. These tubes are of rugged construction, have long life, high sensitivity, good brilliance control, and are not critical in adjustment. In fact, they possess reliability in operation and simplicity of control comparable with the same features of radio receiving tubes.

In the early stage of development work on cathode-ray tubes conducted in the laboratory of the RCA Radiotron Division of the RCA Manufacturing Company, Inc., it was apparent that two lines of attack were feasible; first, an investigation of the characteristics of the gas-filled type, and second, an investigation of the characteristics of the high vacuum type. After a careful consideration of both types, a decision was made in favor of the high vacuum type. A high vacuum tube may be considered as one which is evacuated to a pressure of lower than 10^{-5} millimeters of mercury. It may be well to mention some of the considerations which determined this choice. In a gas-filled tube, positive ion bombardment of the cathode is a troublesome factor, especially where an oxide-coated cathode is used. The use of an oxide-coated cathode is very desirable because it operates at a low cathode temperature and produces practically no illumination to conflict with the luminescent pattern on the screen. Bombardment of this form of cathode shortens tube life appreciably at low voltages and

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seriously at high voltages. In a given tube it is necessary, in order to obtain optimum focus, to adjust the beam current to a critical value for a particular operating voltage condition. This requirement makes control of the light intensity over a wide range difficult. Again, the control of a definite low gas pressure (a few microns) complicates manufacture. Furthermore, because of this small amount of gas, it is not easy to prevent an appreciable change of gas pressure during tube life due to absorption or liberation of gas by the glass walls or metal parts. The high vacuum tube does not present these difficulties, although another means of focusing is necessary to take place of the gas concentration.*

It is known that the direction of a moving electron may be controlled by either a magnetic or electrostatic field, and that either or both may be used to focus a beam of electrons to produce a small spot. The theory of magnetic concentration has been developed by H. Busch[7,8,9] and Thibaud.[10] The application of magnetic focusing to the electron microscope for high magnification is given by E. Ruska.[11,12] The theory of electrostatic focusing has been recently outlined by V. K. Zworykin,[13,14] I. G. Maloff and D. W. Epstein.[15] A number of papers describing experimental work on these methods of focusing have been published by Knoll and Ruska,[16] Ruska,[17] and Johannson.[18] The recent book by E. Bruche and O. Scherzer[19] gives an excellent survey of the field of electron optics, covering electric, magnetic, and gas-focusing methods. To focus fast-moving electrons magnetically requires considerable power in the magnetic coil which is placed coaxially around the tube neck. Because of this, and because of the complications which may arise through the presence of electrostatic accelerating fields and interaction with magnetic deflection fields, it is often desirable to use electrostatic focusing. This method depends almost solely on the ratio of voltages applied to two of the electrodes.

The Electron Gun

Fig. 1 is a cross-section diagram of the electron gun and shows the electrodes, equipotential lines, and beam outline. As mentioned above, the cathode is oxide coated, of the unipotential type, and heated by a noninductive heater coil. The cathode is surrounded by a control elec-

* The exact mechanism of gas concentration is probably not yet clearly understood, but a review of its action may be found in the papers by W. Ende,[1] von Ardenne,[2] and O. Sherzer.[3] The determining factors in the production of a gas-concentrated electron beam are the kind of gas, the gas pressure, the current strength, construction of accelerating electrodes, and the strength and configuration of the resulting electric field. The relations of these factors in the production of a gas-concentrated electron beam are quite clearly understood and are outlined by E. Bruche.[4,5,6]

[1] Numbers refer to Bibliography.

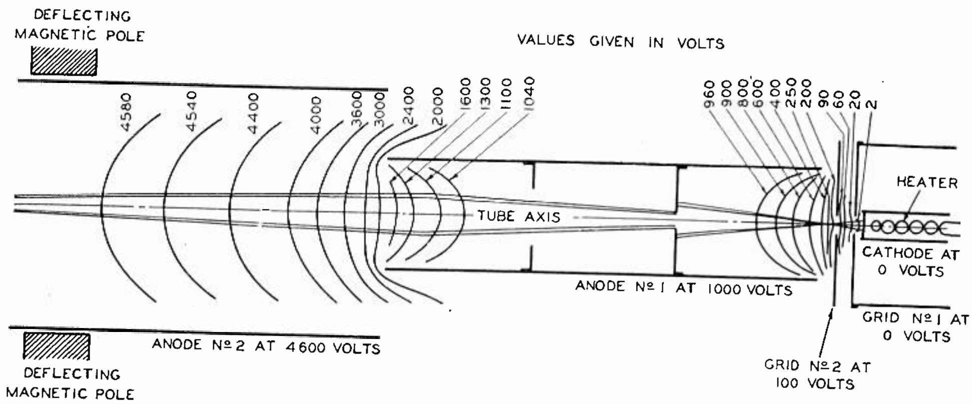

Fig. 1—Typical electron gun diagram, beam outline, and equipotential line plot.

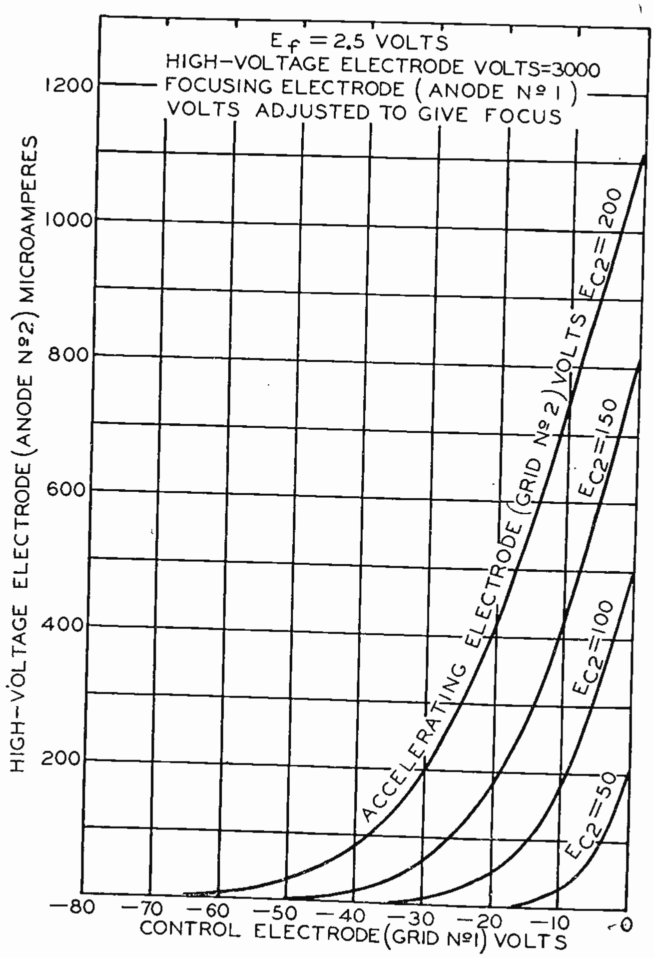

Fig. 2—Typical beam current control characteristics of an electron gun (grid No. 2 voltage as parameter).

trode or No. 1 grid. The amount of current drawn from the cathode is determined by the negative bias on this electrode. A nominal positive potential on the No. 2 grid or accelerating electrode gives the emitted electrons their initial acceleration toward the luminescent screen. The field shape in this region also tends to bring the beam to focus. However, the main focusing field is between the No. 1 and No. 2 anodes.

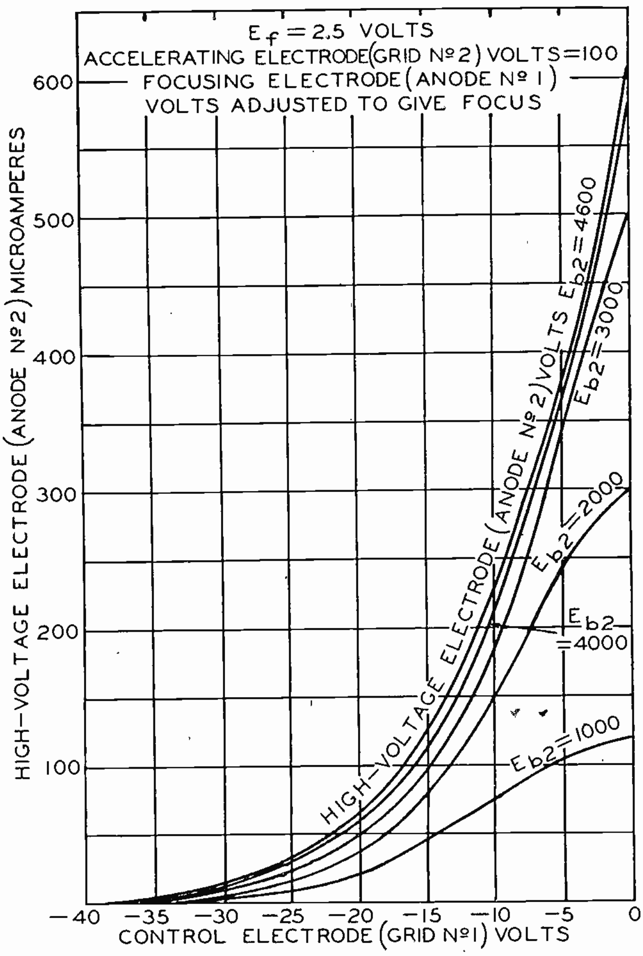

Fig. 3—Typical beam current control characteristics of an electron gun (anode No. 2 voltage as parameter).

The No. 2 anode is the high voltage electrode which gives the electrons their final energy for producing luminescence of the screen. For focus, the No. 1 anode is run at between one fourth to one fifth of the positive potential on the high voltage electrode. For a particular gun structure, the ratio of the anode voltages to produce focus is a constant and is independent of the magnitude of the applied voltage.

In brief, the paths of the electrons from cathode to luminescent screen are initially convergent through the control electrode, and focus in the region of the accelerating electrode. From here, they diverge to the main focusing field where they are again made to converge but this time at a point on the luminescent screen. Those electrons which are too divergent to be focused by the main focus field are stopped by a beam-defining aperture in the No. 1 anode. In some gun structures, the accelerating electrode or No. 2 grid is omitted, and in this case, the No. 1 anode provides the initial accelerating and focusing field for the electron beam.

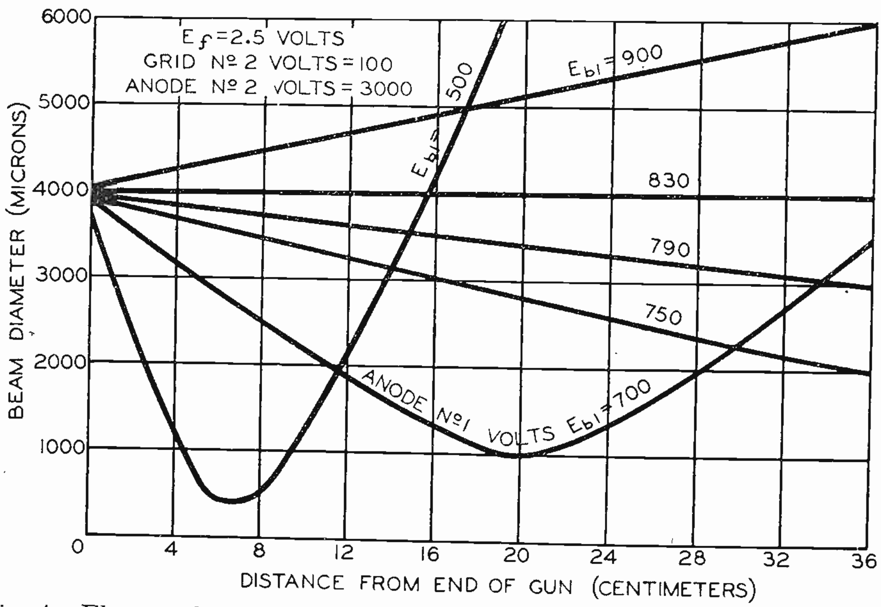

Fig. 4—Electron beam envelope from anode No. 1 to luminescent screen.

Figs. 2 and 3 show typical control characteristics, variations of the beam current with negative voltage on the control electrode. The curves in Fig. 2 were taken with a constant potential of 3000 volts on the No. 2 anode, while those in Fig. 3 were taken with a constant potential of 100 volts on the No. 2 grid. These curves show that the beam current for a given control electrode voltage increases with both the No. 2 grid and the No. 2 anode voltage but is more dependent on the No. 2 grid voltage especially at high No. 2 anode voltages.

The curves in Fig. 4 show the variation in the beam envelope with changes in the No. 1 anode voltage, the No. 2 anode voltage being held constant. With a high value of No. 1 anode voltage, the focusing field between the two anodes is insufficient to converge the beam toward a focal point on the screen, and the beam can be seen to diverge. The divergence of the beam is decreased by lowering the No. 1 anode voltage. At a particular value of No. 1 anode voltage, the beam is almost parallel. Further decrease in the No. 1 anode voltage causes the beam to focus at some distance from the gun. The distance between the gun and the focal point decreases as the No. 1 anode voltage is decreased. It will also be noted that as the focal point becomes closer to the No. 1 anode, the size of the focal spot becomes smaller. This change in spot size follows the laws of geometrical electron optics. The curves in Fig. 5 show that the spot size as measured usually decreases with an increase in No. 2 anode voltage and increases linearly with the square root of the beam current. Therefore, if a very small spot size

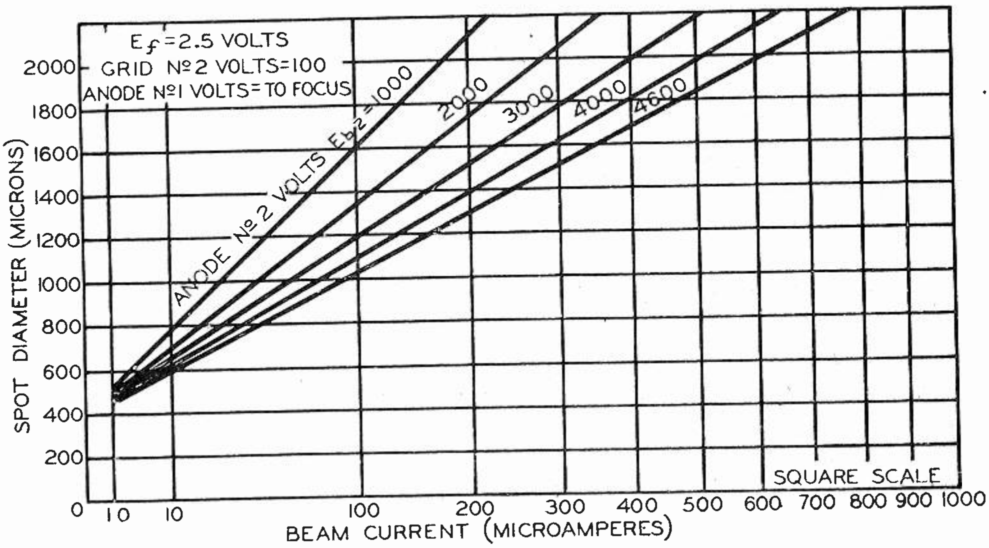

Fig. 5—Typical spot-size characteristics.

is desired, the tube may be operated at a high No. 2 anode voltage with a low beam current.

The Luminescent Screen

The luminescent screen serves to convert the kinetic energy of the electron beam into light. The spectral distribution and persistence characteristics of the light emitted are determined by the kind of luminescent material or phosphor* while the space distribution of the emitted light depends upon the particle size and screen thickness. The efficiency of the screen depends upon the kind of material as well as the particle size, screen thickness, and secondary emission. Zinc orthosilicate, commonly known as willemite, has been selected as a luminescent material suitable for cathode-ray tube screens for numerous applications, largely because of its inherently high efficiency, its stabil-

* Phosphor is a general term referring to the solid material in the screen which produces luminescence when excited. *See* Perkins and Kaufmann, "Luminescent materials for cathode-ray tubes," Proc. I.R.E., this issue, pp. 1324–1333.

ity under electron bombardment, and the fact that the maximum of the emission band falls at about 5250 angstroms which is very near

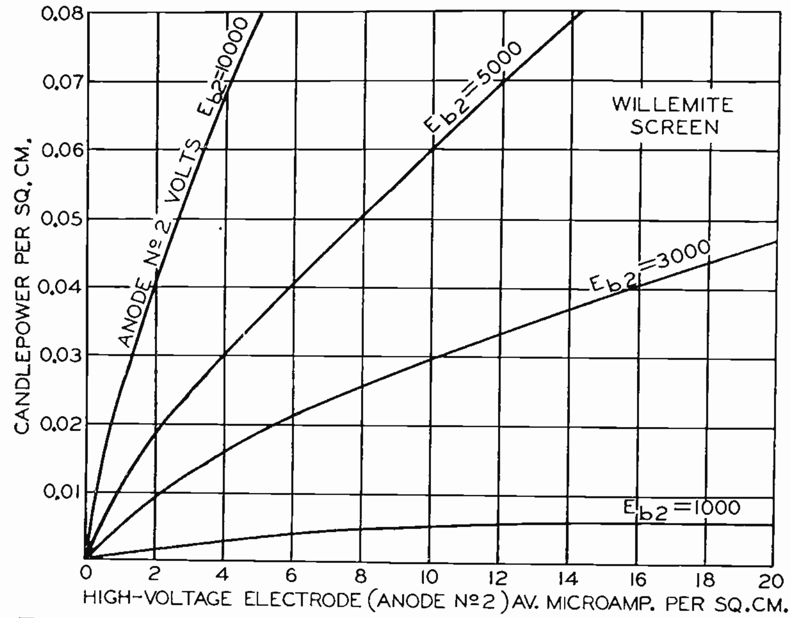

Fig. 6—Average luminescent screen brightness characteristics (anode No. 2 voltage as parameter).

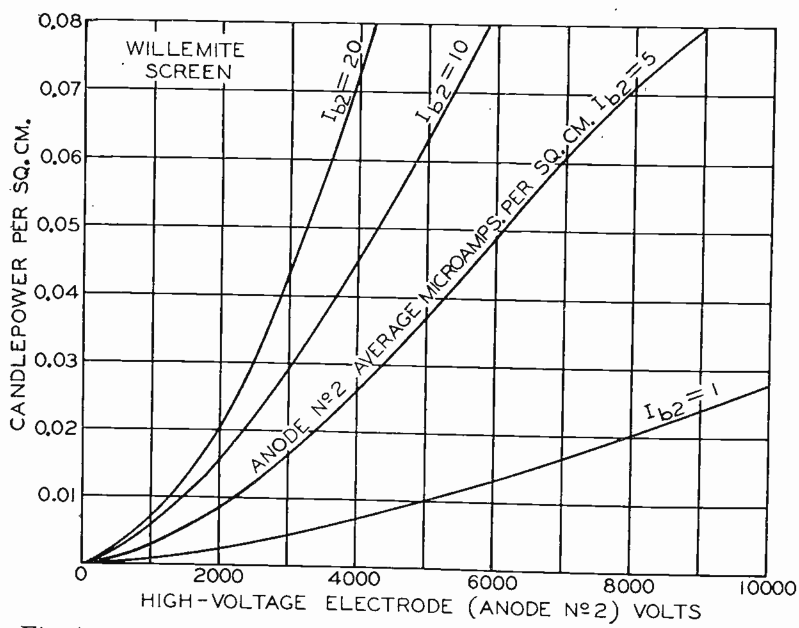

Fig. 7—Average luminescent screen brightness characteristics (anode No. 2 current as parameter).

the maximum sensitivity of the eye (5550 angstroms). This screen material is identified for convenient reference in this paper as No. 1 phosphor.

The following curves, Figs. 6–9, show average characteristics of No. 1 phosphor which are of interest from an operating standpoint.

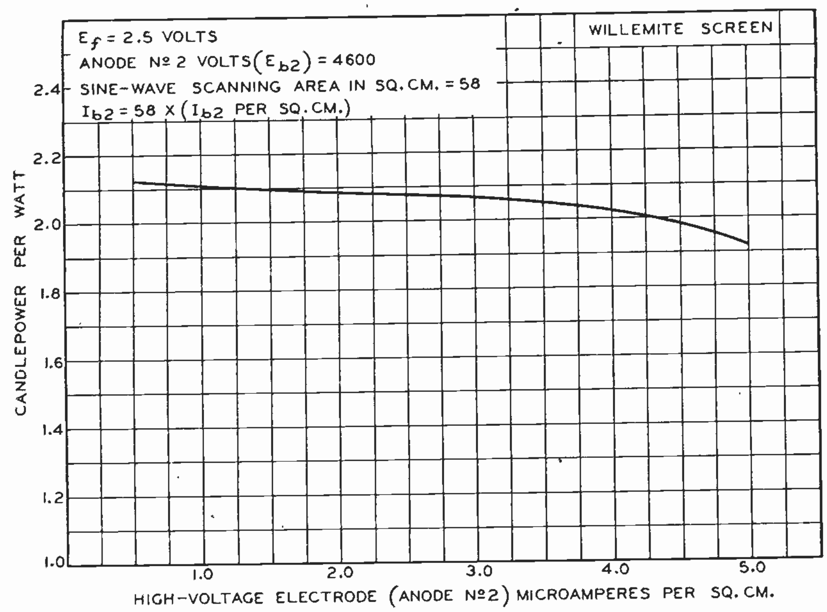

Fig. 8—Average luminescent screen efficiency characteristic (anode No. 2 current varied).

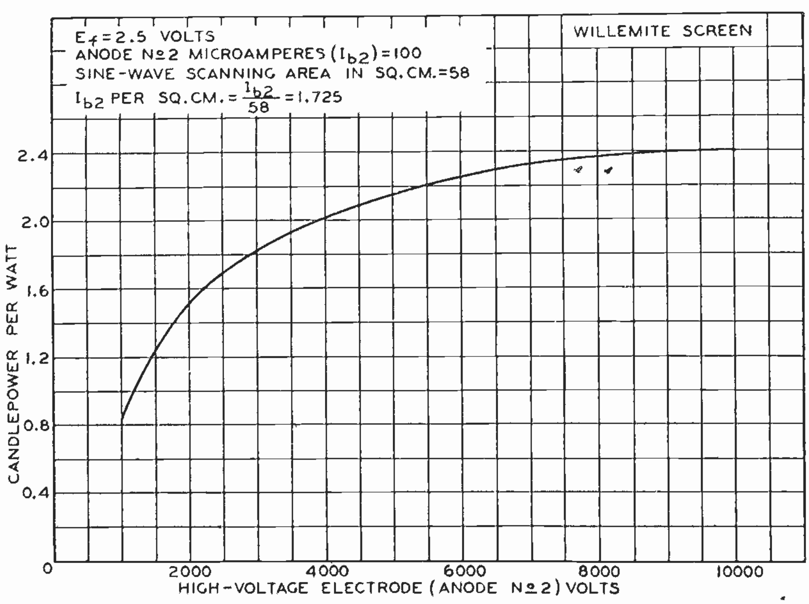

Fig. 9—Average luminescent screen efficiency characteristic (anode No. 2 voltage varied).

The curves in Figs. 6 and 7 show the variation of brightness of the luminescent screen in candle power/cm² with beam current density and No. 2 anode voltage. In Fig. 6, the curves are plotted directly against beam current density, while in Fig. 7 they are plotted against No. 2 anode voltage for different values of beam current density. It will be noted that the candle power increases more rapidly with voltage than with current. The curves in Figs. 8 and 9 show the variation of screen efficiency in candle power/watt with the beam current, as well as with beam current density, and with voltage. These curves show that the screen efficiency decreases with increasing current density, but increases with increasing voltage up to about eight or nine kilovolts where it is practically a constant.

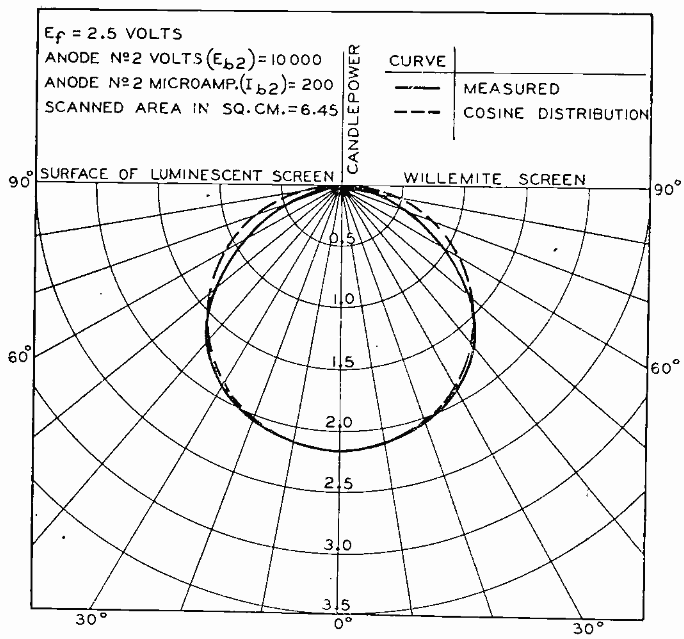

Fig. 10—Typical space distribution of light from luminescent screen.

The curves in Fig. 10 show the space distribution of light emitted from a No. 1 phosphor. The solid curve represents measured data, while the dotted curve follows the cosine-distribution law. The measured distribution is very near the cosine law. However, in some cases the relative intensity perpendicular to the surface is greater than that shown in the above curve. The distribution depends to some extent on the voltage at which the tube is operated, on the particle size of the luminescent material, and on the screen thickness, but in general is not far from the cosine law.

The curve in Fig. 11 shows the persistence characteristic of No. 1 phosphor. The intensity drops to 0.01 of its original intensity in about 0.06 second.

The spectral-distribution characteristic of No. 1 phosphor is shown in Fig. 12 where it will be noted that the maximum is at about 5250 angstroms for practical purposes, the lower limit is about 4800 angstroms and the upper limit is about 6100 angstroms.

Operating Characteristics of the Tube

The electrons impinging on the luminescent screen produce secondary electrons which are emitted in all directions and are accelerated toward the second anode or deflection plates where the beam current is collected. It is this phenomenon of secondary emission which maintains the operation of the tube. Were it not for this phenomenon, the screen would collect negative electrons from the gun until a charge

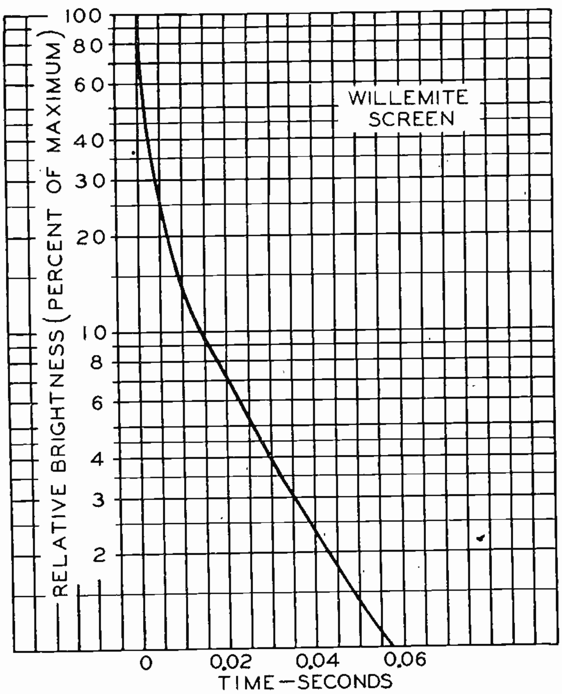

Fig. 11—Typical persistence characteristic of No. 1 phosphor.

would build up on the screen and cause the screen to be at a zero potential or slightly negative with respect to the cathode. However, the secondary emission allows the luminescent screen to float at some potential near second anode potential.

The starting characteristic, when the second anode is above ground, is interesting and may be explained as follows: The random motion of the ions produced by electron collision with the residual gas molecules in the tube causes the screen initially to become charged to near the second anode potential. This allows the gun electrons to strike the screen, after which the tube continues to function because the screen floats at the potential at which the secondary emission equals the primary beam current as described above. Although the tubes are evacuated to such a high degree that the residual gas plays no part in the focusing and does not alter the normal electrical characteristics of the tube under operating conditions, there are always gas molecules in the tube in sufficient quantity to explain the starting characteristics of the tube.

Deflection Characteristics

The beam of electrons as accelerated out of the electron gun and focused on the screen may be deflected either by magnetic or electrostatic fields. In order to move the spot in two directions over the screen, it is necessary to apply two fields whose forces are at right angles.

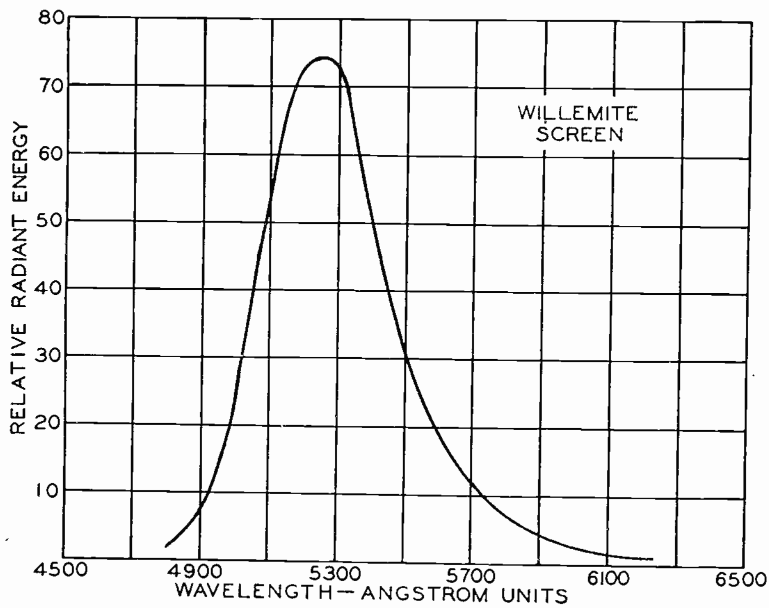

Fig. 12—Typical spectral energy characteristic of No. 1 phosphor.

These separately applied fields of force may either both be magnetic, electrostatic, or a combination of the two. It has been found possible to deflect in both directions magnetically with the fields applied at the same point along the electron trajectory. It is more difficult to obtain a distortionless pattern with electrostatic deflection in both directions when the fields are applied at the same point along the trajectory, but it is quite simple to deflect first in one direction and then the other by displacing the deflection plates along the length of the bulb neck. The combination of electrostatic deflection in one direction and magnetic deflection in the other direction is often best when it is highly necessary to eliminate distortion. When the two methods of deflection are combined both fields may be applied at the same point along the electron trajectory without any trouble from interaction between the fields. The advantages of this are that both the deflection plates and magnetic poles may be made small, and yet the fields can be practically uniform for maximum deflection in any direction.

From an application standpoint, it is well to consider the region over which the different types of deflection may best be applied. Electromagnetic deflection with coils using a core of magnetic material, besides being limited by the distributed capacity of the coils and their resonant frequency, is largely limited by the frequency characteristics of the core material. Electromagnetic deflection with air-core coils may be used throughout the band of high radio frequencies with proper coil design. Electrostatic deflections are usually the simplest when high frequencies are being studied. The potential to be studied may be connected to the deflection plates directly or merely through a coupling condenser in many applications. Where the potential is too high a simple high impedance voltage divider is feasible and where the potential is too low to cause a suitable deflection a voltage amplifier is all that is required. This makes the band of frequencies which may be studied range from direct current and audio frequencies to high radio frequencies. When ultra-high frequencies of the order of magnitude of 3×10^9 cycles are being measured, even this method may give difficulty. This is largely due to a very high-frequency deflecting field being applied to a relatively slowly moving electron beam. The result may be phase distortion, low or even zero deflection sensitivity.

For those who contemplate using the cathode-ray tube for ultra-high-frequency measurements, an expression for the electron velocity may be useful to determine the time of transit of the electron beam between the deflecting plates.

$$\text{Electron velocity} = \sqrt{2E'_{b_2}(e/m)} \text{ cms per second} \qquad (1)$$

where,

E'_{b_2} = second anode potential in absolute electromagnetic units
e = electron charge in absolute electromagnetic units
m = mass of electron in grams.

In practical units where E_{b_2} is measured in volts, the expression becomes

$$\text{electron velocity} = 5.97\times10^7\sqrt{E_{b_2}} \text{ cm per second.}$$

The above equation is not absolutely exact, but for electron velocities of less than ten per cent the speed of light, the error is negligible. Since the cathode-ray beams of most oscillograph tubes move more slowly than this figure, the equation may be taken without any correction.

A good magnetic circuit to give deflection in both directions without spot, pattern, or wave shape distortion requires considerable care in design, but a simple magnetic circuit designed to produce a uniform field over the path of the beam may be used to give good deflection in one direction. In many applications, the frequency under test may be applied to the deflection plates and the lower frequency timing wave may be applied through the magnets. Consideration must be given to the impedance of the circuit under test as compared with that of the cathode-ray tube deflecting-plate circuit. Figs. 13 and 14 show in a typical case how the current in the deflecting plates varies during different phases of the deflecting voltages.

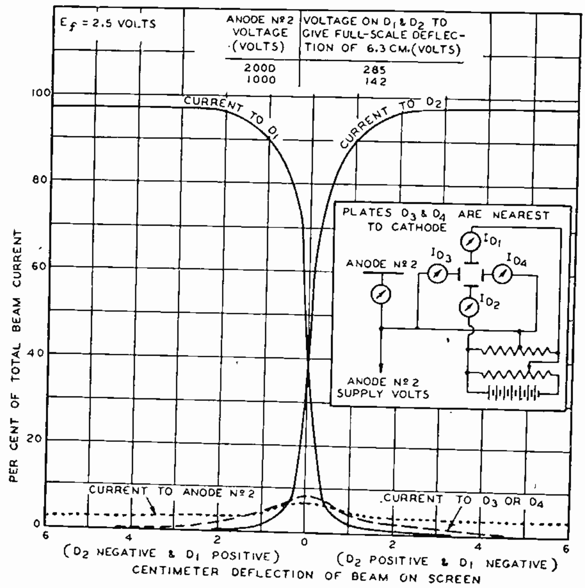

Fig. 13—Deflection impedance characteristics of plates nearest to luminescent screen.

In Fig. 13, the two deflecting plates toward the screen are connected across the center-tapped resistance to be used for the circuit under test. It will be noted that the largest portion of the beam current is collected at the most positive plate. In Fig. 14, the two deflecting plates toward the gun are connected across the center-tapped resistor to be used for the circuit under test. The center tap of the above resistor is connected to the No. 2 anode. Here it will be noted that the impedance of the latter pair of deflecting plates is higher than for the former, especially for small deflecting voltages, because a much larger fraction of the beam current is collected by the pair of plates connected to the second anode. In both cases the second anode collects only a small fraction of the beam current.

The collection of the beam current by the deflection plates which lowers the impedance of this circuit is not an inherent limitation of cathode-ray tubes in general but is only due to lack of shielding of the deflecting plates. For cases where the impedance of deflecting plates is a very important limitation, tubes are made with shielded deflecting plates so that their impedance is high and the beam current is practically all collected by the second anode. The shielding increases the capacitance of the deflecting-plate circuit and, therefore, for very high-frequency work, unshielded deflecting plates are more suitable. There are many different circuit arrangements of the deflecting plates for which different impedance values result. The above figures represent

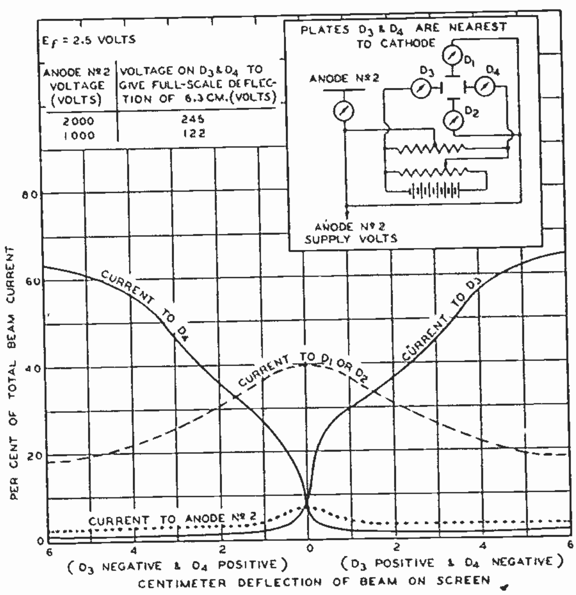

Fig. 14—Deflection impedance characteristics of plates nearest to cathode.

a typical case which may be used to estimate the values for other circuits. It is difficult to generalize on the factors determining the choice of deflection, since each application requires its own analysis. The above discussion may help in shaping the line of attack on these problems.

The factors determining deflection sensitivity of cathode-ray tubes under different conditions of operation should be considered in order to determine the optimum conditions for the sensitivity required for a given application. The electrostatic-deflection sensitivity for a particular pair of deflection plates is inversely proportional to the No. 2 anode voltage and directly proportional to their distance from the luminescent screen. The sensitivity is usually expressed in millimeters deflection of the spot on the screen per volt impressed across the deflection plates.

Magnetic-deflection sensitivity for a particular tube with a particular magnetic circuit is inversely proportional to the square root of the No. 2 anode potential. The magnetic deflection sensitivity is usually expressed in millimeters deflection of the spot on the screen per gauss of magnetic field intensity. Since magnetic-deflection sensitivity is dependent upon the external magnetic circuit and its position along the neck of the tube, the following expression is given to aid in the design and placing of the magnetic circuit.

Magnetic sensitivity is given by the following equation:

$$d/H = (2.98/\sqrt{E_{b_2}})(X_1 X_2 + X_1^2/2)$$

where,

d = deflection in millimeters
H = deflection field intensity in gauss
X_1 = length of pole in centimeters
X_2 = distance from the screen to the nearest edge of the poles
E_{b_2} = second anode potential in volts

This expression is based on the assumption that the magnetic field is uniform and extends only between the pole pieces. From this equation, it becomes apparent that for maximum sensitivity, the poles should be placed as far from the screen as possible along the tube axis. The deflection field should not be applied at the No. 1 anode or closer to the cathode than the No. 1 anode because it may destroy the focusing mechanism. It is, therefore, necessary to apply the magnetic field beyond the No. 1 anode, and it is well to apply this field at a distance equal to the distance between poles to insure that no appreciable stray field penetrates within the first anode.

Conclusion

It may be well to mention a few general items pertaining to the operation of cathode-ray tubes.

The tube may be operated either with the cathode grounded and the No. 2 anode and deflection plates above ground, or with the latter grounded and the cathodes and other electrodes below ground. If the tube is operated with the cathode below ground potential, it is necessary to insulate the filament supply to stand the maximum tube voltage because the voltage difference between the heater and cathode must be not more than a few volts.

The tubes should be electrostatically shielded if the No. 2 anode is run above ground potential and should be magnetically shielded in either case from strong extraneous magnetic fields greater than the earth's field. For weak magnetic fields such as the earth's magnetic field, a small permanent magnet placed at some distance from the tube may be used to center the beam in the tube.

Some of the energy imparted to the fluorescent screen by electron bombardment goes into heat. If an intense beam of the order of a few hundred microamperes is allowed to remain stationary on the screen for a short time, several seconds, local heating may result. If sufficient heat is developed, the screen material may be "burned," resulting in a loss in efficiency of the luminescent screen at that point. It is, therefore, well to have the deflection voltage applied before turning on the cathode-ray beam. The cathode-ray beam may be completely controlled by turning off the No. 1 or the No. 2 anode voltage, or the beam intensity may be varied by varying the negative bias on the control electrode.

Those who are further interested in the fundamental principles upon which the focusing characteristics of cathode-ray tubes are based are referred to the recent book by E. Bruche and O. Scherzer[19] on electron optics.

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LUMINESCENT MATERIALS FOR CATHODE-RAY TUBES*

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Summary—*This paper describes the manufacture, characteristics, and utilization of three kinds of luminescent materials (phosphors) which are employed in cathode-ray tube screens. The phosphors' dependence for efficiency upon minute details of manufacture is brought out, and the subject of activators discussed. The different forms of persistence and spectral distribution characteristics with their corresponding usefulness in the oscillographic and television fields are shown. One material, called phosphor No. 1 is recommended for general-purpose cathode-ray tube screens, particularly for visual observation where the highest brilliancy is desired; another, phosphor No. 2, for work where its brilliant phosphorescence is wanted for observing extremely low-frequency phenomena and transients; and the third material, phosphor No. 5, of very short persistence and highly actinic emission, when photographic records are to be taken of the cathode-ray tube trace.*

THE luminescent screen material, or phosphor, of the cathode-ray tube is a very important part of the tube since it is the means for converting the electrical energy of the cathode-ray beam into light. The efficiency of this conversion depends to a large extent upon the cathode-ray screen material. Work on luminescent materials dates back to the seventeenth century when Bolognese alchemists[1] discovered that barite ($Ba\ SO_4$), with correct treatment, continued to glow after exposure to daylight. In 1764, Canton[1] found that oyster shells heated with sulphur produced the same result. Stokes[1] published the first summary of work done in this field, and Becquerel[1] made the first serious study of these materials. Lenard and Klatt[1] did notable work in 1889 on the manufacture of luminescent materials. A great deal of work on luminescence has been recorded by Nichols and Merritt[2] in 1912 and Nichols, Howes, and Wilber[3] in 1928. During the last few years, however, there has been little published on the accomplishments of those who are striving to obtain the best possible materials and methods of manufacture for the fluorescent screens used in the modern cathode-ray tube. Slight changes in either the screen material or the method of manufacture may result in large changes in efficiency

* Decimal classification: R388. Original manuscript received by the Institute, April 15, 1935; revised manuscript received by the Institute, June 28, 1935.

[1] "Fluoreszenz und Phosphoreszenz" in Wien-Harms' "Handbuch der Experimentalphysik," by P. Lenard, J. Schmidt, and R. Tomaschek, (1928).

[2] E. L. Nichols and E. Merritt, "Studies in Luminescence," published by Carnegie Institution of Washington.

[3] E. L. Nichols, H. L. Howes, and D. T. Wilber, "Cathode-Luminescence and the Luminescence of Incandescent Solids," published by Carnegie Institution of Washington.

of the tube, in the spectral distribution of the light produced by the screen, or in the persistence with which the light continues to emanate from the excited portion of the screen after the electron beam has been removed.

The luminous efficiency of a cathode-ray screen is expressed in candle power per watt. Candle power is a unit of luminous intensity. The luminous intensity in any direction is the luminous flux emitted from a point source in that direction through a unit solid angle, luminous flux being the radiant energy which is visible to the eye. In the case of the cathode-ray screen, the candle power is measured in a direction normal to the screen. The watt is the measure of the electrical energy in the electron beam as determined by the beam current and voltage. The radiant efficiency of the screen, expressed in total radiant flux per watt, is different from its luminous efficiency because the radiant energy in different parts of the spectrum is not equally perceptible to the eye. The spectral distribution of a screen shows the relative amounts of light energy emitted at each wavelength throughout the spectrum. The persistence characteristic of a cathode-ray tube screen discloses its brilliancy variation with the time of phosphorescence (its light emission after excitation); its light emission during excitation is called fluorescence.

Each of these different characteristics has its own importance in the various applications of the cathode-ray tube. Screens which have a short persistence are desired for recording on moving film. For use in viewing transients, or very low-frequency wave forms, a screen having a long persistence is of great benefit since it holds a picture of the trace on the screen long enough to permit easy examination. For still other work, such as television, when it is desired to eliminate flicker and to avoid noticeable blurring of the moving traces, a medium persistence screen is of advantage. There are many separate opinions about the advantage of having a certain spectral distribution, or color, for the viewing screen, but there are definite practical advantages in confining the color to certain spectral bands. Since ordinary photographic films are most sensitive to the blue and ultra-violet, it is highly desirable for making this type of oscillographic records that all the light energy be emitted in this spectral region. However, the greatest visual efficiency is obtained when the light energy output is emitted at a color to which the eye is most sensitive. The average eye is most sensitive to the wavelength of 5550 angstrom units (in the green). If a cathode-ray tube emitted monochromatic light of this wavelength, the brilliancy of the screen would be highest for the same screen conversion efficiency from electrical to radiant energy, and its luminous efficiency would be a maximum. Distributing the energy of light emission over a wider range of colors will decrease the candle power, but may have an advantage for special applications. The luminous efficiency of the screen is also dependent upon its candle power distribution characteristic—the light output of the screen measured at different angles of incidence to the screen. In general, this distribution should not be directional, but should allow good visibility throughout a wide angle of observation. All of the characteristics discussed above have been considered in producing screens for cathode-ray tubes particularly suited for different oscillograph applications.

Of the many materials used for cathode-ray tube screens, the one probably best known is a silicate, synthetic willemite. This is crystalline zinc orthosilicate containing a small proportion of manganese. It is fairly typical of the silicates in particular, and of the so-called "activated" luminescent materials in general: the pure base material (zinc orthosilicate) is not luminescent, but becomes so upon the inclusion in its crystal structure of a small amount of some foreign substance (manganese, in this case), which is called the activator. There is an optimum concentration of the activator for maximum light production efficiency; a large excess of activator produces a material which is completely insensitive to cathode rays. The optimum concentration of a given activator varies for different crystalline bases. Silicates may be formed by heating together suitable proportions of a metal oxide, an oxide of the activator, silica, and a flux. These preparations are usually quite stable and not easily harmed by heating in any atmosphere. A screen using one preparation of this class is identified for convenient reference in this paper as No. 1 phosphor.

A second group of luminescent materials may be classified as sulphides. The term, No. 2 phosphor, is used to identify an example of this type. The sulphides of various bivalent metals can be made luminescent by the presence of activators in the same way as can the silicates; copper is one of the most effective activators. Its optimum concentration, however, is so very small that ordinary "chemically pure" sulphide samples may contain enough copper to suppress the luminescence. The process of manufacture is, therefore, begun by purifying a soluble salt of the metal whose sulphide is to be prepared; this is done to remove the excess of copper and certain heavy metals, such as iron and lead, which render the sulphide nonluminescent. The sulphide may be precipitated from the purified solution. This preparation is mixed with suitable fluxes, sufficient activator is added to bring the total percentage up to the optimum, and the mixture is fused in a controlled atmosphere to prevent oxidation of the sulphide. The product is crystalline and usually more resistant to chemical attack than the corresponding sulphide in the amorphous state.

There is a third group of luminescent materials, including certain tungstates and molybdates, which are generally considered to be self-activated and require the presence of no impurity to produce luminescence. The No. 5 phosphor is an example of this class. It should be remarked that recent evidence[4] indicates that activation of the tungstates may be due to exceedingly minute traces of lead. When luminescent materials were first investigated systematically, the number classified as pure, "self-activated" substances was very large. With later and improved analytical methods, many of these were found to contain traces of foreign substances which were shown to be activators. It is, therefore, not surprising that uncertainty concerning the activation of the tungstates should still exist. The tungstates and molybdates may be formed by the union, at high temperature, of a metal oxide with tungstic or molybdic acid. They are quite stable chemically and certain of them appear to be capable of giving almost unlimited life in a cathode-ray tube.

Many oxides, both bivalent and trivalent, can be activated by some of the rare earth elements and by certain other metals, for example, chromium. A well-known example of this very broad class is the ruby, which is aluminum oxide activated by chromium; its luminescence is of the beautiful red color for which the gem is so highly prized.

The No. 1 phosphor, used in RCA tubes numbers 903, 904, 905, and 906 produces a general purpose screen having high candle power per watt, brilliant green color, good photographic qualities when verichrome or panchromatic type films are used and a medium persistence sufficient to aid the eye in eliminating flicker when determining wave form of certain low-frequency phenomena. The No. 2 phosphor is a long persistence screen whose special quality is its ability to display the trace of the electron beam's path long after it has left the screen. This screen, therefore, finds a wide field where ordinarily the electron beam travels too slowly for the persistence of the eye to visualize the shape of the wave formed by the trace or too fast for the eye to comprehend the detail of phenomena. A short persistence screen is found in the No. 5 phosphor, which is used in cathode-ray tubes, RCA-907 and RCA-908. This screen is well adapted to all photographic work because its spectral distribution lies in the blue and ultraviolet and because its especially short persistence allows the photographing of high-frequency phenomena on film moving at high speed.

The candle power distribution characteristic which affects the lumi-

[4] F. E. Swindells, *Jour. Opt. Soc. Amer.*, vol. 23, p. 129; April, (1933).

nous efficiency of all the screens has been found to depend only slightly upon the electrode voltages of the tube, the particle size of the phosphor, and the method of screen manufacture. With increasing electrode voltages, larger particle size, and thinner screens, this characteristic tends to become more directive along the normal to the screen. Fig. 10 of the Orth, Richards, and Headrick paper[5] illustrates that, in general, it follows the cosine distribution law very closely. This form of distribution, which is often called perfect diffusion, is highly desirable since the brilliancy of the screen is theoretically the same at any viewing angle between zero and ninety degrees from normal.

The outstanding features of the above three phosphors are their differences in persistence. The comparative curves of persistence* for these screens are shown on Fig. 1, where the luminous intensity of each screen at the time of excitation is given an arbitrary value of 100. If these curves were plotted on an absolute scale, the ordinates of curve *A* would be only fifteen per cent of the values shown for curve *B*, while those for curve *C* would be only one per cent of those for curve *B*. Even with these absolute ordinates the area under curve *A* is greater than that for either curves *B* or *C*, which confirms the fact that the luminous efficiency for No. 1 phosphor is greater than for either of the other screens. No. 5 phosphor (*C*) shows no persistence whatever by any method of measurement yet applied to it. Determinations thus far indicate that the luminous intensity of the screen decreases to zero immediately after excitation. However, the smallest measurable time interval in this determination was 0.000027 second and the smallest percentage of the initial brilliance which could be accurately measured was one per cent. Therefore, the conclusion is that the luminous intensity of this phosphor decreases to less than one per cent of its excitation value in less than thirty microseconds. This means that this phosphor will not blur the trace on a moving film which travels at two hundred centimeters per second. Tests have further indicated that, at speeds at which a record can be made on the most sensitive moving film, no detrimental effects will be caused by the persistence of this phosphor. No. 1 phosphor, curve (*A*) shows a definite persistence extending over a period of nearly a fifteenth of a second before its brilliancy has decreased below ordinary visibility. However, in a darkened room, the phosphorescence (a term which refers to the luminous radiation given out after the excitation is removed) is apparent immediately after

[5] Orth, Richards, and Headrick, "Development of cathode-ray tubes for oscillograph purposes," Proc. I.R.E., this issue, pp. 1308–1323.

* The methods used in obtaining the curves representing the persistence and spectral characteristics are discussed by T. B. Perkins, "Cathode-ray terminology," Proc. I.R.E., this issue, pp. 1334–1344.

the electron beam is removed from the screen although the intensity of light is so small as to be of no practical value. No. 1 phosphor aids the eye in visual observations, and due to its high efficiency can be used for photographic recording with all nonmoving film cameras. However, its persistence characteristic causes blurring when moving films are employed. The persistence of No. 2 phosphor as illustrated in curve (*B*) of Fig. 1, is apparently less than that of No. 1 phosphor

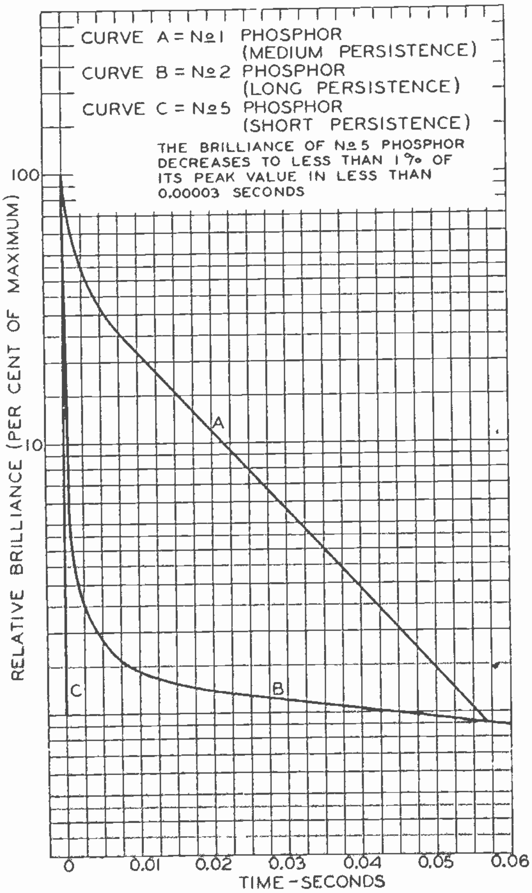

Fig. 1—Persistence characteristics of three cathode-ray tube phosphors.

since its brilliance decreases extremely rapidly in 0.0005 second after excitation and drops to less than ten per cent of its initial brilliance within this time. This fact is of benefit in photographic work; this phosphor can be used with film moving at moderate speed where No. 1 is not suitable. However, after 0.0005 second, the rate of decay of the phosphorescent brilliance is very slight; the actual percentage brilliancy becomes greater for No. 2 than for No. 1 phosphor before a fifteenth of a second has elapsed. It might appear that this persistence would be hard to utilize because of its low percentage of the initial brilliancy.

However, since the instantaneous brilliancy of the fluorescent spot for No. 1 phosphor is approximately fifteen per cent of that for No. 2 phosphor, at times equals zero, the absolute brilliancy of No. 2 is quite appreciable well after one fifteenth of a second. The persistence of No. 2 phosphor is illustrated better by Fig. 2, which shows this characteristic for three intervals of time, namely, 0–0.05, 0.05–0.5, and 0.5–10

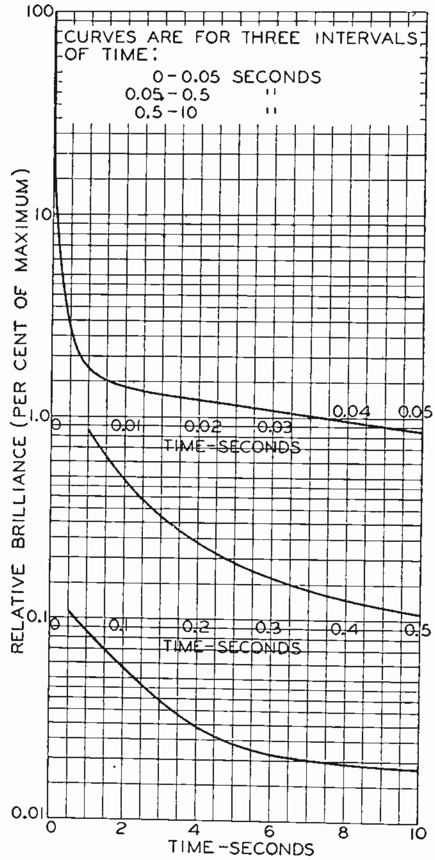

Fig. 2—Persistence characteristics of No. 2 phosphor. This phosphor is used in cathode-ray tubes where a highly persistent trace is desired.

seconds. Here it is seen that the brilliancy of this screen's phosphorescence is measurable as long as ten seconds after excitation. In a darkened room phosphorescence can be observed a number of minutes after excitation but its brilliancy is small. In interpreting these curves, it must be borne in mind that a certain time is required for the phosphorescence brilliancy to build up so that, if the screen is excited for too short a period, the phosphorescent brilliancy will be much lower than that obtained when the screen is excited to saturation. Saturation is reached in a shorter time when a more intense excitation is applied to the screen. It is, therefore, advisable when transients of a very high speed are to be viewed that a tube capable of putting the highest possible wattage into the electron beam be utilized. Repetition of excitation on the same portion of the screen will sometimes increase the persistence brilliancy. This is only true if saturation has not been reached before re-excitation and if the time interval between excitation and re-excitation is not too great. Therefore, high speed recurrent patterns will persist longer than nonrecurrent ones.

The spectral distributions of the light emitted from No. 1, No. 2, and No. 5 phosphors are shown in Figs. 3 to 5. These curves are all drawn to relative energy scales of the same units but to different

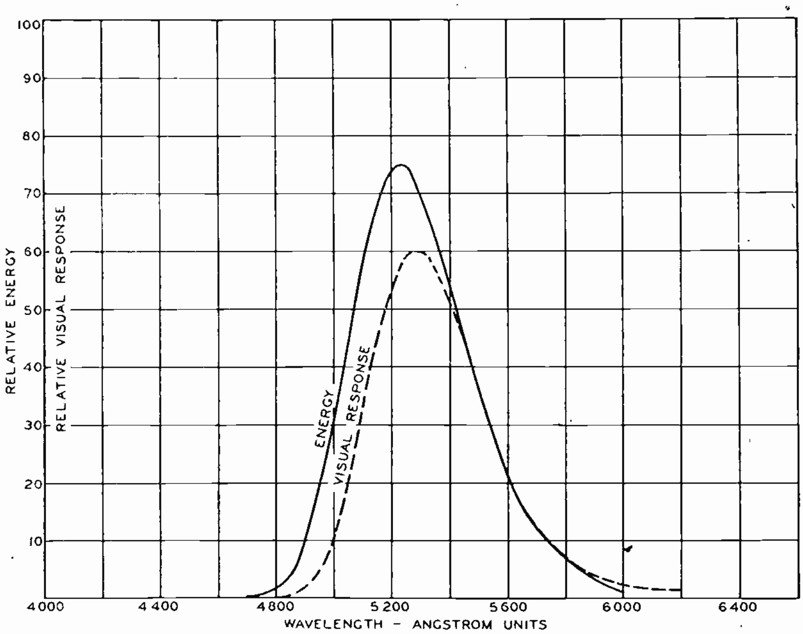

Fig. 3—Spectral characteristics of No. 1 phosphor. This phosphor is used in cathode-ray tubes where high visual efficiency is desired.

magnitudes to illustrate better their several shapes. The general purpose No. 1 phosphor (illustrated in Fig. 3) proves to have a spectral characteristic whose peak is near the sensitivity peak of the eye. Its general shape is of the eye visibility curve but it is more narrow, cutting off more sharply in the blue and red. This form of spectral curve is very nearly ideal as can be seen by the corresponding visual spectral curve, shown dotted in Fig. 3, whose ordinates are obtained by multiplying the values on the spectral curve by the percentage sensitivity values of the average eye for each wavelength. The ratio of the areas under the two curves, illustrated in Fig. 3, or the ratio of the total energy as evaluated by the eye to that emitted in the visible spectrum, called the utilization factor, is approximately 0.68 which shows that only thirty-two per cent of the visible light emitted from the screen is not utilized to its fullest advantage by the eye. This high utilization factor is one of the reasons that No. 1 phosphor has a higher luminous efficiency than either of the other two screens. The average luminous efficiency of No. 1, under certain definite conditions, is two candle power per watt. The color of No. 2 is a bluish white in contrast with the green of the No. 1 phosphor. The spectral characteristic of No. 2 phosphor, as illustrated by Fig. 4, is not only more peaked in the blue, than for No. 1, but it has a far more irregular shape and gives emission throughout the entire visible spectrum. No. 2 phosphor curve shows three

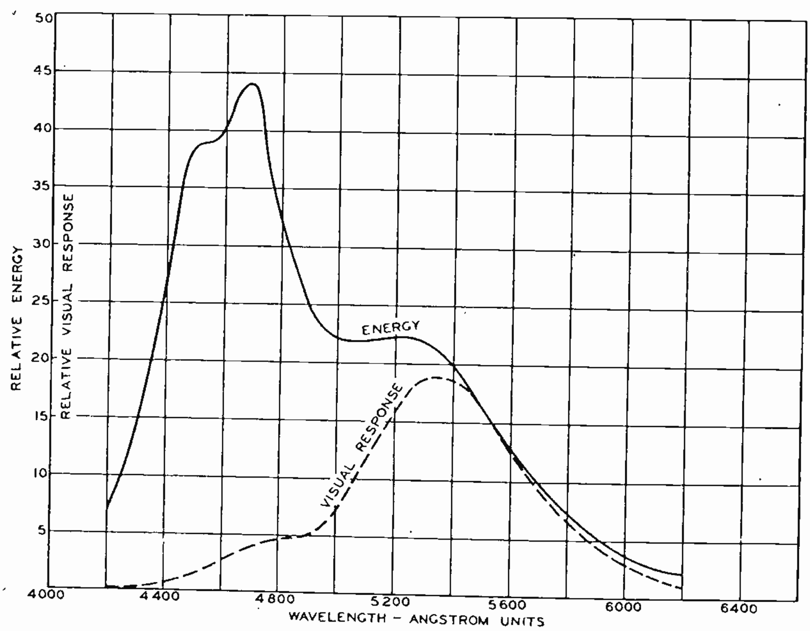

Fig. 4—Spectral characteristics of No. 2 phosphor. This phosphor is used in cathode-ray tubes where a brilliant phosphorescent trace is desired.

peaks, one in the green close to that of No. 1 and two in the blue beyond No. 1's low wavelength cutoff.[6] As might be expected from its high blue peak, the green peak energy value of No. 2 phosphor is far below that of No. 1. Because the eye is quite insensitive in the blue, the No. 2 visibility curve is reduced considerably below that for No. 1. The utilization factor for No. 2 is approximately 0.36, which contributes strongly to the fact that the luminous efficiency of this screen is about one fifth of that of No. 1. This decrease in efficiency is also due in part to the fact that the brilliance of the phosphorescent spot in phosphor No. 2 decreases much more rapidly than that for

[6] No. 1 phosphor does not actually cut off at 4800; it continues to emit weakly into the ultra-violet; however, this light is so weak that 4800 is usually considered the practical cutoff.

No. 1. The spectral characteristic of No. 2 phosphor's phosphorescence does not contain the blue peaks but retains the green portion of the curve, so that the color of this screen's persistence is approximately the same as that for No. 1. The limitations of the spectrophotometer system of measuring spectral intensitives gives a partial characteristic for No. 5 phosphor. As shown in Fig. 5, the light energy emitted from this screen increases towards the ultraviolet and very likely has a peak in the ultraviolet although the glass transmits negligible energy below 3200 angstrom units. This characteristic shape accounts for the violet hue of No. 5 phosphor fluorescence. Its high energy in the violet

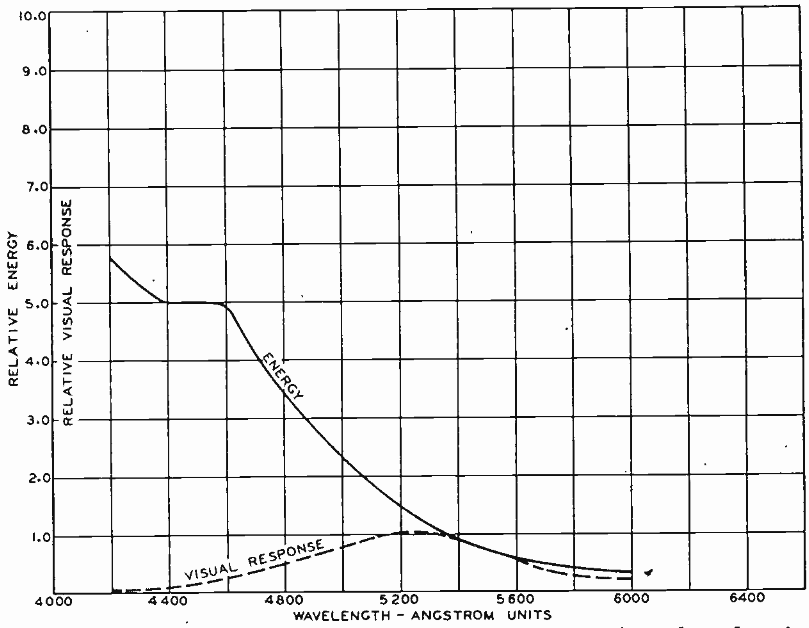

Fig. 5—Spectral characteristics of No. 5 phosphor. This phosphor is used in cathode-ray tubes where a high photographic efficiency is desired.

end of the spectrum where the eye is very insensitive predicts a low luminous efficiency rating for the screen. The utilization factor which covers only the visible spectrum as calculated for the areas under the two curves of Fig. 5 amounts to 0.2. However, if the values of relative energy in the ultraviolet were considered, in the utilization factor, it would be actually far less. This statement is borne out by the fact that the luminous efficiency of No. 5 is one tenth that of No. 2 phosphor. On the other hand, the candle power of this phosphor is of very little importance since its ultra-violet emission is more useful for photographic work than its visible emission. If photographic utilization factors for ordinary film were calculated it would be found that the order of utilization would be reversed. No. 5 phosphor close to one, No. 2 next, and No. 1 least.

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CATHODE-RAY TUBE TERMINOLOGY*

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Summary—*This paper offers tentative definitions for many terms commonly used in cathode-ray tube work. The terms defined are magnetic deflection sensitivity, electrostatic deflection sensitivity, gun-current efficiency, electron gun, screen, screen luminous efficiency, screen actinic efficiency, screen radiant efficiency, spectral characteristic, relative radiant energy, color characteristic, visual spectral characteristic, actinic spectral characteristic, persistence characteristic, relative brilliance, candle power distribution characteristic, luminescence, phosphorescence, fluorescence phosphor, defocus, spot distortion, pattern distortion, spot size, spot diameter, line width, apparent spot size, apparent spot diameter, and apparent line width. The methods of measurement of gun-current efficiency, screen luminous efficiency, candle power, screen, actinic efficiency, screen radiant efficiency, spectral characteristic, persistence characteristic, candle power distribution characteristic, spot size, spot distortion, and pattern distortion are also included for clarifying purposes.*

SINCE the advent of the cathode-ray tube in the industrial field new terms or newly applied terms have been used, the definitions of which have never been established. It is, therefore, the purpose of this paper to offer some tentative definitions of the terms most commonly applied to the cathode-ray tube. In the preparation of these definitions the author has profited from the information gained from many publications[1–9] and from conversations with his associates in his company. Through the courtesy of the I.R.E. subcommittee on cathode-ray tubes these definitions have been tentatively approved as not being in conflict with their proposed standards.

Deflection Sensitivity

In a cathode-ray tube, the deflection sensitivity is a very important rating. It is expressed in two ways which depend upon the kind of field used for deflecting the electron beam.

Magnetic Deflection Sensitivity. The magnetic deflection sensitivity of a cathode-ray tube is the ratio of the distance which the electron beam moves across the screen to the change in the flux density producing the motion. This sensitivity may be expressed in millimeters per gauss, but due to the difficulty of determining flux density, it may often be more practical to express it in millimeters per ampere-turn or merely millimeters per ampere. The sensitivity varies with different forms and positions of the magnets; it is affected by the width of the

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[1] Numbers refer to Bibliography.

magnet poles, the leakage flux, and the distance from the field axis to the screen. It varies inversely as *the square root* of the beam voltage.

Electrostatic Deflection Sensitivity. The electrostatic deflection sensitivity is the ratio of the distance which the electron beam moves across the screen to the change in potential difference between the deflection plates. This deflection sensitivity is usually expressed in millimeters per volt. It varies with the deflection plate dimensions and positions but since these factors are generally fixed by the construction of the tube, they do not have to be considered in the tube sensitivity rating. This sensitivity will vary inversely as *the first power* of the beam voltage.

Efficiency

The cathode-ray tube's efficiency is rated in terms of gun and screen efficiency. The *electron gun* comprises all the tube's electrodes which contribute to the generation and focus of the electron stream, while the *screen* is that portion of the tube where the electron beam produces visible and near-visible radiation.

Gun-Current Efficiency. The electron gun is considered perfectly efficient when all of the initial cathode current reaches the screen. In other words, the gun-current efficiency of a cathode-ray tube is the ratio of the beam current to the current which leaves the cathode. *Beam current* is the current in the electron beam at the screen, and is usually measured in microamperes. When both evaluations of current are made in the same units, the efficiency is readily expressed in per cent.

Screen Luminous Efficiency. The screen's efficiency can be expressed in many different forms. The most generally used rating is the screen luminous efficiency, which is a measure of the screen's ability to produce visible radiation from the electrical energy of the beam. Since the lumen is a measure of the rate of passage of radiant energy evaluated in luminous units, this efficiency should be measured in lumens per watt. However, for the sake of ease in measurement, this is usually expressed in candle power per watt, as candle power is a measure of the luminous flux per unit solid angle in a given direction and can be converted to lumens whenever the candle power distribution characteristic of the screen is known. When the cathode-ray tube is used for photographic purposes, however, the luminous efficiency rating of the screen is not sufficient.

Screen Actinic Efficiency. The screen actinic efficiency is a measure of the screen's ability to convert the electrical energy of the beam to radiation which affects a certain photographic surface. This term should be expressed in microwatts per watt but may often be expressed for ease of measurement in terms of actinic power per watt relative to a well-known screen. Neither of the above efficiencies express the over-all efficiency of the screen.

Screen Radiant Efficiency. The screen radiant efficiency is a measure of the screen's ability to produce luminescence from the electrical energy of the beam. This is a measure of the total radiation produced at the screen irrespective of the spectral sensitivity of the measuring device. This should be also expressed in microwatts per watt, but due to the difficulties of making absolute measurements will be more often expressed in radiant energy per watt relative to some well-known screen. Since the screen efficiency of the cathode-ray tube varies with the electrode and screen conditions, as well as the direction of utilization, these factors should always be stated unless it can be assumed that the measurement was made under optimum conditions.

Characteristics

Since vacuum tube characteristics are familiar to anyone in the radio field and since cathode-ray tube characteristics involving electrode voltages hold closely to the forms used in vacuum tube data, they are really self-explanatory. On the other hand, cathode-ray tube characteristics which describe various functions of the screen alone are relatively new.

Spectral Characteristic. The spectral characteristic of the cathode-ray screen shows the relation between the radiant energy per element of wavelength and each wavelength of the spectrum. It is generally shown in a curve plotted with relative radiant energy against wavelength in angstroms, microns, or millimicrons. *Relative radiant energy* is expressed in arbitrary units of radiant energy. This characteristic may be expressed in different ways, such as, color, visual spectral, and photographic spectral characteristics.

Color Characteristic. When the spectral characteristic of the cathode-ray tube screen covers only the part of the phosphor's total emission spectrum which lies within the visible spectrum, this characteristic will more accurately be called the color characteristic.

Visual Spectral Characteristic. The visual spectral characteristic of the cathode-ray tube screen shows the relation between the luminous energy per element of wavelength and each wavelength of the spectrum. It is generally shown in a curve plotted with relative luminous energy against wavelength in angstroms, microns, or millimicrons. *Relative luminous energy* is obtained by multiplying the relative radiant energy values for each wavelength from the screen's spectral characteristic by the relative response of the eye at that wavelength.

Actinic Spectral Characteristic. The actinic spectral characteristic of a cathode-ray screen shows the relation between the energy per element of wavelength which affects a certain photographic surface and each wavelength of the spectrum. It is generally shown in a curve plotted with relative actinic energy against wavelength in angstroms, microns, or millimicrons. *Relative actinic energy* is obtained by multiplying the relative radiant energy values for each wavelength from the screen's spectral characteristic by the relative sensitivity of a given photographic surface at that wavelength. In the more comprehensive data, the units are always the same for each screen's characteristic but data are sometimes given with the maximum point on each curve at the same energy level.

Persistence Characteristic. The persistence characteristic of the cathode-ray screen is a relation showing the brilliance of light emitted by the screen, as a function of the time after excitation. This characteristic is generally shown in a curve where relative brilliance as the ordinate is plotted on a logarithmic scale against time on a linear scale. *Relative brilliance* is used to describe readings of luminous intensity per unit area evaluated in arbitrary units. On account of the nature of the persistence characteristic and its measurement, the brilliance of the screen at the time of excitation is often given the arbitrary value of 100 on each curve; however, it is being felt more and more strongly that these brilliance units should at least be identical for each characteristic if not expressed in absolute units of candle power per square millimeter.

Candle Power Distribution Characteristic. The candle power distribution characteristic is a relation which when plotted is invariably represented by a polar curve; this illustrates the luminous intensity of the cathode-ray tube in a plane of the axis of the tube and with the screen at the origin. This characteristic shows how the candle power of the luminescent screen varies when viewed at different angles to the screen. Unless the above characteristics are known not to vary with change in electrode voltage and screen conditions, these details should be recorded with each characteristic.

Luminescence

There are three terms which have long been used quite loosely and interchangeably but have really definite and separate meanings of their own. These are luminescence, fluorescence, and phosphorescence.

Luminescence. Luminescence covers all forms of visible and near-visible radiation which depart widely from the black-body radiation law. It can be divided according to the means of excitation into many classes such as candoluminescence—the luminescence of incandescent solids; photoluminescence—the luminescence created by exposure to radiation; chemiluminescence—the luminescence created by chemical reactions; electroluminescence—the luminescence given off by ionized gas; bioluminescence—the luminescence emitted by living organisms; triboluminescence—luminescence created by the disruption of crystals; crystalloluminescence—luminescence created by formation of crystals; radioluminescence,—luminescence excited by emissions from radioactive materials; galvanoluminescence—the luminescent phenomena observed at electrodes during some electrolyses; cathodoluminescence —the luminescence produced by the impact of electrons, etc. In cathode-ray tubes we deal principally with cathodoluminescence, and therefore, the luminescence of the screen refers to that radiation which is produced by the electron beam. Luminescence may also be divided into two classes according to its time of emission.

Fluorescence. Fluorescence is that luminescence emitted by the phosphor during excitation, which in a cathode-ray tube refers to the radiation emitted by the screen during the period of beam excitation.

Phosphorescence. Phosphorescence is the luminescence emitted after excitation, and in a cathode-ray tube this is the radiation which persists after the electron beam excitation has ceased.

Phosphor. The solid material in the screen which produces luminescence when excited is called the phosphor.

Luminescent Spot

When the focused electrons hit the cathode-ray screen, a luminescent spot is formed. A defocused beam will cause what is known as spot or pattern distortion.

Defocus. Defocus may be defined as any condition of the spot other than optimum with respect to size and shape.

Spot Distortion. Spot distortion is any condition of the spot other than optimum with respect to shape, irrespective of size.

Pattern Distortion. When the electron beam is moved by changing fields a pattern is formed and the wave form of the spot movement will be identical with the resultant wave forms of the electrical phenomena producing these fields unless there is pattern distortion present. This distortion takes many forms such as amplitude, frequency, phase, brightness persistence, spot size, etc.

Spot Size. Spot size may be measured under various conditions, and is commonly designated by such names as spot diameter or line width. When the spot is stationary, its width can be measured in any direction, but is usually determined by the width of the spot along its longest and shortest axes.

Spot Diameter. In the case of the undistorted spot, two designations of spot size are not necessary, but the term spot diameter can only be applied to a circular spot.

Line Width. In the pattern, spot size is usually referred to as line width, or line width may be defined as the width of the moving spot measured at right angles to the direction of motion. The above definitions of spot size refer to the width of the spot regardless of the method of measurement. It is true, however, that most cathode-ray luminescent spots are not of uniform brilliance; usually the brilliance of the spot decreases towards its edges.

Apparent Spot Size. Apparent Spot Diameter. When the spot size is measured visually or from a photographic record, the resultant spot size is not necessarily the true spot size and should be called the apparent spot size or apparent spot diameter.

Apparent Line Width. The apparent line width, the visible or recorded width of the moving spot, may be different from the apparent spot size of the stationary spot because screen luminescence is dependent upon the duration of excitation.

Appendix

On Methods of Measurement

Some of the suggested definitions may be made more clear by describing their method of measurement.

Gun-Current Efficiency. In cathode-ray tubes where the high voltage electrode current is a measure of all the secondary emission from the screen rather than primary electrons to this electrode, the gun-current efficiency may be measured by the ratio of the high voltage electrode current to the cathode current. In some cathode-ray tubes where the greater part of the high voltage electrode current is direct current from the cathode, other means of measuring this efficiency will have to be employed.

Screen Luminous Efficiency. The luminous efficiency of the screen is measured by determining the horizontal candle power of the cathode-ray tube in the direction normal to the screen at a given high voltage electrode potential and current, whose product may be used as a measure of the beam energy. There are several methods of measuring candle power, the most important is that in which the inverse square law of illumination is used.

Candle Power. The candle power of a cathode-ray tube is determined by measuring the illumination received in a normal direction from the scanned luminescent screen at a distance where the inverse square law is known to hold. This may be done by means of any form of photometer[4,10] or illuminometer.[4,11] Since the luminescence from cathode-ray screens has varied spectral characteristics which are far different from that of the standard generally used, some form of heterochromatic photometry[4] must be used. The most direct method is one using the flicker photometer[4,12] for the measurement of illumination, but one of equal accuracy and measurement ease is that using the color filter[13] to correct the standard to the color of the luminescent screen. The latter method requires an accurate method of determining the transmission[4] of the filter employed for the measurements.

Screen Actinic Efficiency. The most direct method of measuring screen actinic efficiency is to compare the density[14] produced on a photographic surface by the luminescent screen excited by a certain beam energy on the same surface with that produced by a standard source of radiation.[15] The standard must have the same actinic characteristic as that of the cathode-ray screen. The difficulties of obtaining a standard source of radiation, correcting it to have the same photographic spectral characteristic, and of taking the densitometer measurements, make relative actinic power per watt measurements more practical. This is done by comparing the beam energy required to produce a given density on a given photographic surface with that required by some well-known screen to produce the same density. These density measurements are usually made at some low density and compared by eye with the densities at juxtaposition, or by a simple form of densitometer. Either of these methods must use identical optical conditions, for comparison between tube and standard and measurements must be taken in a definite direction and under cathode-ray tube conditions stated in the resultant efficiency.

Screen Radiant Efficiency. The screen radiant efficiency can be measured directly with a calibrated thermopile[16] or similar radiant energy sensitive device[17] of sufficient spectral response by utilizing the inverse square law of radiation. However, the more practical method uses this radiant energy sensitive device merely for comparing different screens in relative radiant energy per watt units. This method requires that the measuring device not only cover the entire radiant-energy emission spectrum of all cathode-ray screens, but that it be uniformly responsive throughout this range, unless the spectral characteristics of both the cathode-ray tube and measuring device are known accurately and compensations in the calculations can be made. However, when the spectral characteristic of the cathode-ray tube screen is known, the area under the spectral curve represents the total radiant energy emitted by the screen. Therefore, measurement of the areas under spectral curves using identical units, either by graphical or mathematical methods can be used as relative radiant energy values. Similarly, the areas under visual and photographic spectral characteristic curves drawn to the same units can be used as relative luminous energy and relative actinic energy values, respectively.

Spectral Characteristic. The spectral characteristic of a cathode-ray tube screen is taken by comparing the amount of radiant energy emitted from the screen at each element of wavelength with that emitted from a standard source of radiation. The spectra of both cathode-ray screen and standard source are divided into spectral bands by means of a spectrometer or monochromator, and are compared in intensity by means of some form of radiant energy sensitive device. When the eye is used with a spectrophotometer[4,18] only the visible part of the spectrum is measured; however, by means of a phototube or thermopile, a wider spectral range can be covered. When it is desired that each spectral characteristic use the same units, the apparatus must be set up in exactly the same manner for each measurement and the beam energy and screen scanning conditions must be identical. The standard should be checked also to give the same energy level and distribution before each test.

Color Characteristic. Visual Spectral Characteristic. Actinic Spectral Characteristic. The color, visual spectral, and actinic spectral characteristics are derived from the spectral characteristic as described in their definitions. It is possible to take such characteristics directly. For example, the spectrograms of the cathode-ray tube screen's emission may be compared to that of a standard radiation source by means of a densitometer. However, such a method is so complicated and difficult as to make it rarely practical.

Persistence Characteristic. The persistence characteristic is taken by the use of some form of stroboscopic apparatus[2] by means of which one is able to measure the relative brilliance of the screen at definite times after excitation. One method of accomplishing this involves the use of a stroboscopic disk with a single rectangular aperture or slit geared to a synchronous motor. The width of the slit and speed of the disk govern the time interval which can be covered by a single measurement and that covered by the entire test. Thus, if the slit is located four inches from the center of the disk, the slit width is one sixteenth of an inch, and the gear reduction is two to one from a motor whose speed is 1800 revolutions per minute, the slit width covers an interval of time equal to 0.00017 second while the total time covered by the test can only equal the time of a single rotation of one fifteenth of a second. When the field of the motor can be rotated with respect to the armature the phase or position in time of the slit can be determined by the amount of this rotation. Thus the number of degrees through which the field turns to complete one revolution of the disk divided into the interval of time covered by the disk's revolution represents the interval of time covered by moving the field one degree. The cathode-ray tube screen is scanned and the pattern synchronized so that the trace recurs once and only once for every rotation of the disk. This can easily be accomplished by a synchronized sweep-circuit voltage in one direction of deflection and a spreading voltage equal to or a multiple of the sweep voltage in the other direction. Now a persistence characteristic can be measured by a brilliance determination at each field setting by means of a photometer, or illuminometer, as often as there are slit-width intervals in time of disk rotation, provided the field rotation is calibrated finely enough. This method does not allow accurate measurements over a time greater than one fifteenth of a second, since an objectionable flicker is obtained from the luminescent spot when observed at intervals greater than one fifteenth of a second. Therefore, another method utilizing a traveling slit which follows the luminescent spot in its travel across the tube, so as to obtain a continuous brilliance from the screen at all speeds may be used. This is accomplished by utilizing an opaque film having slits cut at distances equal to the travel of the luminescent spot on the screen. Now the speed of the film and spot are synchronized mechanically by generating the voltage which sweeps the spot across the screen from a potentiometer geared to the film pinions. The film speed controls the test interval and the phase difference is determined by a calibration of the potentiometer stator rotation. As in the spectral measurements, if this characteristic is desired to be taken in identical units for each screen, then the apparatus and screen must be operating under the same conditions for each test.

Candle Power Distribution Characteristic. The candle power distribution characteristic measurement involves the same method of measuring candle power as described above, except that in this determination the illumination from the screen is measured not only in the normal direction but at all angles to the screen. In this work the scanned area of the cathode-ray screen is made as small as possible and placed in the exact center of the screen in order that the curvature of the bulb might not give an erroneous value to the candle power measurement at right angles to the axis of the tube.

Spot Size. Spot Diameter. Line Width. The spot size, spot diameter, and line width for the cathode-ray screen may best be measured by an optical micrometer[19] which is calibrated in microns and has a linear magnification from one to twenty-five and a field of view of at least two millimeters. Due to nonuniform brilliance in the luminous spot the apparent spot size will vary somewhat with the magnification power of the micrometer; therefore, for accurate comparison of data this factor should be stated with the measurement results.

Spot Distortion. Spot distortion may be measured by determining the amount that the spot size varies from symmetry by means of the optical micrometer. This should be expressed in per cent distortion or the ratio of the actual spot size in the given direction to that size which the spot would have if it were symmetrical.

Pattern Distortion. Pattern distortion may be measured by many different methods, the use of which depends upon the form of distortion and applied wave form present. No matter what form of distortion is measured, however, it should be stated in per cent distortion, or the ratio of the actual condition of the pattern to the optimum condition. For example, when the amplitude pattern distortion is measured, the amplitude of the cathode-ray trace due to the applied voltage is measured in a given direction over each portion of the screen. Then the amplitude distortion of the tube is determined by the percentage change in amplitude response of the cathode-ray tube at each portion with respect to its response with the trace in the center. Methods for measuring all forms of pattern distortion have not been established, but they will all follow the same general principle of measuring the output of a cathode-ray screen as compared with that which is optimum for the working condition.

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RADIOMETEOROGRAPHY AS APPLIED TO UNMANNED BALLOONS*

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Introduction

WIDESPREAD surface observations, taken at the same hour, have long been the basis of synoptic meteorology. Modern methods of weather analysis and forecasting require, in addition, synoptic data on the extent, motion, and characteristics of air masses within which and between which the weather takes place. If a complete synoptic picture is desired, vertical soundings up to four or five kilometers must be made twice daily, or oftener, at stations separated by not more than a few hundred miles; at each station the desired data include the direction and velocity of upper air winds, and the vertical distribution of pressure (p), temperature (t), and relative humidity (f). In some synoptic situations, and in many research problems, it may be desirable to extend observations far above the five-kilometer level.[1]

Present sounding methods have definite limitations. Complete pilot-balloon observations require clear weather and good visibility. Sounding balloons, while independent of weather, are slow and unreliable. In levels below five kilometers airplane sounding, being more rapid and far more certain, is usually preferable to balloon sounding. But the airplane cannot operate safely in unfavorable weather, when soundings are most urgently needed; and above five kilometers, which is a practical ceiling for ordinary airplanes, free balloons are necessarily used.

Within recent years the application of radio technique to meteorological balloons has developed new sounding methods which may largely transcend the limitations indicated above. In fair weather or foul, radio transmission can convey instantly to the surface observing station either the balloon's position, or the air characteristics being encountered at any moment during its ascent, or both. In practice either the simple pilot-balloon or the meteorograph-equipped sound-

* Decimal classification: R553. Original manuscript received by the Institute, December 10, 1934; revised manuscript received by the Institute, May 24, 1935. This paper is an abridged version of that published in *Monthly Weather Review*, vol. 26, pp. 221–226; July, (1934).

[1] Important military applications of radiometeorography are the determination of ballistic wind and ballistic density for artillery use.

ing balloon is combined with a small radio transmitter light enough to be conveniently lifted by the balloon. In the case of a radio pilot-balloon, the radio transmitter is excited either intermittently or contінously during the ascent, while bearings are taken on it by one or more directional radio receivers on the ground. In the case of the radio sounding-balloon, the meteorograph continuously indicating p, t, and f as the balloon ascends is caused so to vary some element of the radio transmission, or to interrupt the transmission in such a way, that nearly simultaneous records of p, t, and f during the ascent are converyed to a receiver on the ground. A radio sounding-balloon can also serve simultaneously as a radio pilot-balloon.

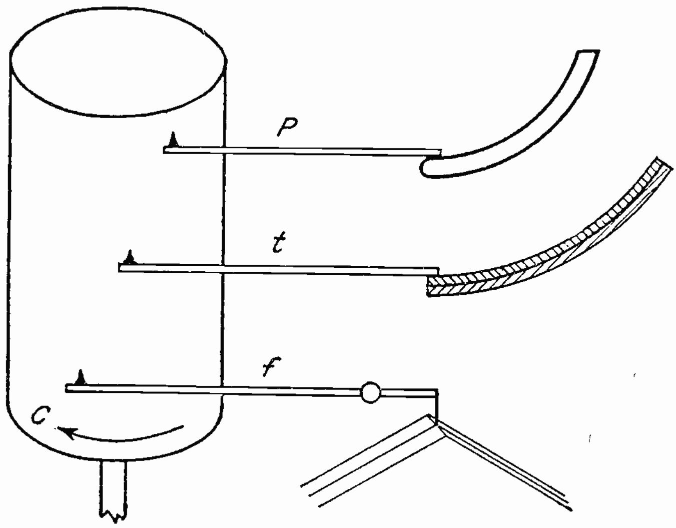

Fig. 1—Standard meteorograph principle.

Historical

Standard Meteorograph

The standard three-element metoerograph, recording p, t, and f as simultaneous ordinates against time as abscissa, was available in completed form before research on the radiometeorograph began. Its basic elements are shown in Fig. 1. C is a cylinder revolved at a suitable rate by clockwork; p is the pressure arm, activated through suitable levers by either a sylphon vacuum chamber or a Bourdon vacuum tube; t is the temperature arm, activated by a bimetallic strip; f is the relative humidity arm, activated by a multiple strand of human hairs. One or more elements of the standard meteorograph have been included in every radio-sounding balloon.

[2] A fairly complete bibliography on radiometeorography up to 1933 is given in a paper by the writer in the *Monthly Weather Review*, vol. 62, p. 221; July, (1934).

Radio Design

Balloon radiometeorography became fully practicable when small, efficient, short-wave transmitters were developed in the radio field. Thereafter, radio design by the various investigators in telemeteorography proceeded along similar and conventional lines. The most efficient generator of stable electric oscillations, and hence of receivable radio waves, is a single triode electronic tube connected to suitable external inductance and capacitance, with or without electromechanical (crystal) stabilization of frequency.

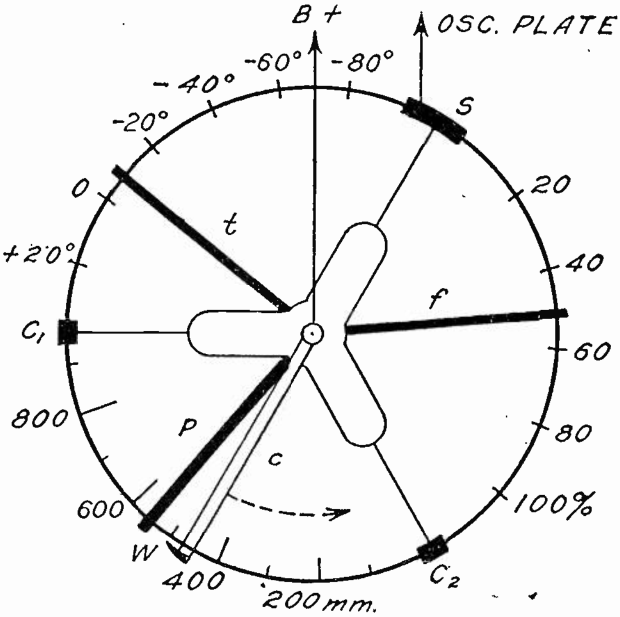

Fig. 2—Olland telemeteorograph principle.

In order that appreciable amounts of radio-frequency energy may be radiated toward the receiver, the oscillator is suitably coupled to an antenna, which may be a single wire one-half wavelength long, with the entire transmitter suspended at its center. Wavelengths so far used for radio sounding-balloons have ranged from 20 to 150 meters (15,000 to 2000 kilocycles).

Telemeteorography

Outstanding among the pioneers of telemeteorography was Olland, a Dutch instrument maker. About 1875 he invented a system for the electrical indication of one or more meteorological elements. The same principle is used today in modern radiometeorography (Fig. 2). It embodies: (*a*) indicating arms, arranged on a common center, which move radially in response to meteorological changes, one indicating arm serving for each meteorological element; (*b*) fixed reference marks on a circle, between (or in some definite relation to) the indicating arms and their limits of movement; (*c*) a revolving contact arm, pivoted concentrically with the indicating arms, and making electrical contact with them and with the fixed reference marks in turn during its clockwork-driven progress around the circle.

Another method was developed more recently by R. Bureau in France. The principle of Bureau's apparatus is shown in Fig. 3. *C-I* is a cylinder formed partly of conducting material (*C*) and partly of insulating material (*I*), and rotated by clockwork. *M* is an eccentric cam, rotated at a much faster rate than *C-I*. *A* is an arm indicating temperature, pressure, etc., by the vertical position of its contact point *P*. When *P* is over the conducting portion of the cylinder, dots are transmitted as the cam *M* makes and breaks contact. When the point *P* is over the insulating portion of the cylinder, no dots are transmitted.

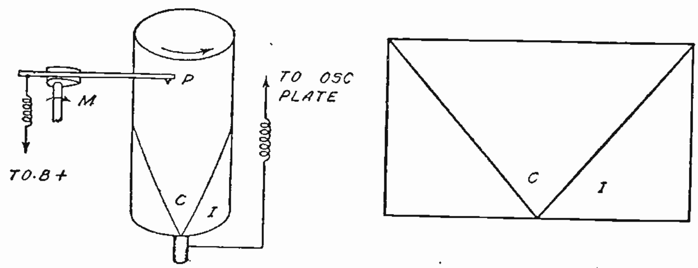

Fig. 3—Bureau radiometeorograph principle.

The number of dots in a series therefore indicates the vertical position of *P*, and hence the temperature, pressure, etc. The system is not limited to the indication of one meteorological element; temperature, pressure, etc., arms can be connected into the circuit in turn by switching cams. For temperature soundings this device gave an accuracy of ± 0.7 degree between $+20$ and -60 degrees centigrade.

A third method was developed by P. Duckert in Germany. Fig. 4 shows the principle used by Duckert for continuous indication of temperature. The bimetallic strip *T*, by means of the lever connection *A*, varies the capacity of the condenser *C* in the oscillating circuit, which in turn varies the emitted frequency. A similar principle was used by Blair in the United States.

Duckert used wavelengths between 30 and 60 meters. The accuracy of his apparatus was ± 0.2 degree for temperature and ± 1.5 millimeters for pressure. In addition to the more practicable temperature-varied condenser system, he devised a system of continuous temperature indication based on the fact that an oscillating crystal changes its frequency with temperature. In addition, he developed lightweight transmitters and improved methods of thermal insulation, such as enclosing the entire transmitter in a glass vacuum bulb. Duckert's latest radiometeorograph is manufactured commercially by the Telefunken Company in Germany.

The original method of Olland was applied to sounding balloons by P. Moltchanoff in Russia and Germany. His radiometeorograph, now manufactured commercially by Askaniawerke in Germany, is shown in Fig. 2. This device keys the transmitter in such a way that the time intervals between dots and dashes, which can be automatically recorded, give indication of pressure, temperature, and relative humidity in turn. In Fig. 2, p, t, and f are concentrically pivoted and move radially, within the limits of their respective scales, in response to pressure, temperature, and humidity. Separating the p, t, and f scales are the fixed contacts c_1 and c_2 and the synchronizing contact S. p,

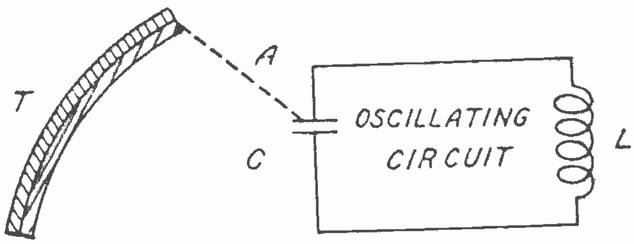

Fig. 4—Duckert temperature indicating principle.

t, f, c_1, c_2, and S are all electrically connected to the oscillator plate. Concentrically pivoted with p, t, and f, but insulated from them, is the contact arm C, rotated by clockwork at about two revolutions per minute and electrically connected to $B+$. In the course of C's revolution, the platinum wire W contacts p, c_2, f, S, t, and c_1 in turn, transmitting a dot for each contact except S, which results in a dash. The time interval between a fixed contact such as c_1 and a movable contact such as p indicates the pressure, etc., being encountered by the device at the time. The entire radiometeorograph, which operates on wavelengths between 25 and 100 meters, weighs 1.4 kilograms. Reception and evaluation are best accomplished by a standard facsimile receiver, whose recording cylinder is synchronized with the contact arm C. Curves having time (altitude) as abscissa and p, t, and f as ordinates are thus directly recorded as the sounding is made.

In Finland Väisälä, in the course of radiometeorograph research, devised some notably light batteries of lead-acid type. His four-volt, 15-ampere-minute A battery weighed 38 grams; and his 72-volt, one-ampere-minute B battery weighed 72 grams.

Possibilities of Improvement in Radiometeorography

Direction of Future Development

Present models of the radiometeorograph give satisfactory performance. Future development will probably be directed chiefly toward lightness, simplicity, and low cost. For radiometeorographs operating up to six kilometers altitude, the following weights are desirable upper limits: Batteries and other power supply, 150 grams; transmitter (including thermal insulation and antenna), 100 grams; meteorograph and clockwork, 150 grams; balloon, 150 grams; total weight, 550 grams. For radiometeorographs operating up to thirty kilometers: Batteries, etc., 250 grams; transmitter, etc., 150 grams; meteorograph, etc., 150 grams; balloon, 250 grams; total weight 800 grams. Simplicity is to be gained by standardizing the best and simplest telemeteorograph system available, and by using the simplest radio transmitter circuit that will give adequate performance (output and frequency stability). Low cost can be obtained partly through this simplification, and partly through manufacture of standardized units in fairly large quantities. A desirable upper limit of cost is twenty-five dollars (exceeded at present) which would enable the radiometeorograph to compete economically with the sounding-airplane, even assuming the loss of the balloon.

Possible Methods of p, t, etc., Indication

The possible ways in which meteorological changes can be translated into intelligible radio indications are four in number, as follows:

(*a*) By varying the intensity of the transmitted signal. Unsatisfactory, as the signal intensity at the receiver is also subject to undesired and uncontrolled variation.

(*b*) By varying carrier frequency (bimetallic strip forming or mechanically coupled to one plate of a condenser in the oscillating circuit, etc.). This system is used and advocated by Duckert, who stresses the advantage of continuous temperature indication. At fairly uniform rates of ascent, however, a smooth temperature curve can easily be plotted from points intermittently recorded. Moreover, where the carrier frequency itself indicates the meteorological element, inordinate precautions are necessary to avoid slight undesired frequency changes that would otherwise be negligible.

(*c*) By modulating the carrier frequency and varying the modulation frequency. Believed to be too uncertain, and too cumbersome from the viewpoint of weight and cost.

(*d*) By interrupting, or keying, a signal of reasonably constant amplitude and frequency, which may be either continuous wave or tone-modulated. This system, as exemplified in the Bureau and Moltchanoff devices, is believed by the present writer, in view of the known characteristics of radio apparatus and radio transmission, to be most worthy of future development.

Improvement of Indicating Meteorographs

In the opinion of this writer, the telemeteorograph as invented in 1875 by Olland and adapted to radio sounding-balloons more recently by Moltchanoff, has not been surpassed in simplicity and completeness of working principle by any later development. It is light, relatively simple, and permits the direct recording of practically continuous p, t, and f curves against time or altitude. It is possible that some better telemeteorographic principle will be invented. But as the Olland principle is now amply satisfactory, it would appear more profitable to concentrate immediate research in radiometeorography on other subfields such as radio technic.

Radio Improvements—General

In this field much progress in the direction of lightness, simplicity, and low cost remains to be made. It should come chiefly from radio engineers interested in and conversant with modern meteorology. Conventional radio apparatus, as so far applied to radiometeorography, can be considerably improved. It appears to the present writer that the most favorable short waves (five to twenty meters) have not been used. Moreover, the field of ultra-short waves (one to five meters) and microwaves (<one meter) has not even been touched.

Power Supply Improvements

Small lead-acid batteries of A and B types, as developed by Väisälä, are probably near the ultimate in lightweight design. Dry-cell batteries are more convenient to use, and would probably give satisfactory electrical capacity in relation to weight. Dry cells which have been stored for any length of time, however, are in uncertain condition, whereas lead-acid batteries can be filled and charged before use. Also, the constant-voltage characteristic of the lead-acid A battery is particularly suitable for electronic filament operation. Most batteries are affected by low temperature. Below about −20 degrees centigrade, ordinary dry cells become practically inoperative. Thermal insulation is therefore mandatory. As a substitute for the B battery, the buzzer transformer is worthy of careful consideration, particularly as its produces a relatively broad, tone-modulated signal which is easily received. Its performance, however, depends on the adjustment of its vibrating contact, which may change considerably in response to large temperature changes. Enclosure of the contact or the entire device in a glass vacuum bulb, which might also enclose the entire transmitter, would be worth considering.

Improvements in Ordinary Short-Wave Technique ($\lambda > 5$ *Meters*).

In contrast to the 20- to 150-meter spectrum so far used in radiometeorography, wavelengths below fifteen meters have three advantages: (*a*) Effective transmitting antennas are smaller and lighter; (*b*) oscillator circuit arrangements are smaller, lighter, and often simpler; (*c*) the five- to fifteen-meter region is removed from the high power interference which blankets the long-rage wavelengths above fifteen meters.

Receivers

Receivers for five to fifteen meters are of conventional design, somewhat refined, and can easily be combined with recording devices. Moreover, it is possible to use loop aerials for tracking radio pilot-balloons at wavelengths down to five meters. Whether these shorter waves would suffer greater directional vagaries than longer waves, will have to be determined by experiment.

Transmitters

Before optimum transmitter design in radiometeorography is reached, one important question must be answered: Is crystal frequency stabilization necessary, or desirable? The crystal does insure better frequency stability. But it also entails slightly greater weight and complexity, and considerably greater cost. Using tone-modulated waves, as produced by buzzer-transformer plate supply, self-stabilized circuits will certainly suffice, and with correct circuit design and adequate thermal insulation, they may suffice for continuous waves. The tuned-plate—tuned-grid circuit, or the simpler tuned-plate variation of it, is inherently more stable than the ordinary Hartley circuits so far used in radiometeorography. With suitable thermal insulation, it might give frequency stability adequate for all practical purposes. In addition, simple and cheap methods of frequency stabilization, such as the resistance-stabilized oscillator of Kusunose and Ishikawa, are available.

Antennas

So far most investigators have used a half-wave dipole antenna, the upper end being tied to the balloon, the transmitter being connected in the middle (current feed), and the lower end hanging free. A more stable arrangement, from the viewpoint of swinging, etc., in shifting air currents, might be hanging the transmitter (perhaps equipped with damping vanes of light pasteboard) at the lower end of the half-wave dipole, which in this case would be voltage-fed.

Use of Ultra-Short Waves and Microwaves ($\lambda < 5$ meters)

This entire field, entirely unexplored so far as meteorography is concerned, shows considerable promise in certain directions. According to Beverage, Peterson, and Hansell wave propagation in this region is in general optical with considerable diffraction at the higher wavelengths, becoming strictly optical at about λ one meter. In balloon radiometeorography this optical characteristic is no disadvantage, as there will always be a direct air path between a normally rising balloon and a properly located ground station.

Receivers

Receivers for this wavelength region are highly specialized, being mostly of superregenerative and superheterodyne types. They can easily be adapted to radiometeorographic indication and recording. For radio tracking of pilot-balloons, several possibilities appear. The half-wave dipole, giving a minimum signal when the transmitter is on a line with it, is compact and easily movable at wavelengths below ten meters. Antenna array or "beam" systems, as developed by Yagi and others become reasonably small and wieldy at about λ one to two meters, and have distinct direction finding possibilities. Finally, in the microwave region below λ 0.2 meter, solid reflectors (and even lenses) of optical type are feasible, offering some interesting possibilities in three-dimensional direction finding.

Transmitters

In the ultra-short wave and microwave regions several types of transmitters are available, though probably not all suited to radiometeorography. The magnetron oscillator, which reaches very short wavelengths, is definitely too heavy. The Barkhausen-Kurz oscillator circuit, also capable of producing microwaves, requires relatively high grid and plate voltages; whether the requisite power supply could be made light enough is questionable. Regenerative oscillators, which operate at low plate voltage, are limited to wavelengths above two meters when ordinary electronic tubes are used.

Very small electronic tubes have recently been developed by Thompson and Rose. These tubes enable the regenerative oscillator to reach wavelengths less than one meter and may permit other circuit developments of importance in radiometeorography.

Probable Future Usefulness of Radio Balloons

Classification of Types

Radio balloons will naturally be grouped into two types, pilot-balloons and sounding-balloons, as they are used (*a*) to determine upper-air motion or (*b*) to determine upper-air characteristics. In addition, the writer envisions pilot-balloons and sounding-balloons of two general classes: (1) low altitude (up to six kilometers) and (2) high altitude (up to thirty kilometers). These two classes may be exactly similar in principle, but must differ in certain details such as power of radio equipment, thermal insulation, scale of meteorographic instruments, etc.

Radio Pilot-Ballons

Any radio sounding-balloon can, of course, serve simultaneously as an ordinary pilot-balloon in clear weather; or as a radio pilot-balloon in poor visibility, provided that adequate direction finding receivers are available on the ground. It is possible that, due to wave transmission characteristics, radio pilot-balloons will develop along separate lines, using a frequency spectrum widely different from that suited to radio sounding. It is also possible, however, that the same frequency range may serve for both sounding and tracking; in which case one might envision a combined radiometeorographic instrument of either low altitude or high altitude type, which could be used complete for sounding with or without tracking, or used for tracking only without its detachable meteorographic unit. In any case, it is clear that for upper-wind determination in poor visibility the radio pilot-balloon has no practical competitors, and its use is clearly indicated under these conditions.

Radio Sounding-Balloons

Subdivided into probable low altitude and high altitude classes.

Low Altitude Class (Up to Six Kilometers)

This class competes with airplane sounding, which is already satisfactory, getting the data back with certainty for evaluation one or two hours after they are taken. In weather dangerous to flying, which is often the form of weather most deserving of investigation by soundings, the radio balloon is markedly superior to the airplane, and its use is indicated even at present cost levels. Even in fair weather, the radio balloon gets the data back more quickly than the airplane, and may, with standardization and mass production, prove superior from a cost viewpoint, particularly in settled regions where the percentage of returned balloons is reasonably high.

High Altitude Class (Up to Thirty Kilometers)

This class competes with the ordinary sounding-balloon, and should practically supplant the latter in most investigations. The radio development is incomparably quicker, the data being evaluated as they are taken rather than days or weeks later. Moreover, the radio balloon is far more certain, particularly in sparsely settled regions where the balloon may never be recovered. The balloon radio installation need add but very little to the total high altitude sounding-balloon cost, and the ground radio installation should be more than justified by vastly superior results, particularly where many soundings are made from a single fixed or mobile station.

Need for Standardization

All radiometeorographic possibilities should be explored, as rapidly as possible, in order to determine : (*a*) optimum frequency (wavelength) ranges for various types of radiometeorographs; (*b*) optimum apparatus design for most efficient use of these frequencies. Considering past and present trends in the development of both meteorographic and radio technique, it seems probable that such optimum design, once reached in the present state of the art, will remain reasonably efficient for some years. This will permit the standardization and large-scale production which alone can make the full benefits of radiometeorography available to modern meteorology.

Proceedings of the Institute of Radio Engineers
Volume 23, Number 11 *November, 1935*

ECLIPSE EFFECTS IN THE IONOSPHERE*

By

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Summary—*It is concluded from measurements of virtual heights and critical ionization frequencies of the various regions of the ionosphere which were made during two solar eclipses at Deal, New Jersey, that ultra-violet light is an important ionizing agency in the E, M, F_1, and F_2 regions.*

As a result of pulse measurements made at Deal, New Jersey, during the partial eclipse of the sun February 3, 1935,[1] and during the total eclipse of the sun of August 31, 1932,[2] we now have data which show that the passage of the moon's shadow across the earth is accompanied by a decrease in ionization in four of the ionized regions of the ionosphere (E, M, F_1 and F_2).[3]

During the 1932 eclipse the ionic density in the E and F_1 regions was found to decrease, with the maximum effect occurring shortly after the eclipse maximum. Since the ionization in these two regions ordinarily changes uniformly with time, and since the variations observed during the eclipse were much larger than normal variations, we believe that the decrease in ionic density was actually caused by the eclipse. As regards the changes observed in the F_2 region, our 1932 results were not conclusive because the maximum effect in this region did not coincide with the eclipse but occurred somewhat later. The ionic density in this region is known to fluctuate on at least some non-eclipse days and did in fact undergo comparable variations on several occasions during the two days preceding and the two days following the eclipse. Other observers have reached the same conclusions as regards the F_2 region during the 1932 eclipse from thier own data.[4]

The data from which our conclusions were drawn are shown in Fig. 1.

* Decimal classification: R113.61×R113.55. Original manuscript received by the Institute, April 19, 1935. Presented before joint meeting, I.R.E.—U.R.S.I., Washington, D.C., April 26, 1935.

[1] Letter to *Nature*, vol. 135, p. 393; March 9, (1935).

[2] Mention has already been made of the results of our 1932 eclipse experiments in the following publications:

Science, November 11, (1932); *Proc. Fifth Pacific Science Congress*, vol. 3, pp. 2171–2179, (1934); *Nature*, September 30, (1933); *Bell Lab. Record*, March, (1935).

The data have never been published and we are therefore including some of of it in this paper as it may be of interest to other investigators in this field.

[3] M refers to the intermediate region between E and F_1.

[4] Kirby, Berkner, Gilliland, and Norton, Proc. I.R.E., vol. 22, p. 246–265; February, (1934); Henderson, *Canadian Jour. Res.*, January, (1933).

The double maximum in virtual height with a minimum between for 2398 and 3492.5 kilocycles[5] is interpreted by us to have been caused by a decrease in ionic density in the F_1 region, which resulted in a change in reflection from the F_1 to the F_2 layer during the central part of the eclipse.

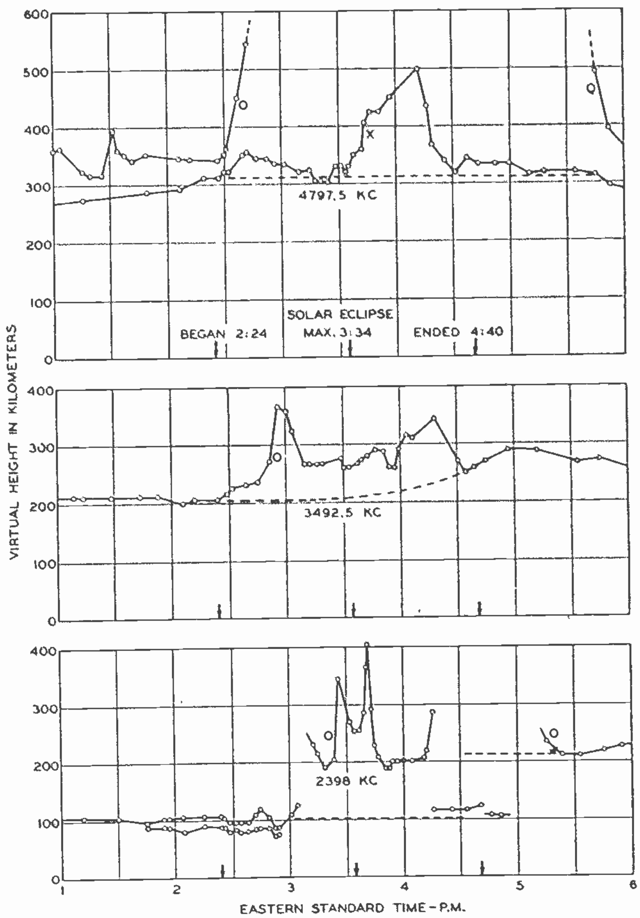

Fig. 1—Virtual height values during the total eclipse of August 31, 1932.

From the curves of Fig. 1, it is possible to plot the virtual height contour map shown in Fig. 2. Since we know as a result of data taken on several hundred days that the X-component curve in Fig. 1 for 4797.5 kilocycles is nearly equivalent to the O-component[6] curve for a frequency approximately 750 kilocycles lower (i.e., 4050 kilocycles), there are in effect O-component curves for four different frequencies

[5] Mimmo and Wang, Proc. I.R.E., vol. 21, pp. 529–545; April, (1933), and Kenrick and Pickard, Proc. I.R.E., vol. 21, pp. 546–566; April, (1933), obtained similar double maximum curves.

[6] The expressions "O-component" and "X-component" are used in place of the terms "ordinary ray" and "extraordinary ray" used by other writers.

available for plotting the contour map. The dotted lines represent regions of maximum ionization. These curves show in a rather striking manner the sharp decrease in ionization of the E and F_1 regions near the time of the eclipse maximum.

The uncertainty of the 1932 results as regards the F_2 region led us to concentrate our efforts on this region during the 1935 eclipse. Improved technique now made it possible to measure the critical ioniza-

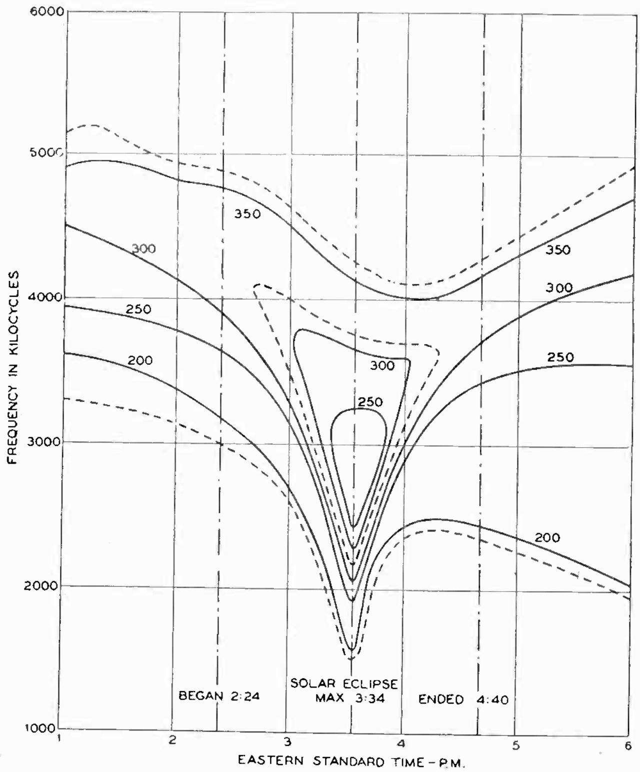

Fig. 2—Virtual height contour map drawn from data for Fig. 1, for August 31, 1932.

tion frequencies directly instead of making virtual height measurements on fixed frequencies as had been done during the 1932 eclipse. The critical frequencies for the E and M regions were measured in addition to those for the F_2 region.

We found that this eclipse was accompanied by a decrease in the maximum ionic density in all three regions and that the minimum ionization occurred at or very shortly after the eclipse maximum. The percentage decrease was progressively greater from the lowest to the highest region, being approximately twenty per cent for the E region, twenty-two per cent for the M region, and twenty-five per cent for the F_2 region.

Some such progressive change might be expected from the fact that the eclipse had a magnitude of forty per cent at the ground and approximately forty-three per cent in the F_2 region (250 kilometers over Deal). These magnitudes are in terms of the sun's diameter, and on the basis of eclipsed area these figures become twenty-nine and thirty-one per cent, respectively.

Fig. 3 gives the critical ionization frequencies for the three days on which data were obtained. The curves for the E and M regions are for

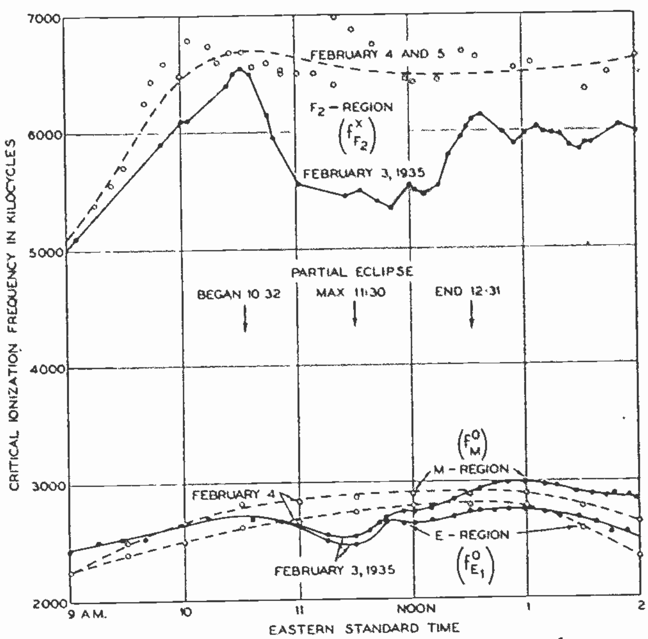

Fig. 3—Critical ionization frequencies during the partial eclipse of February 3, 1935.

the *O*-component while the curves for the F_2 region are for the *X*-component. For the *O*-component, the ionic density, N, is proportional to f_c^2, where f_c is the critical ionization frequency, while for the *X*-component curve in Fig. 3, N, is proportional to $(f_c - 750 \text{ kc})^2$.

The decreases in ionic density of the various regions may be compared with a fifty to sixty per cent decrease in the E region ionization during the eclipse of August 31, 1932, when the eclipse magnitude was ninety-five to one-hundred per cent.

The 1935 measurements give a more definite synchronism than those of 1932, between the eclipse occurrence and the time of decrease in ionic density of the F_2 region.

In view of the variable nature of the F_2 region, it is a possibility that the decrease in ionic density at the time of the eclipse was a mere coincidence and was actually due to some noneclipse agency. We believe that this was not the case and that the decrease in F_2 ionization was a bona fide eclipse effect, as the decrease began within a few minutes after the first contact, the density attained its lowest value shortly after the maximum of the eclipse and recovered to a more or less constant higher value a few minutes after the last contact. At no time during these measurements on the eclipse day or the days after was there any other variation of a comparable magnitude.

These results, therefore, indicate that ultraviolet light[7] is an important ionizing agency in the E, M, F_1 and F_2 regions of the ionosphere.

[7] While ultraviolet light is probably the ionizing agency responsible for the effects noted, any other solar emanation which travels substantially at the velocity of light, should not be precluded from consideration. *See* E. A. W. Müller, *Nature*, February 2, 1935, who suggests Roentgen type radiation.

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ON THE CORRELATION OF RADIO TRANSMISSION WITH SOLAR PHENOMENA*

By

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Summary—*A daily character figure for radio transmission is obtained from the data of the short-wave transatlantic telephone circuits of the American Telephone and Telegraph Company which shows a positive correlation with character figures of terrestrial magnetism and earth currents. Charts of these new radio character figures show the twenty-seven-day recurrence tendency and are similar in general appearance to such charts of terrestrial magnetism and earth currents.*

A method of prediction, based on the twenty-seven-day recurrence tendency is discussed. Such data as are available at present indicate that long-range predictions based on the eleven-year solar cycle will prove fairly reliable.

Attempts to link particular solar disturbances with the individual radio disturbances of several sequences of twenty-seven-day intervals were not very successful. It is indicated that the most promising line of investigation for further study is a more continuous watch for sudden activity of solar disturbances.

Short-Wave Radio Character Figure

It appears that, in order to secure a character figure for short-wave radio transmission, which is representative for the day, data must be taken over the whole twenty-four hours. Since the New York-London short-wave telephone circuits are in practically continual use, suitable data may be obtained from them.

It has been found that the field strengths of the received signals taken at regular intervals do not furnish the best indication of the over-all character of the day. If, however, the ratio of uncommercial time to total time, or the percentage of time lost to traffic is taken, a fairly good daily character figure is obtained.

In order to facilitate the plotting of these figures in chart form they were reduced to four classifications or group indexes. The following table gives the relation of these character figures (T) and the group indexes (T_1).

T_1 Classification	T* Per Cent Time Uncommercial
0	0–24
1	25–49
2	50–74
3	75–100

* Since the letter R is used for relative sunspot numbers, T is here used for the radio character figure designation.

* Decimal classification: R113.5. Original manuscript received by the Institute April 25, 1935; revised manuscript received by the Institute, August 22, 1935.

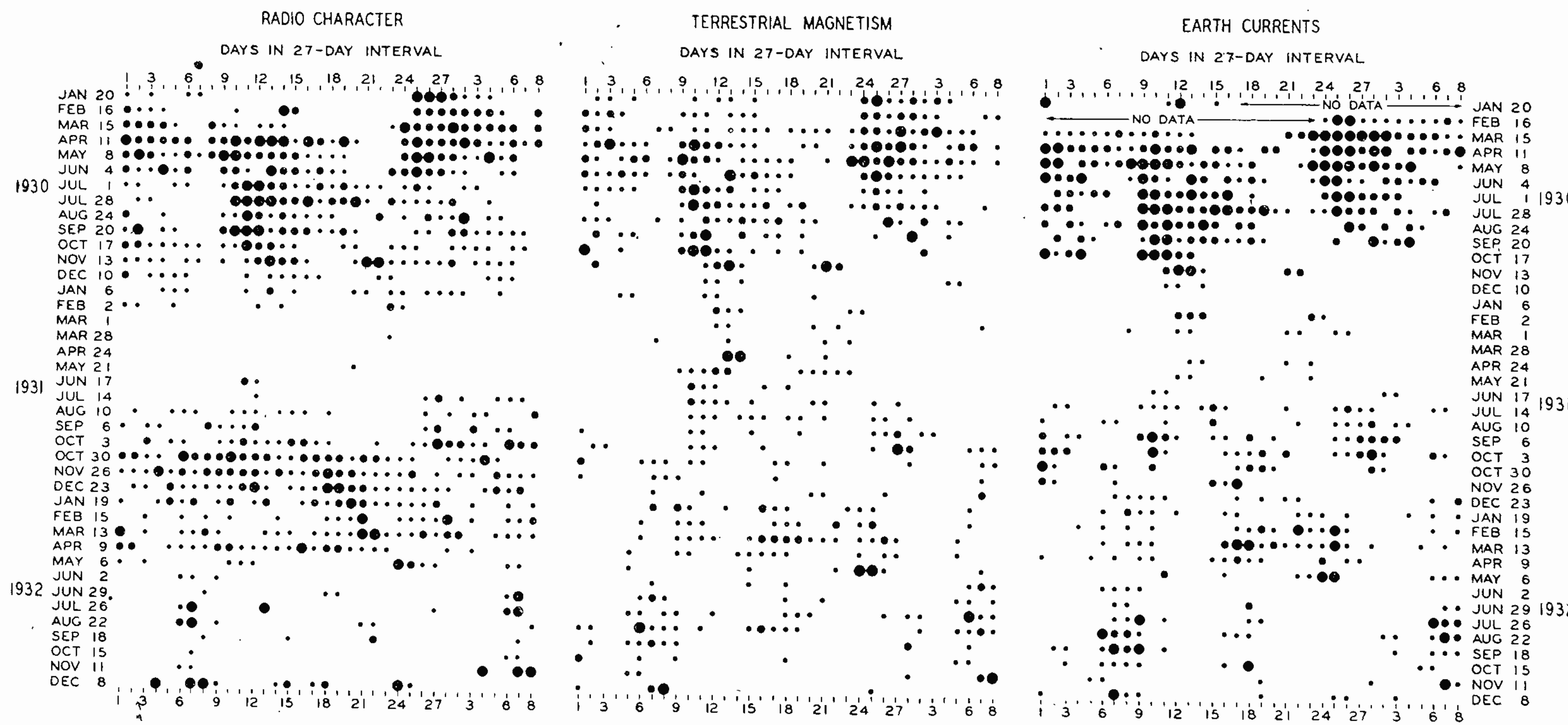

Fig. 1—Day-by-day record of short-wave radio transmission over the North Atlantic Ocean, terrestrial magnetism activity, and earth current activity for 1930, 1931, and 1932, demonstrating the twenty-seven-day recurrence tendency. The dot size corresponds to the severity of disturbance. The magnetic indexes were derived from the daily ranges of horizontal intensity as recorded by the U. S. Coast and Geodetic Survey at the Cheltenham Observatory (Maryland); and those for earth currents were obtained in New York State over lines of the American Telephone and Telegraph Company for 1930 and 1931 and at Bell Telephone Laboratories laboratory at Deal, New Jersey, over a short line for 1932.

Fig. 1 shows these group indexes arranged to bring out the twenty-seven-day recurrence tendency. This is demonstrated by the apparent bunching of the spots into more or less vertical columns. Terrestrial magnetic and earth current data[1] are shown alongside in similar form for comparison.

The twenty-seven-day recurrence tendency is somewhat better demonstrated on the radio chart of Fig. 2. This is because the radio character figures (T) have been arbitrarily altered in a manner which places greater emphasis on minor disturbances when they occur during a period for which the average disturbance is of minor character. This was done as follows: Starting with the daily uncommercial percentage figures an average for each month was taken. The figure T_2 for each day of that month was then determined on the strength of its deviation from this average. Thus for days on which the percentage of time lost was less than the average $T_2=0$. If it fell between the average and $1+\frac{1}{2}$ times the average $T_2=1$, between $1+\frac{1}{2}$ and twice the average $T_2=2$ and above twice $T_2=3$. This resulting group index T_2 (used in Fig. 2) corresponds more closely than does T_1 to the magnetic character figures from which the international character figure C is derived. (C is the world average of the character figures from the individual magnetic observations.)

Predictions

In all of the charts of Figs. 1 and 2, the twenty-seven-day recurrence tendency is well enough marked so that useful predictions of future behavior may be made. For instance, if such a chart is kept up to date, predictions may be made by inspection for any day not more than twenty-seven days distant. Some idea of the probable accuracy may also be obtained from the chart by noting whether the day in question falls, for instance, in the middle of a major sequence or on the ragged edge of a poorly defined one. Such probable accuracy is expressed by modification of the prediction with the words "probably" and "possibly."

Predictions based on the charts of Fig. 1 will give the better forecast of the absolute magnitude of the activity while those based on the charts of Fig. 2 may be expected to give a more accurate idea of the relative magnitude of one day with respect to its neighbors. These two properties could, of course, be obtained from one chart of the type of Fig. 1 if the number of group indexes were so greatly increased as to show all minor variations.

[1] The writer is indebted to Dr. G. C. Southworth for the earth current data of 1930 and 1931.

The correlation between the different phenomena is good enough so that predictions of activity of one nature may be made from the chart of another type of activity. For instance it would be possible to predict the radio behavior from the magnetic chart alone.[2] This method has been found to be of the same order of accuracy as that using the radio chart alone. The reason for this is not, obviously, that the

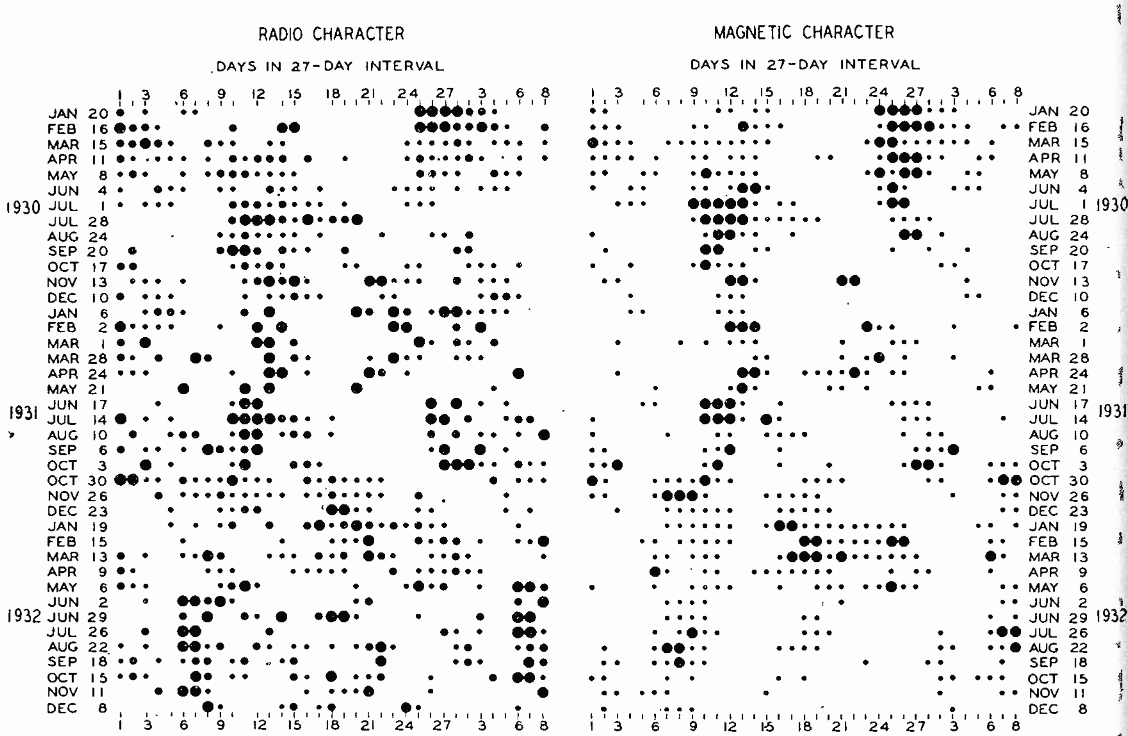

Fig. 2—Relative day-to-day record of short-wave radio transmission over the North Atlantic Ocean, and terrestrial magnetism activity for 1930, 1931, and 1932, demonstrating the twenty-seven-day recurrence tendency. The dot size corresponds to the magnetic character figure as determined at Tucson, Arizona, and to the relative severity of disturbance of the radio transmission.

correlation for the radio and magnetic data is perfect but probably that the radio data as they are obtained at present are subject to greater inaccuracy than the magnetic and such inaccuracy apparently just about compensates for the imperfect correlation between the two phenomena.

In order to determine which type of data may be expected to prove the more useful in making predictions of radio behavior the correlation coefficients between magnetic and radio and earth current and radio data were determined. These are given below.

[2] Such predictions have been made by R. M. Morris and W. A. R. Brown, Proc. I. R. E., vol. 21, 1, pp. 63–80; January, (1933).

Year	*Magnetic-Radio	*Earth Current-Radio
1930 (first half)	+0.66 ± 0.03	+0.76 ± 0.02
1930 (whole year)	+0.63 ± 0.02	+0.67 ± 0.02
1931	+0.30 ± 0.03	+0.44 ± 0.03
1932	+0.44 ± 0.03	+0.46 ± 0.03
Average	+0.46 ± 0.03	+3.52 ± 0.03

* These correlation coefficients are for magnetic activity with severity of radio disturbance and of earth current activity with severity of radio disturbance. They are thus positive.

These coefficients indicate that predictions based on earth current measurements should prove the more accurate. Whether this is true can only be determined, however, by a more comprehensive and detailed study of the data. They also indicate that the correlation is best when the radio disturbances are greatest, which is, fortunately, also the time when predictions are needed the most.

Daily predictions of the behavior of the radio circuits from either the radio or magnetic chart have been correct sixty-two per cent of the time. Similar predictions of the magnetic data from the magnetic chart have been correct seventy-one per cent of the time. These figures have been determined solely on the basis of disturbed or undisturbed days and modification of the forecasts by the words "probably" and "possibly" have not been taken into account.

This method of making predictions, even in its present state, is of definite use commercially. Special forecasts of the same nature have proven useful in planning certain experimental studies. Among the latter were those made for the magnetic and radio behavior during the solar eclipse[3] of August, 1932, and the Leonid meteor shower[4] of November 16, 1932.

Terrestrial-Solar Correlation

The solar data for correlation purposes were obtained mainly from the spectrohelioscopic observations made at Deal, New Jersey. They were supplemented by data published by the observatories mentioned below.

The solar phenomena studied included: (1) sunspots, (2) prominences, and (3) bright and dark hydrogen flocculi. Sunspots are the dark spots which appear on the white solar surface as viewed by an ordinary telescope. The other phenomena could only be seen with the spectrohelioscope. Prominences are the red flaming clouds of gas which project out from the edge or "limb" of the sun and flocculi are bright and dark clouds which appear on the surface of the sun when viewed by the light of the hydrogen line H_α. A flocculus often becomes a prominence when the solar rotation carries it to the limb.

[3] A. M. Skellett, *Science*, vol. 76, p. 169; August 19, (1932).
[4] A. M. Skellett, *Science*, vol. 76, p. 434; November 11, (1932).

Well-defined sequences of activity of approximately twenty-seven-day periods are apparent in both the solar (see below) and terrestrial (Figs. 1 and 2) data. An attempt to link these two phenomena together in a cause-and-effect relationship was not, however, very successful. For several well-marked radio (and magnetic) sequences it was found that no single type of solar activity could be identified in such a manner as to exhibit a clear-cut relationship. For some of the sequences there was some form of solar activity near the center of the sun at the time of each radio disturbance but such activity varied between recurrences in heliographic latitude and longitude and in kind (bright and dark hydrogen flocculi, activity of these as shown by radial velocities and rapid motions, and sunspots).

A similar indefinite result was found by starting with the solar observations. For instance an area on the sun approximately in heliographic latitude of +10 degrees and longitude of 315 to 329 degrees exhibited the presence of either hydrogen flocculi, prominences, or sunspots or combinations of these on each transit across the face of the sun from October 21, 1932, to February 9, 1933, a total of five transits. Sunspots appeared in this region on the last four transits and their identity over this period of time was noted at Mt. Wilson Observatory.[5] From a spectroheliogram taken at the Solar Physics Observatory at Cambridge, England, on December 12, 1932, C. P. Butler[6] determined that this disturbance extended over a great area some thirty degrees in diameter. Although the times of central meridian crossing of this area fall within a well-defined sequence on the radio chart (between days 6 and 9 on the left at the bottom of Fig. 2) the absence of activity on this solar area for earlier recurrences of the radio sequence tends to vitiate the relationship between radio disturbances and those types of solar activity which were observed. Nevertheless, the reality of the twenty-seven-day period in radio is strong indication that solar activity is responsible, even though not convincingly identified in detail.

If the relationship is not simple and clear cut, the statistical method of correlation must be resorted to. A few years' data are not adequate for such a study but from them it was attempted to obtain at least an indication of the results to be expected from a larger amount of data.

For the period from June 4, 1931 to January 6, 1933 there were sixty days for which the radio character figure T_2 was three (severely disturbed). For twenty of these days cloudy weather prevented ob-

[5] *Publications of the Astronomical Society of the Pacific*, vol. 45, p. 53; February, (1933).

[6] *Monthly Notices Royal Astronomical Society*, vol. 93, p. 174; January, (1933).

servation. For each of the remaining forty days[7] there was a disturbed region on the sun and for all but three this active region was within thirty degrees of the center of the sun's disk. For each of the three exceptions this disturbed region, though not near the center, was a large one. However, cases are recorded of such active solar areas passing the central meridian of the sun without a terrestrial disturbance of appreciable magnitude taking place within five days. As far as such statistical data go, we might assume that the presence of an area whose activity may be seen with a spectrohelioscope is a necessary though not a sufficient condition for a radio disturbance of character figure $T_2=3$. On the other hand the foregoing discussion on the basis of the twenty-seven-day period suggests that detection of a disturbed area has not even been a necessary condition though it might be if the observations were made continuously.

A study of the solar distribution of flocculi and spots on terrestrially disturbed and quiet days shows a maximum and minimum, respectively, about thirteen degrees west of the center (one day past); see Fig. 3.[8] These curves are interpreted as indicating that the most probable position of flocculi and spots on disturbed days is thirteen degrees west of the center of the solar disk and that on quiet days it is other than in this region. The one-day interval from the center is interpreted as the time taken for the propagation of the disturbance from the sun to the earth. Probably it should be noted here that the positions taken from the spectrohelioscopic data were made with the aid of a measuring device used visually at the instrument and are not likely to have as great an accuracy as similar measurements made on spectroheliograms. These measurements are accurate enough, however, to justify confidence in the reality of the shape of the curves referred to above.

A number of groups of days was selected with respect to solar disturbance or lack thereof. The percentages of selected days falling into the three classes determined by radio character—severely disturbed, moderately disturbed, and quiet—were compared with the average distribution of all days falling into these classes. A positive correlation was found for the severely disturbed days ($T_2=3$) for both the solarly active selection and the solarly quiet selection of days (actually a negative correlation here) but not generally for the moderately disturbed ($T_2=2$ or 1) and quiet ($T_2=0$) days. The same study for the magnetic data gave a positive correlation throughout for the solarly active days.

[7] On six of these days there were no spots visible on the sun.

[8] The curves of Fig. 3 are corrected for foreshortening. The uncorrected curves show the same general characteristics as to the dip referred to above.

Similar studies were made of the radio data for the presence of flocculi exhibiting radial velocities. These are generally short-lived phenomena and most such occurrences must necessarily be missed since the sun is actually under observation for only a small per cent of the time. The data in this case were very meager but gave a positive correlation.

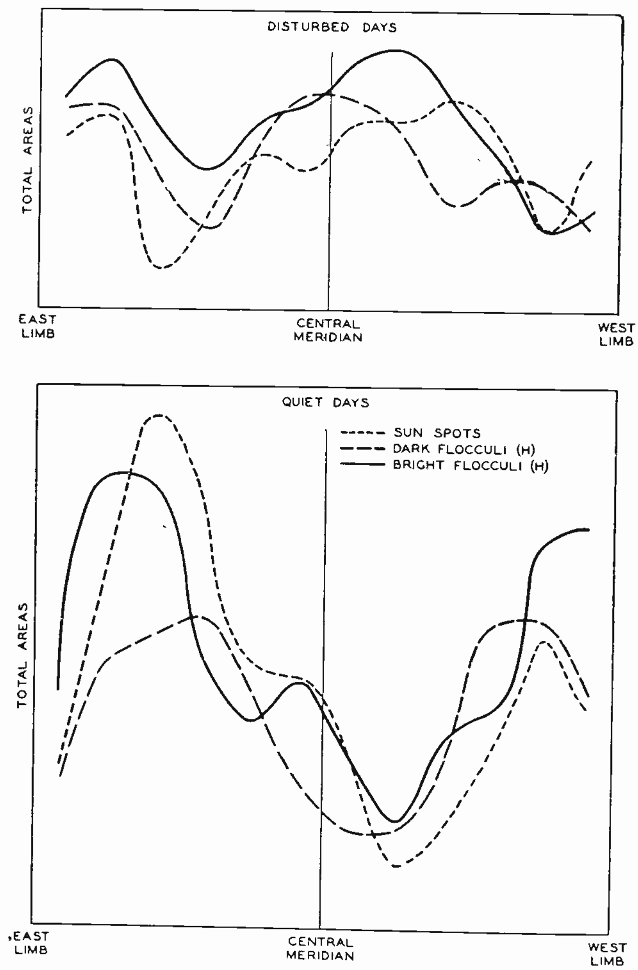

Fig. 3—Distribution across the solar disk of solar phenomena on terrestrially disturbed and quiet days.

This study was continued for days on which flocculi with radial velocities were observed near the center of the sun. These gave the best correlation of all, though here the number of days was so few (actually only thirteen) as to place little faith in the result. It is mentioned here, however, because it supports the idea that could the solar disk be watched continuously on a world-wide program of observation, as Hale[9] has suggested, to record all solar outbursts and so to increase

[9] *Astrophys. Jour.* vol. 73, p. 408, (1931).

the completeness of the solar data until it approaches that of the terrestrial data, it seems likely that considerably more success might be obtained in determining the solar-terrestrial relationships.

It is of interest to record the data for the most severe radio disturbance and magnetic storm (May 29, 1932) experienced during this period. They occurred during the passage across the face of the sun of one of the most active solar regions observed. The largest radial velocity that has been measured since observations were begun was exhibited by a mass of gas in this region. The magnetic storm was accompanied by an aurora visible as far south as Mexico. The gases over this solar region exhibited extreme turbulence on the face of the sun and when it had reached the western limit an enormous prominence of great brilliancy and activity was seen to blow off into space. It was over 100,000 miles in length and completely disappeared within an hour.

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HIGH POWER OUTPHASING MODULATION*

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As long as the power of broadcast stations was not over a maximum of some tens of kilowatts, the operating expenses as far as power supply and tubes are concerned remained small in comparison with the total expenses (first cost of plant, artists, copyright charges, staff, etc.). However the situation rapidly changed when the power on carrier-wave operation came to exceed 100 kilowatts,[1] and when the same program was to be relayed simultaneously by several stations.

In fact, it is possible to state that at the present time, the power supply and the high power tube expenses already represent the greatest part of the operating cost of the larger stations and that this situation will certainly continue to develop along these lines.

It is therefore important to design stations in which, for a given power output, the power required for tube operation will be a minimum and one in which the over-all efficiency will be as high as possible.

The modulation system hereinafter described and called "outphasing modulation" (a system in operation in several high power stations, with as much as 150 kilowatts in the antenna) entirely meets this double condition. From the operating point of view, and without impairing the quality which above all a broadcast station must have, it confers economic advantages that no other system at the present time can pretend to possess.

Standard Modulation Systems and Principles of the New Method

All the present modulation systems can be classified in two categories: either the anode voltage of the high-frequency power stage is controlled by means of a low-frequency amplifier of suitable power in such a manner as to vary the output of this stage, or, the anode voltage remaining constant, the output power is controlled by an action on the grids of the high-frequency power amplifier.

* Decimal classification: R410. Original manuscript received by the Institute, January 8, 1935; translation received by the Institute, March 8, 1935.

[1] In Russia and in the United States notably, there are in operation stations using 500 kilowatts in the carrier wave which the present state of the art permits to be easily attained.

In the first case the efficiency of the power stage remains high and its tubes always work under excellent conditions but the low-frequency amplifier must be able to supply a considerable power (half the power applied to the power stage) thus requiring an important supplementary tube installation and a more or less considerable supplementary power expense.[2]

In the second case, the complication of the low-frequency high-power amplifier is avoided and the power consumed in tube operation is kept down to the instantaneous power required, but the efficiency of the power stage on carrier wave only reaches half the maximum efficiency.

Consequently, for various reasons, the systems derived from these two categories cannot solve the stated problem.

However there is another possibility as to modulation. This is to control the load (without any supplementary power expense) within wide limits, under constant grid excitation and constant anode voltage; that is to say, to control the *load impedance* of the output circuit during the modulation cycle.

In fact, when a tube with a high negative bias is strongly excited (class C operation) the *total apparent power* is obtained under good conditions of efficiency (for instance 75 per cent) and the real efficiency in *real power* is the preceding efficiency figure multiplied by the power factor of the output circuit. Therefore, if the latter always remains around one when the *load impedance* varies, it will be possible to keep a high efficiency for the system during the modulation cycle while at the same time the tube will supply a useful power which will vary from nil to the maximum power that the tube is able to supply.

This principle of operation could be realized in practice by coupling, for example, the antenna circuit to the output circuit of the amplifier and then controlling this coupling by means of low-frequency currents (saturation inductance or magnetic modulator, for instance).

However along these lines important difficulties would be encountered due to the presence of an iron core and its utilization at high frequencies, and this control would probably also require considerable power necessitating supplementary tubes. Another path leading to the same results consists in dividing the amplifier, according to the principle diagram of Fig. 1, into two parts each including one or several tubes according to the power required and each part having its own

[2] According to the operating conditions of the tubes of the low-frequency power amplifier (operation in the straight parts of the characteristics, called class A, or in the bent parts with polarization at the origin of the anode current, called class B) the importance of this installation varies but nevertheless it always remains rather considerable as will be shown further on.

output circuit while the load circuit is differentially coupled to both.

The variable load is then obtained by acting on the phase difference between the grid excitations of the two parts of the final amplifier, whence the name of "outphasing" modulation given to the system.

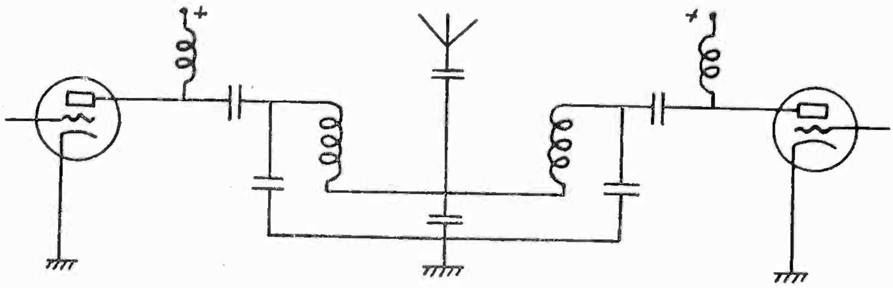


Fig. 1

We shall first study the reaction caused by the load circuit when the relative phase of the two generated voltages varies.

Active and Reactive Components of Load

Let us consider the diagram of Fig. 2.

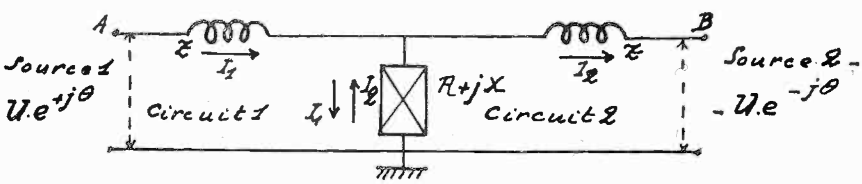


Fig. 2

$(Ue^{+j\theta})$ and $(-Ue^{-j\theta})$ represent two high-frequency voltages of equal amplitude (those generated between the terminals of the output circuits of the last stage tubes) but of relative phases varying with θ; Z represents the equal reactances of the output circuits inductances (the capacities of these circuits are not yet shown); $R+jX$ is the impedance of the load circuit (antenna) called, from its location, the middle branch (R resistance, X reactance); j the symbol of the imaginary quantity $\sqrt{-1}$.

Then two equations can be written

$$jZI_1 + (R + jX)\cdot(I_1 - I_2) = U\cdot e^{+j\theta}, \tag{1}$$

and,

$$-jZI_2 + (R + jX)\cdot(I_1 - I_2) = -U\cdot e^{-j\theta}, \tag{2}$$

or by addition and subtraction,

$$I_1 - I_2 = 2jU\frac{\sin\theta}{2R + j(Z + 2X)}, \tag{3}$$

and,

$$I_1 + I_2 = 2U \frac{\cos\theta}{jZ}. \tag{4}$$

It will be noticed from (3) that the current in the load circuit starts from zero for $\theta=0$ then increases in proportion to the sine of angle θ.

As long as θ is rather small, the corresponding law is linear. If θ reaches somewhat high values, it will still be possible to have a linear law if U has the form $U_0\cdot\theta/\sin\theta$; or in other words, if the amplitude also varies very slightly during the modulation cycle, it will still be possible to have a linear law in respect to modulation if θ follows a suitable law in relation to the latter. This question will be discussed later.

Equations (3) and (4) give for I_1 and I_2 the values

$$I_1 = \frac{U}{jZ}\left[\cos\theta + j\sin\theta \frac{jZ}{2R + j(Z+2X)}\right],$$

$$I_2 = \frac{U}{jZ}\left[\cos\theta - j\sin\theta \frac{jZ}{2R + j(Z+2X)}\right].$$

By making $2R/Z=\alpha$, and $2X/Z=\beta$, those expressions can be easily transformed into

$$I_1 = \frac{U}{jZ}\left[\left(\cos\theta - \frac{\alpha\cdot\sin\theta}{(1+\beta)^2+\alpha^2}\right) + j\sin\theta\frac{1+\beta}{(1+\beta)^2+\alpha^2}\right], \tag{5}$$

$$I_2 = \frac{U}{jZ}\left[\left(\cos\theta + \frac{\alpha\cdot\sin\theta}{(1+\beta)^2+\alpha^2}\right) - j\sin\theta\frac{1+\beta}{(1+\beta)^2+\alpha^2}\right]. \tag{6}$$

The active and reactive power displayed will be obtained for each of the two circuits by multiplying the currents I_1 and I_2 by the corresponding voltages, changing however, according to a well-known law in the calculus of imaginaries, the sign of the term in j in the expression for I or U.

Designating by $W_{1\,\text{real}}$ and $W_{2\,\text{real}}$ the real power in the two circuits, we obtain for $W_{1\,\text{real}}$ and $W_{2\,\text{real}}$ the following values:

$W_{1\,\text{real}}$ = real part of

$$\frac{U^2}{jZ}\cdot e^{-j\theta}\left[\left(\cos\theta - \frac{\alpha\cdot\sin\theta}{(1+\beta)^2+\alpha^2}\right) + j\sin\theta\frac{1+\beta}{(1+\beta)^2+\alpha^2}\right], \tag{7}$$

$W_{2\,\text{real}}$ = real part of

$$\frac{U^2}{jZ} \cdot e^{+j\theta}\left[\left(\cos\theta + \frac{\alpha \cdot \sin\theta}{(1+\beta)^2+\alpha^2}\right) - j\sin\theta\,\frac{1+\beta}{(1+\beta)^2+\alpha^2}\right]. \tag{8}$$

Or, using the relations $e^{\pm j\theta} = \cos\theta \pm j\sin\theta$,

$$\left.\begin{matrix} W_{1\,\mathrm{real}} \\ W_{2\,\mathrm{real}} \end{matrix}\right\} = \frac{U^2}{Z}\left[\mp \sin\theta\cdot\cos\theta + \frac{\alpha\cdot\sin^2\theta}{(1+\beta)^2+\alpha^2} \pm \sin\theta\cdot\cos\theta\,\frac{1+\beta}{(1+\beta)^2+\alpha^2}\right]. \tag{9}$$

The real powers in each circuit can only be equal to each other when the first and third terms balance each other which means making

$$(1+\beta)^2 + \alpha^2 = 1 + \beta.$$

If β^2 is negligible in comparison with unity, that is in fact if α^4 can be neglected in comparison with unity, this may be written

$$\beta = -\alpha^2. \tag{10}$$

This condition is always practically fulfilled. Due to the minus sign, (10) implies that reactance X must be a capacity.

Then transforming Fig. 2 into the equivalent Fig. 3,

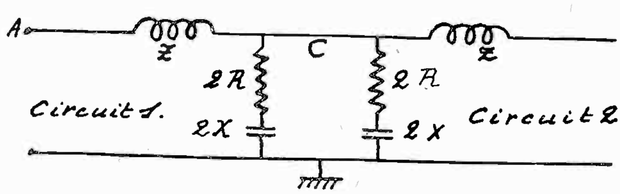


Fig. 3

we see that condition

$$\beta = -\alpha \quad \text{or} \quad \beta/\alpha = -\alpha$$

that is to say,

$$\frac{2X}{2R} = -\frac{2R}{Z},$$

implies that the middle branch containing the useful load must be mistuned in such a way that the current flows ahead of the voltage by an angle α complementary to that which the current would have in relation to the voltage if the two circuits were split at point C.

It is evident that if resistance R is small in comparison to reactance Z the necessary mistuning of the middle branch tends towards zero.

By adjusting the middle branch to these conditions, (9) takes the value

$$W_{1\,\text{real}} = W_{2\,\text{real}} = \frac{U^2}{Z} \cdot \frac{\alpha}{1 - \alpha^2} \cdot \sin^2 \theta = \gamma \cdot \frac{U^2}{Z} \cdot \sin^2 \theta \tag{11}$$

where,

$$\gamma = \frac{\alpha}{1 - \alpha^2}.$$

The imaginary or reactive powers in the two circuits can also be found by working out the imaginary parts of (7) and (8).

The result of this calculation can then be written:

$$j \frac{U^2}{Z} [1 \mp \gamma \cdot \sin \theta \cdot \cos \theta]. \tag{11a}$$

Noticing that, without any load circuit (middle branch short-circuited) the reactive power would have been for each circuit,

$$j \cdot \frac{U^2}{Z},$$

it can be seen that the reaction of one circuit on the other introduces imaginary loads: $W_{1\,\text{imag}}$, $W_{2\,\text{imag}}$ given by

$$\left.\begin{matrix} W_{1\text{imag}} \\ W_{2\text{imag}} \end{matrix}\right\} = \mp j \cdot \frac{U^2}{Z} \cdot \frac{\gamma}{2} \cdot \sin 2\theta. \tag{12}$$

In other words, the load circuit brings back *capacitive energy* on the side where θ is *positive,* that is to say where the voltage leads when the load increases, and *inductive energy* on the side where θ is *negative,* that is to say where the voltage lags when the load increases and this for angles varying from 0 to 45 degrees.

Equation (12) shows that the reactive energies due to the load circuit are also equal to each other.

The middle branch being adjusted as has been specified in (10) it is possible to calculate currents I_1 and I_2.

$$\left.\begin{matrix} I_1 \\ I_2 \end{matrix}\right\} = \frac{U}{jZ} [\cos \theta \mp \gamma \cdot \sin \theta \pm j \cdot \sin \theta] \tag{13}$$

and more specially the ratio $[I_1/I_2]$ of the r-m-s values

$$\frac{I_1}{I_2} \cong \sqrt{\frac{1 - \gamma \sin 2\theta}{1 + \gamma \sin 2\theta}} \tag{14}$$

or,

$$\frac{I_2 - I_1}{1/2(I_1 + I_2)} = \gamma \cdot \sin 2\theta = \frac{\Delta I}{I_{\text{mean}}}. \tag{15}$$

Equations (13), (14), and (15) show that when θ increases, current I_1 decreases and current I_2 increases, and permit the difference expressed in per cent of the mean current to be easily found. Last, it will be noticed that although the currents are not equal, the powers from (11) and (12) are equal in the two circuits.

Compensating Reactive Energy

If reverting to Fig. 2 we insert between A and the ground on the one hand and B and the ground on the other hand, equal capacities C designed to be in resonance with inductances Z, we shall exactly balance the reactive power, U^2/jZ, and the supplies will deliver exactly the powers calculated in (11) and (12). The power factor equal to 1 for $\theta=0$ will progressively decrease with the load, which must be avoided. Let us, therefore, operate in another manner, decreasing condenser C of circuit 1 by a small capacity c which is shifted to circuit 2; the respective capacities of the circuits will then become $C-c$ and $C+c$.

If c is not too large it is always possible to express the reactive energy absorbed by condenser c (or $U^2c\omega$) by a relation of the form

$$-\frac{1}{2}\cdot\frac{U^2}{jZ}\cdot\gamma\cdot\sin 2\theta_0$$

θ_0 being a certain angle given by

$$\frac{c}{C}=\frac{1}{2}\cdot\gamma\cdot\sin 2\theta_0. \tag{16}$$

The total reactive energy to be supplied by source 1 will be decreased by that quantity and that to be supplied by source 2 will be increased by the same quantity.

Calling $W_1'{}_{\text{imag}}$ and $W_2'{}_{\text{imag}}$ these new values, we have

$$\begin{aligned}
W_1'{}_{\text{imag}} &= -\frac{U^2}{jZ}\cdot\gamma\,\frac{\sin 2\theta}{2}-\left(-\frac{U^2}{jZ}\cdot\gamma\cdot\frac{\sin 2\theta_0}{2}\right)\\
&= -\frac{U^2}{jZ}\cdot\gamma\cdot\sin(\theta-\theta_0)\cdot\cos(\theta+\theta_0)\\
W_2'{}_{\text{imag}} &= +\frac{U^2}{jZ}\cdot\gamma\,\frac{\sin 2\theta}{2}+\left(-\frac{U^2\omega}{jZ}\cdot\gamma\cdot\frac{\sin 2\theta_0}{2}\right)\\
&= +\frac{U^2}{jZ}\cdot\gamma\cdot\sin(\theta-\theta_0)\cdot\cos(\theta+\theta_0).
\end{aligned} \tag{17}$$

In other words, for a given dephasing corresponding to angle $\theta = \theta_0$ the reactive energy is entirely balanced and consequently the power factor is equal to 1.

For all other angles, the power factor may be calculated and from relations 11 to 16 is found equal to

$$\eta = \frac{W_{\text{real}}}{\sqrt{(W_{\text{real}})^2 + (W'_{\text{imag}})^2}},$$

or,

$$\eta = \frac{\sin^2\theta}{\sqrt{(\sin^2\theta)^2 + \sin^2(\theta - \theta_0)\cdot\cos^2(\theta + \theta_0)}}. \tag{18}$$

The general shape of all the curves plotted with (18) is the following: The power factor starts from zero for $\theta = 0$ and increases very rapidly up to 1 for $\theta = \theta_0$. It then decreases slightly (the decrease being smaller, the greater the value of θ_0 chosen), reaches a minimum, and then comes back to 1 for $\theta = \pi/2 - \theta_0$. On Fig. 4 will be found the values for $\theta_0 = \pi/10 = 18°$. From this curve we see that the power factor remains always high as soon as the power is somewhat high. The efficiency can therefore only become bad for very small instantaneous powers which is of no great importance.

Thus, it is seen that by adjusting *below resonance* the circuit where the voltage leads when the power increases (θ positive), and *above resonance* the circuit where the voltage lags when the power increases (θ negative), the reactive energy is balanced, thus permitting satisfactory operation throughout an extended range of variation of θ.

Example. Let us suppose according to Fig. 4, the balance at $\theta_0 = \pi/10$ and the carrier-wave operation at a value of θ, θ_p. If θ_p is taken to be 25 degrees, then the figure shows the power factor will be 0.9.

The intrinsic efficiency of the tubes for the relatively small power of the carrier being taken equal to 75 per cent and the efficiency of the circuits in relation to the antenna being taken equal to 0.9 the overall efficiency on carrier-wave operation will be $0.9 \times 0.9 \times 0.75 \cong 0.6$. For other values of θ_p the efficiency follows the variations of Fig. 4. The angle in carrier-wave operation being thus known, let us determine the other parameters of interest by taking as a basis the new parameter,

$$s = \frac{K.V.A.}{K.W.},$$

defined as the ratio of the *total apparent power* involved in the installation to the real power utilized in the carrier wave.

From (11) and (11a) this ratio is

$$s = \frac{1}{\gamma \cdot \sin^2 \theta_p},$$

whence,

$$\gamma = \frac{1}{s \cdot \sin^2 \theta_p}. \tag{19}$$

For reasons of economy in the installation and in order not to increase too much the losses in circuits 1 and 2 it is necessary to make s as small as possible. If we remark that the value of s just defined for carrier-wave operation falls down to $s/4$ during the instantaneous peaks of 100 per cent modulation, we may take $s=20$, that is to say a power ratio of only five during the peaks of deep modulation. From (19) we then have $\delta=0.28$, whence $\alpha=0.26$.

Consequently, the circuit of the middle branch will have to be untuned by an angle whose tangent is 0.26, that is to say by about 15 degrees.

Relation (15) gives

$$\frac{\Delta I}{I_{\text{mean}}} = \gamma \cdot \sin 2\theta_p = 0.21.$$

In the same manner from (16)

$$\frac{c}{C} = \frac{\gamma}{2} \cdot \sin 2\theta_0 = 0.084.$$

Hence, roughly, the currents of the principal circuits will differ by ten per cent over or under the average value and the circuits will be untuned, one in default, the other in excess of eight per cent in capacity that is to say four per cent in frequency. These various values have been verified very approximately in practice.

Operation During Modulation

If we revert to Figs. 4 or 5[1], the curve called apparent power can represent as well, on another scale, the power absorbed by the tubes, that is to say, the anode current as the operation takes place at a constant anode voltage.

It is easily seen that the curve of this current is the shape of a U, the anode current not being nil at the origin, that is to say, for a no-load condition (due to the untuning of the circuits). In fact, it reaches a minimum for an angle a little smaller than the angle chosen for the balance, then, it remains almost proportional to the useful load and consequently varies according to a square law with the useful current (antenna current). Starting from this curve it is possible to plot the anode current curve (or that of the input power) during the modulation cycle in relation to the sine modulation of the antenna current.

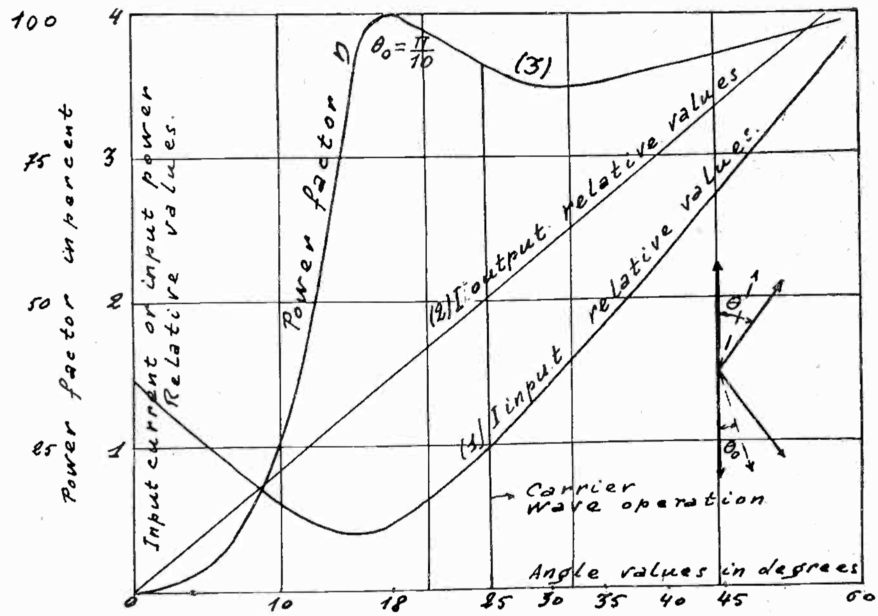


Fig. 4

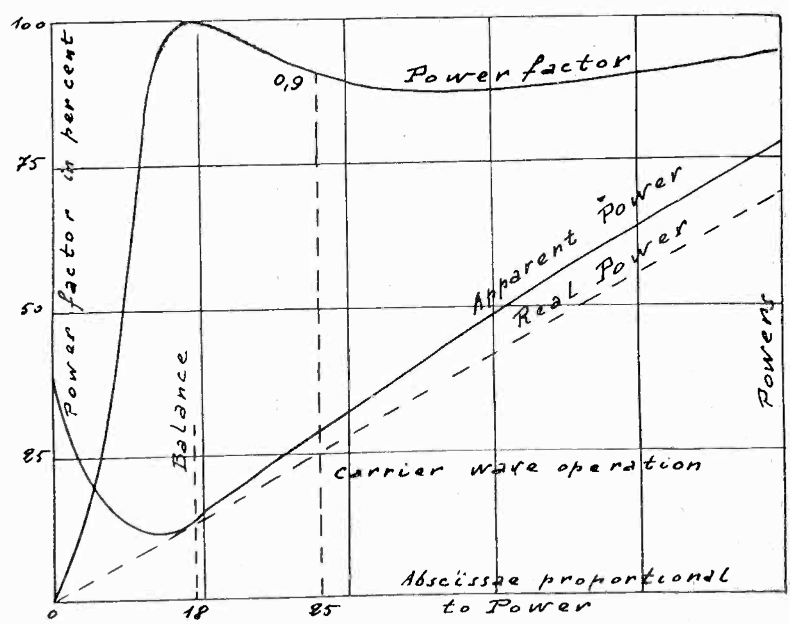


Fig. 5

This can be done for several modulation ratios that is to say for more or less important variations of angle θ around the value corresponding to carrier-wave operation (here $\theta_p = 25$ degrees). Such curves have been plotted for the following modulation ratios 30, 60, 80, and 100 per cent. (Fig. 6.) These curves show the anode current during half the modulation cycle, the other half being symmetrical in relation to the vertical axis. From these curves, it is possible to deduce the mean current, that is to say the mean input power taken from the mains for various modulation ratios. When the modulation ratio K increases this power evidently increases as the effective power in the antenna increases according to law, $1+K^2/2$.

It even ought to increase linearly with this law, if the efficiency really remained constant during the complete modulation cycle. In fact, in the following table the resulting figures can be found when $i_{mean}=1$ on carrier-wave operation.

K	0.3	0.6	0.8	1
i_{mean}	1.045	1.27	1.48	1.76
$1+\frac{K^2}{2}$	1.045	1.18	1.32	1.50

It can be seen that for all average modulation ratios the law of the antenna power is exactly followed and that for deep modulations the efficiency slightly decreases due to the incomplete balance under small loads. For instance, the efficiency for 80 per cent sustained modulation falls down to $1.32/1.48=0.88$ of its value on carrier operation. This has practically no importance as deep modulations only take place during a very short time as compared with the total time, the average modulation of a transmitter scarcely reaching 30 per cent.

Specification and Action of the Driving Stages

Two waves chiefly phase-modulated and of reversed rotations will be obtained by combining in some low power stage an unmodulated carrier wave with an entirely modulated wave (or almost entirely modulated), 90 degrees out of phase with the first one. We mean by "almost entirely modulated wave" a wave having its carrier almost entirely suppressed. Such a wave is represented vectorially by (a) of Fig. 7 where $O'A$ represents the remaining portion of the carrier and $AB=AB'$ the amplitude of the modulated wave.

Diagrams (b) and (c) of Fig. 7 show the results of the above-mentioned combination. Phases vary inversely by a total quantity $\Delta\theta$ of the order of 50 degrees for deep modulations while the amplitude varies only slightly passing from OO' to OB (about 10 per cent for instance). Each of the waves (b) and (c) will then be amplified by separate stages up to the final excitation level (two stages only).

By systematically inserting an outphasing device or simply by acting on the tuning of the circuits of the stages, it will also be possible to give to the grid excitations of the last stage of amplification the desired angles in carrier operation (angle between straight lines OA of Fig. 7 (b) and (c).

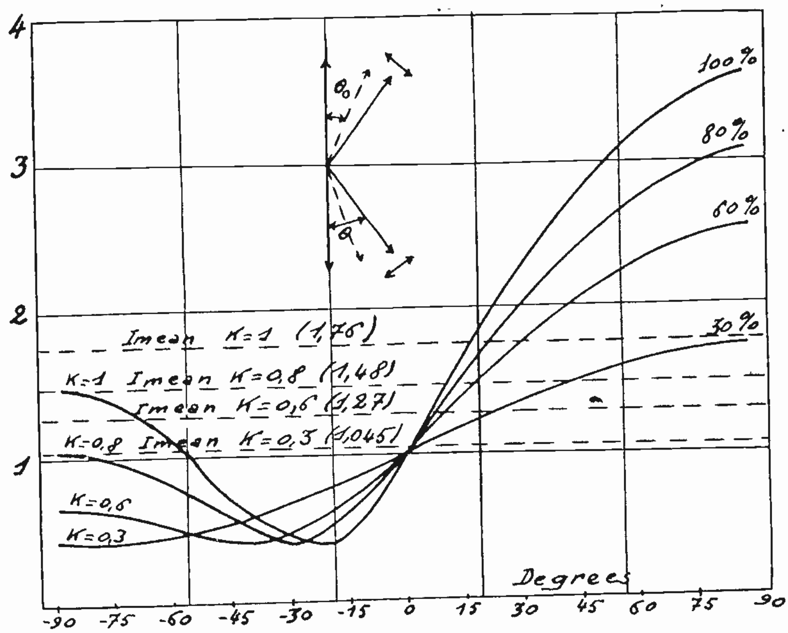


Fig. 6

The vectorial combinations of Fig. 7 can be easily obtained by the bridge connections of Fig. 8, which shows a bridge fed on the one hand at 11′ by an unmodulated voltage and on the other hand at 22′ by a modulated voltage such as that of Fig. 7 (a). It is seen that this bridge

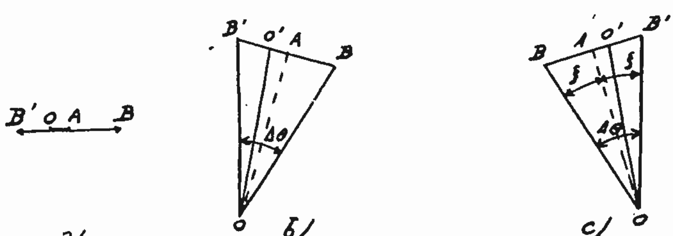


Fig. 7

composed of four equal inductances is tuned simultaneously in respect to both of the excitations by the two condensers shown. Designating by full arrows the currents due to excitation 11′ and by dotted arrows the currents due to excitation 22′ we see that the action of excitation 22′ is the reverse of that of excitation 11′ in the coupled windings 33′.

There will thus appear in these windings an induced voltage such as that indicated in Fig. 7 (b) and (c), the straight lines OO' of Figs. 7 (b) and (c) coinciding, when the currents represented by the arrows will be at right angles.

Excitation $22'$ is itself obtained as the result of a wave amplitude modulated in any manner (by anode control for instance) from which we subtract (in phase with it) an unmodulated wave.

In practice, the complete low power stages of an outphasing modulated station are composed as follows:

(a) A quartz-controlled master oscillator placed in a thermostat, including together with the quartz stage a first buffer stage

(b) A second buffer stage, driving

(c) A modulated stage with its modulator

(d) An unmodulated stage (the whole of (c), and (d) supplying the wave represented in Fig. 7 (a))

(e) The balance bridge driven on the one hand directly by the second buffer (b) and on the other hand by a stage (f) amplifying the wave resulting from (c) and (d).

(g_1) (g_2) Two identical stages amplifying the inversely rotating phase-modulated waves of diagrams Fig. 7 (b) and (c).

All the stages from (b) to (e) included are fitted with fifty watts (nominal) power tubes and the output power of (g_1) and (g_2) is of the order of from 500 to 1000 watts.

Only one intermediary stage is therefore sufficient for driving the last power stage.

Linearity of Modulation Versus Amplitude

Let us call x the vector $O'B$ of Fig. 7 (a) representing the modulated wave's instantaneous amplitude. Then the antenna current ought to vary linearly with x.

Let us assume first that $O'A=0$, in other words that the carrier has been entirely suppressed, for each value of x will correspond an angle ξ of phase variation such that $x=tg\xi$ by taking OO' as unity.

If phase variations in the plate circuits follow phase variations in the grid circuits, and if the small amplitude variations of grid excitation are faithfully reproduced in the plate circuits, angle ξ will represent in fact the variation of angle θ in relation to the value θ_p corresponding to carrier-wave operation.

Noticing that the amplitude of U increases by the ratio $\sqrt{1+x^2}/1$ for extreme angle values $(\theta_p \pm \xi)$, we obtain from (3) a total relative variation of the antenna current equal to

$$\frac{\Delta(I_1 - I_2)}{I_1 - I_2} = \frac{U\sqrt{1 + x^2}\cdot \sin\,(\theta_p + \xi) - U\sqrt{1 + x^2}\cdot \sin\,(\theta_p - \xi)}{U\cdot \sin\,\theta_p},$$

or,

$$\frac{\Delta(I_1 - I_2)}{I_1 - I_2} = \sqrt{1 + x^2}\cdot \sin\,\xi\cdot \text{ctg}\,\theta_p. \tag{20}$$

and as,

$$\sin\,\xi = \frac{x}{\sqrt{1 + x^2}}.$$

$$\frac{\Delta(I_1 - I_2)}{I_1 - I_2} = x\cdot \text{ctg}\,\theta_p. \tag{21}$$

The antenna current variation is therefore linear with x.

In the case where the carrier wave of vectorial diagram 7 (a) is not entirely suppressed, the result is slightly different and we see without calculation that a slight dissymetry is introduced in the modulation curve; this dissymetry is adjustable and can eventually compensate other effects.

Practically the results are less simple due to the amplitude limitation implied by the class C operation of the last amplifying stages, and mainly to the nonlinearity of angular displacement in the grid and plate circuits brought about by the detuning of the last circuit.

Outphasing Between the Grid and Plate Alternating Voltages and its Correction

Circuits (1) and (2) of Fig. 2 being detuned as has been stated, the diagram equivalent to either of the parts of the last amplification stage can be represented by Fig. 9 where, E is the grid voltage.

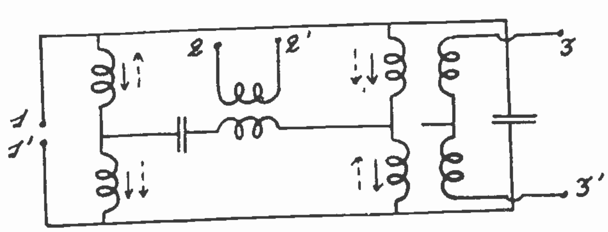


Fig. 8

Then U is the plate voltage feeding the resistance

$$\rho = \frac{Z}{\gamma\cdot \sin^2\,\theta}, \tag{22}$$

and the reactance

$$z = \mp \frac{Z}{\gamma \cdot \sin(\theta - \theta_0) \cdot \cos(\theta + \theta_0)}. \tag{23}$$

These values, (22) and (23), derive immediately from (11) and (12). As θ and θ_0 are to be taken in absolute value, that is to say, independently of their sign, equation (23) must be taken with the minus sign for the voltage, $Ue^{+j\theta}$ and with the plus sign for the tension, $Ue^{-j\theta}$.

If we call r the *apparent resistance* of the tube for the operation considered, we can write

$$r(i_1 + i_2) + \rho i_1 = E, \quad \rho i_1 = jzi_2, \quad \text{and} \quad \left(r + \rho + \frac{r\rho}{jz}\right)i_1 = E.$$

Hence,

$$\frac{E}{\rho I_1} = \frac{E}{U} = 1 + \frac{r}{\rho} + \frac{r}{jz}.$$

Therefore, we shall have a certain angle between excitations E and U, given by

$$\operatorname{tg} \chi = \frac{-j(r/z)}{1 + r/\rho}. \tag{24}$$

r/ρ is nothing else than the ratio of the voltage drop in the tube to the operating voltage. r/ρ is then always small compared to 1, except for peak instantaneous values corresponding to deep modulations.

Angle χ being itself small, in first approximation we get

$$\chi = \pm j \cdot \gamma \cdot \frac{r}{Z} \cdot \sin(\theta - \theta_0) \cdot \cos(\theta + \theta_0). \tag{25}$$

For the voltage $U \cdot e^{+j\theta}$ we must take the plus sign and we then see that for $\theta < \theta_0$, χ is negative, the phase angle $\psi = \theta + \chi$ of the grid voltage must therefore be smaller than θ, that is, $\psi_1 < \theta_1$. For $\theta > \theta_0$, χ becomes positive phase angle: $\psi = \theta + \chi$ of the grid voltage must then be larger than θ so that $\psi_2 > \theta_2$. The range of variation of ψ is then for these two reasons wider than the range of variation of θ.

$$(\psi_2 - \psi_1) > (\theta_2 - \theta_1), \quad \text{or} \quad \Delta\psi > \Delta\theta.$$

Identical reasoning with respect to the voltage $Ue^{-j\theta}$ gives the same results, but with reversed signs. The diagram of Fig. 10 makes these conditions clearer.

If we call α the quantity $\gamma r/Z$, or better $\gamma r/Z \cdot 1/(1+r/\rho)$ which we shall suppose constant by giving to r/ρ its mean value, we have

$$\psi = \chi + \theta = \theta + \alpha \cdot \sin(\theta - \theta_0) \cdot \cos(\theta + \theta_0)$$

whence,

$$\frac{\partial\psi}{\partial\theta} = 1 + \alpha\cdot\cos 2\theta, \qquad \frac{\partial\theta}{\partial\psi} = \frac{1}{1 + \alpha\cdot\cos 2\theta}. \tag{26}$$

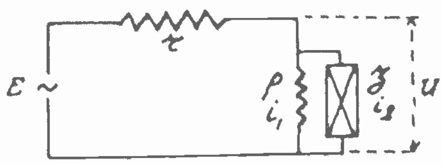


Fig. 9

If we plot the curve of θ versus ψ (Fig. 11) we see that its slope increases when θ increases but always remains smaller than 1 as long as $\theta < \pi/4$, that is, in the useful region. In other words the response to phase variations varies but little as long as θ is small and then increases. It is possible approximately to evaluate α as follows:

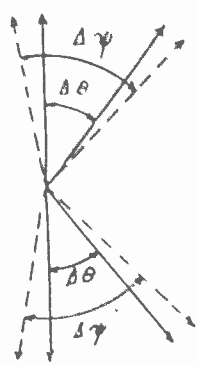


Fig. 10

During peak operation and for 100 per cent modulation $r/\rho = \epsilon$ reaches a value around 20 per cent, the tubes working at the maximum power. On carrier, ϵ is then of the order of 0.05 (one fourth of the preceding power) and this for $\theta = \theta_p$. From (22) we have

$$\alpha = \gamma/Z\cdot\frac{r}{1 + r/\rho} = \gamma\,\frac{\rho}{Z}\cdot\frac{r/\rho}{1 + r/\rho} = \frac{1}{\sin^2\theta}\cdot\frac{\xi}{1+\xi}. \tag{27}$$

We may thus approximately write, taking the point corresponding to carrier operation,

$$\alpha \cong \frac{0.05}{\sin^2\theta_p}. \tag{28}$$

That is, for the chosen example

$$\theta_p = 25 \text{ degrees}, \quad \alpha \cong 0.28.$$

$\partial\theta/\partial\psi$ considered for θ around zero is then almost equal to 0.8 and progressively increases up to 0.85 on carrier and then tends towards one for high instantaneous powers.

It is to be noticed from (28) that this effect is all the more marked as θ_p is chosen smaller.

This effect which tends to increase the antenna current faster than the linear law, is in fact not entirely bad as it efficiently fights against the drop of U for high powers, a drop due to the increase of the voltage drop in the tubes. It can however be too large and we proceed to mention the means practically used for bringing it to the desired value.

The diagrams of Figs. 1 and 2 show but one of the numerous possible ways of obtaining outphasing modulation. In particular it would

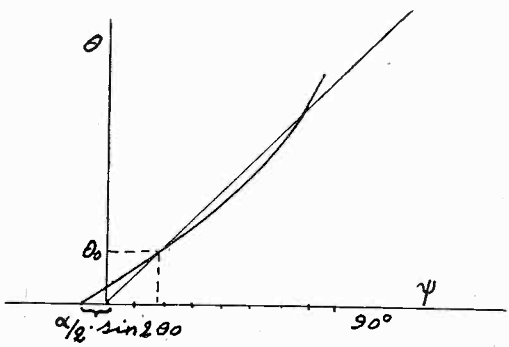


Fig. 11

have been possible to short-circuit the middle branch (Fig. 2) and to place the load between points A and B (between the plates) (Fig. 12). The cancellation of load would then take place for $\theta=\pi/2$. Calculation would then show that for suitable value of R' meeting the condition $R\cdot R'=Z^2$ the load on the output tubes would have been expressed by terms including resistance and reactance such that the resistance term would have the value given by (22) (replacing θ by the complementary angle) and the reactance term the value given in (23) with opposite sign.[3]

It would also have been possible to compensate the reactive energy for angle $\theta=\theta_0$, but the balancing impedance would have a changed sign, χ would still have the value given by (25) but with a reversed sign, so that in the end, the curve of Fig. 11, instead of bending towards the top, would have bent in direction of the bottom.

[3] So that if we combine Figs. 2 and 12 (Fig. 13) with load circuits R and R', the real power supplied is independent of θ, and the reactive power due to the total load is null, in other words, each circuit operates under constant load and without any reactive power variation.

It is thus seen that if we constitute the whole of the last two stages of amplification according to Fig. 14, that is to say by applying Fig. 12 to the stage before last, it will be possible to balance partially or totally the angle χ by an angle χ' obtained by suitably mistuning the stage before last in the *opposite direction* to the last one, and by adjusting resistance r.

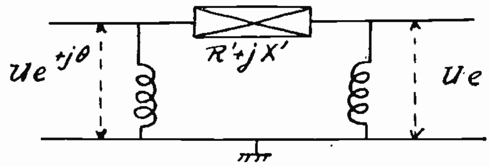


Fig. 12

In particular, if $\chi-\chi'=0$ the voltages U will constantly be in phase with excitations E applied at A and B.

The presence of a resistance r, useful on the other hand to load the excitation circuit, increases the general stability of the system.

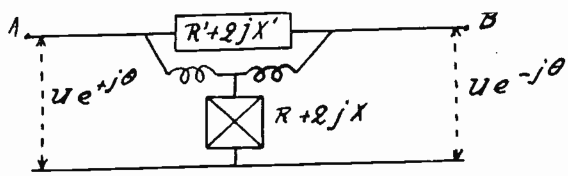


Fig. 13

From what has already been said, the whole of the two stages of Fig. 14 may be directly connected to the output of the low power stages.

It is to be understood that in practice the antenna circuit is not directly connected between C and D but through circuits for filtering harmonics.

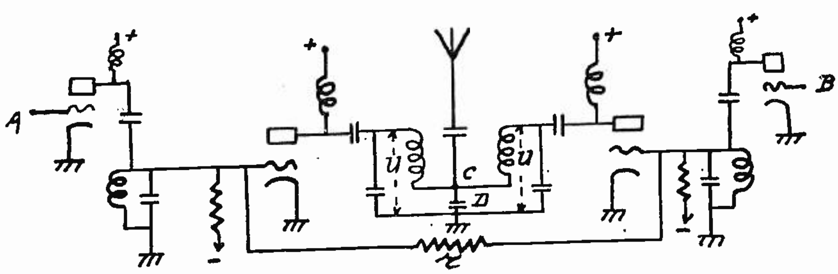


Fig. 14

Regulating the Power

As in all high efficiency systems where the load varies with the modulation ratio, it is to be feared that, due to the voltage drops in the rectifiers or even in the mains, the carrier-wave power weakens during the "forte" of modulation. To avoid this, which does not impair quality as the voltage variations take place only slowly as compared to the period of modulation, the outphasing modulated transmitters are fitted with the following device. Tapped on a potentiometer placed across the direct high voltage, (Fig. 15) at a point where the voltage is small (a few tens of volts) a rectifier is placed in series with a resistance 2 whose cathode is brought to a given positive potential. Resistance 2 is then submitted to a voltage proportional to the difference between the rectified high tension and a certain voltage threshold adjustable by means of the potentiometer tap. This direct voltage is

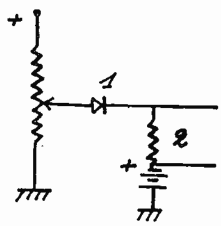


Fig. 15

amplified and it controls (within small limits), the voltage applied to the stage whose aim is to suppress the carrier (the stage called (d), p. 1382) and the output of which is thus varied. We see that for any variation of the main voltage there corresponds a movement of point A of Fig. 7, therefore a variation of angle θ_p and a variation of the carrier-wave power. By proper adjustment, the antenna current drop is cancelled and at the same time we get rid of the power variations due to variations of the mains. The variations of angle θ_p moreover are very small (for instance from 23 to 25 degrees) and instantaneously follow the variations of the direct voltages, consequently no transients are to be feared.

Characteristics and Advantages of the Outphasing Modulation

The linearity of modulation with amplitude, (the question studied at length above), can be made sufficiently perfect so that the prescriptions of the C.C.I.R. (Comité Consultatif International de Radio-Électricite) as regards low-frequency harmonics may be obtained, for instance, 28 decibels at 80 per cent modulation.

For this purpose it is only necessary that the maximum intrinsic power of the tubes permits at the operating voltage the attainment of the possible 100 per cent modulation.

If the power of the tubes is liberally calculated, still better curves may be obtained, for instance 28 decibels at 90 per cent modulation.

The linearity of modulation with frequency can also be obtained without any additional difficulty with this modulation system as long as all the circuits possess wide enough pass bands.

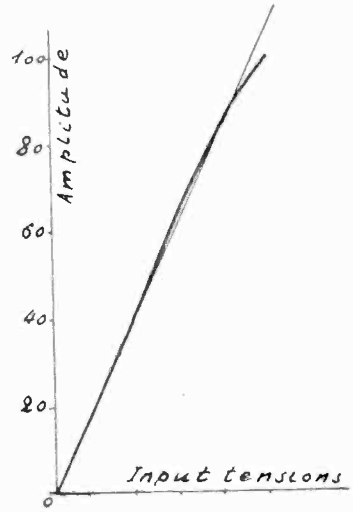


Fig. 16

The other qualities of modern transmitters such as freedom from high-frequency harmonics, scintillation, stability, ground noise, can be obtained perfectly within the limits imposed by the C.C.I.R.

Fig. 16 and 17 show the amplitude versus amplitude and amplitude versus frequency characteristics of a transmitter established with this system.

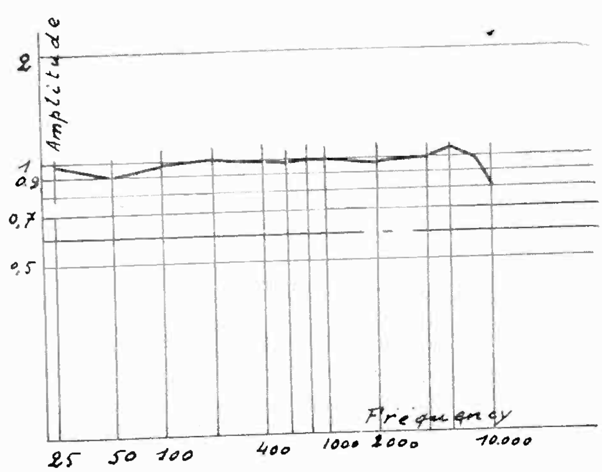


Fig. 17

On Fig. 18 are also given oscillograms taken at Radio-Paris (80 kilowatts carrier wave). These oscillograms show the high-frequency modulated wave versus the low-frequency modulation level.

Besides the principal advantage of important saving in the input power and in the tube power installed, the modulation system just described possesses a certain number of additional advantages.

First, the power tubes operating with high efficiency and with a mean power practically equal to one fourth of the maximum power dissipate very little. This kind of operation is very favorable, as the formation of hot spots on the anode is avoided, the life of the valve is increased, and the cooling problem is simplified.

Second, grid excitation being always very important, we entirely avoid negative resistance or dynatron effects generally met with in transmitters designed along the modulated high-frequency amplification principle, these effects being due to secondary emission of the grid.

Third, as may be easily realized from Fig. 7, the modulation percentage in the antenna is practically independent of the modulation percentage in the modulated stage. It is therefore possible, for instance, to obtain a 100 per cent modulation in the antenna without exceeding 70 per cent in the modulated stage. This enables one to obtain correct curves more easily.

Economic Advantages as Compared with Other Systems

Let us compare the actual broadcast transmitting systems:

(a) The so-called modulated high-frequency amplification system, modulation taking place at small power or modulation being carried on the last driving stage as in the system named "series modulation."

(b) The so-called power modulation system, the low-frequency amplifier operating with high efficiency (class B or push-push amplifier).

(c) The outphasing modulation system.

It is found from the efficiency point of view that the over-all efficiency considered as the ratio of antenna power on carrier wave to the power derived from the mains, taking into consideration all the auxiliaries, is as an average 22 per cent for system (a), 30 per cent for system (b), and 34 per cent for system (c).[4]

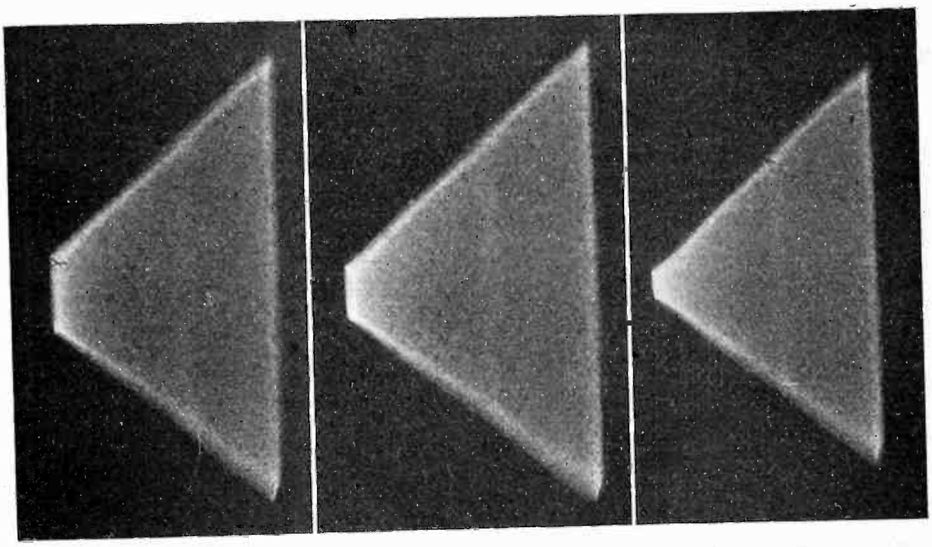
If A is the price of the kilowatt-hour and P the power of the station on carrier, the power expenses will then be equal respectively to 4.5 AP, 3.33 AP, and 2.93 AP.

We find in addition that systems (a) and (c) require the same number of power tubes.[5]

[4] The figures taken on the Luxembourg broadcast radio-transmitter (150 kilowatts carrier wave) have given almost 35 per cent.

[5] Under the condition that in system (a)—modulated high-frequency amplification—the plate dissipation is not a limiting factor this being a supplementary eventual advantage of system (c)—outphasing—in comparison with (a).

System (b), on the other hand, (power modulation) requires besides the tubes necessary for systems (a) and (c), additional tubes for the low-frequency power amplifier. It can be shown theoretically as confirmed by published results that the necessary nominal power of these extra tubes is at least two thirds of that of the tubes necessary for the high-frequency amplifier alone. As however it may be admitted that the instantaneous input voltage can be slightly increased in the case of anode control, we shall assume that the supplementary expense in tubes is only one half of the main expense. Calling B the price of the tube kilowatt-hour expressed in terms of carrier-wave power, that is



(a) (b) (c)

Fig. 18—Radio Paris (80-kilowatt carrier wave). Modulated high frequency versus low frequency (line current).
(a) about 70 per cent modulation.
(b) about 80 per cent modulation.
(c) about 90 per cent modulation.

to say, the hourly depreciation of the tube installation divided by the carrier-wave power, we arrive at the following tube expenses in cases (a), (b), and (c):

These quantities permit us to plot the curves of Fig. 19 which show in relation to A/B:

Curve 1—The saving in maintenance expenditure of the outphasing modulation as compared with the modulated high-frequency amplification system.

Curve 2—The same saving when outphasing modulation is compared with class B anode control modulation (push-push).

Curve 3—The economic advantage of so-called class B modulation in comparison with the high-frequency amplification system.

From these curves, we see that the outphasing modulation always brings an important advantage over modulated high-frequency modulation in the shape of a 20 to 25 per cent saving for average conditions (cost of power about half of cost of tubes). This same 25 to 20 per cent saving is also found again when comparison is made with power modulation (class B).

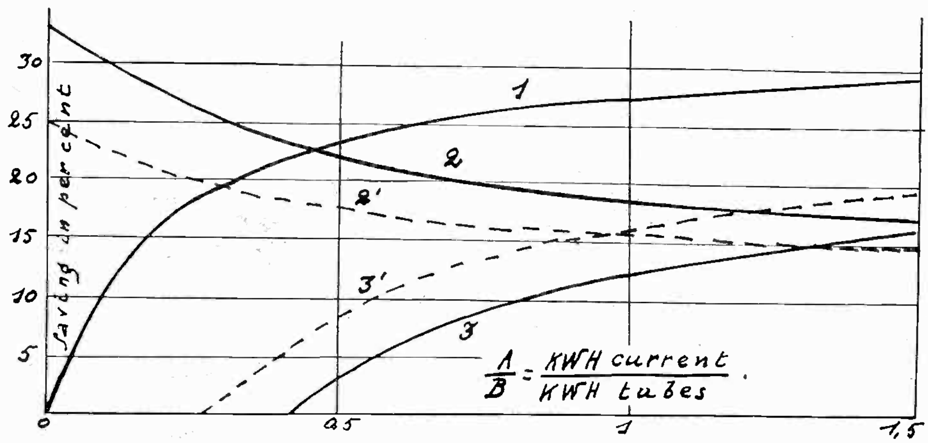


Fig. 19

In fact, it must be pointed out that for these quite average values of A/B, anode control modulation only gives a very small saving as compared with the modulated high-frequency amplification system.

We have plotted in dotted lines (curves 2′ and 3′) curves equivalent to 2 and 3 but calculated under conditions still more favorable to power modulation, by assuming that the additional tube expense is only one third of that necessary for the outphasing system.

Although these conditions seem difficult to fulfill, the saving brought by the outphasing method over the power modulation (class B) is still around 20 to 16 per cent for a variation of the ratio A/B within the limits (0.25 to 1 approximately).

We can conclude from these curves that the modulation method described in the preceding pages and applied to a certain number of high power installations in operation or under construction, brings a very important economic advantage (reaching several times ten thousand dollars a year for big installations), when it is compared with the other existing systems.

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THE STEADY-STATE RESPONSE OF A NETWORK TO A PERIODIC DRIVING FORCE OF ARBITRARY SHAPE, AND APPLICATIONS TO TELEVISION CIRCUITS*

By

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Summary—*This paper deals with the problem of determining the steady-state response of a network to a periodic driving force of arbitrary form, without the use of Fourier analysis. The periodic driving force is treated as a series of repeated transients, and the response of the network is found as the sum of the separate responses to each of the repeated transients. Formulas are developed for the general response of a network to several types of driving force encountered in scanning circuits for cathode-ray television, in terms of the general indicial response of the network. Applications to some typical scanning circuits are shown.*

Scanning circuits for cathode-ray television have brought to the fore a type of problem that has not been treated in great detail. This is the determinaion of the wave form of the steady-state response of a network to a periodic electromotive force of arbitrary form. The classical solution of this problem by Fourier analysis is not of great value, since the synthesis of the harmonic components of the response is not obvious, and to obtain the wave form, one must resort to laborious and inaccurate graphs.

For the case of lumped networks, the steady-state response to a periodic electromotive force of arbitrary form may be put in closed form by an application of the superposition theorem, treating the periodic electromotive force as an infinite succession of periodically repeated transient electromotive forces, and summing the responses of the network to each transient. Besides the analytic formulation of the electromotive force we need to know only the indicial response of the network, or the response caused by a suddenly applied unit electromotive force.

Consider first a periodic electromotive force of the type shown in Fig. 1. We propose to find the response of a network to this electromotive force between $t=o$ and $t=\tau$, assuming an infinite number of periods preceding $t=o$. This response will fall naturally into two parts, the first, $R_0(t)$, due to the electromotive force $E_0(t)$ acting during this period, and the second, $R'(t)$, the sum of the responses due to $E_1(t)$,

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$E_2(t)$, etc., acting during the preceding periods. This latter response will be represented by

$$R'(t) = R_1(t) + R_2(t) + R_3(t) + \cdots \tag{1}$$

where $R_1(t)$ is the response due to the period $-\tau<t<0$, $R_2(t)$ due to the period $-2\tau<t<-\tau$, etc. Since the electromotive forces acting in these periods are of the same form the individual responses of (1) will differ only in their arguments. Thus,

$$\left.\begin{aligned} R_2(t) &= R_1(t+\tau) \\ R_3(t) &= R_1(t+2\tau) \\ &\text{etc.} \end{aligned}\right\}. \tag{2}$$

Hence it is necessary only to determine $R_1(t)$ for the particular network and electromotive force, and then to obtain $R'(t)$ from (1) and (2).

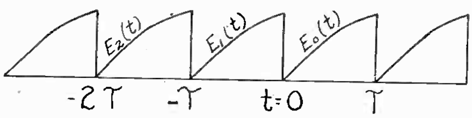

Fig. 1

If the indicial response of the network is $A(t)$, the response produced by $E_0(t)$ is given by one form of the superposition theorem[1]:

$$R_0(t) = E_0(0)A(t) + \int_0^t A(t-\lambda)\frac{d}{d\lambda}E_0(\lambda)d\lambda. \tag{3}$$

To find $R_1(t)$, the response due to the preceding period, $-\tau<t<0$, we must introduce an electromotive force $-E_1(t)$ at $t=0$. The transient electromotive force of this period is then made up of the electromotive forces $E_1(t)$ introduced at $t=-\tau$, and $-E_1(t)$ introduced at $t=0$.

The response due to $E_1(t)$ introduced at $t=-\tau$ is, from (3),

$$R_1'(t) = E_1(-\tau)A(t+\tau) + \int_{-\tau}^t A(t-\lambda)\frac{d}{d\lambda}E_1(\lambda)d\lambda \tag{4}$$

and that due to $-E_1(t)$ introduced at $t=0$ is

$$R_1''(t) = -E_1(0)A(t) - \int_0^t A(t-\lambda)\frac{d}{d\lambda}E_1(\lambda)d\lambda. \tag{5}$$

[1] V. Bush, "Operational Circuit Analysis," p. 56, John Wiley and Sons.

Adding (4) and (5), we obtain for the response $R_1(t)$ for $t>0$,

$$R_1(t) = E_1(-\tau)A(t+\tau) - E_1(0)A(t) + \int_{-\tau}^{0} A(t-\lambda)\frac{d}{d\lambda}E_1(\lambda)d\lambda. \quad (6)$$

The complete steady-state response is then, for $0<t<\tau$,

$$R(t) = R_0(t) + R_1(t) + R_1(t+\tau) + R_1(t+2\tau) + \cdots. \quad (7)$$

Generalized Network Response to a Saw-Tooth Electromotive Force

In television scanning circuits, the most important voltage wave form is the saw-tooth voltage shown in Fig. 2. For this voltage function

$$\begin{aligned} E_0(t) &= Kt \\ E_1(t) &= K(t+\tau). \end{aligned} \quad (8)$$

The indicial response of most networks commonly encountered is given by Heaviside's expansion theorem in the form[1]

$$A(t) = A_0 + \sum_{i=1}^{n} A_i \epsilon^{p_i t} \quad (9)$$

where the summation is taken over all of the n roots of the determinantal equation of the network. Equation (9) does not hold for the case of equal roots. Let us apply the electromotive force of (8) to the generalized network whose indicial response is given by (9).

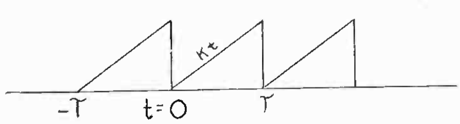

Fig. 2

Substituting the above form for $A(t)$ in (3), and noting that $E_0(0)=0$, we obtain

$$\begin{aligned} R_0(t) &= \int_0^t \left(A_0 + \sum_{i=1}^{n} A_i \epsilon^{p_i(t-\lambda)}\right) K d\lambda \\ &= A_0 Kt + \sum_{i=1}^{n} \frac{KA_i}{p_i}(\epsilon^{p_i t} - 1). \end{aligned} \quad (10)$$

Substituting $A(t)$ in (6), and noting that $E_1(-\tau)=0$ and that $E_1(0) = K\tau$, we obtain

[1] *Loc. cit.*, p. 86.

$$R_1(t) = -K\tau\left(A_0 + \sum_{i=1}^{n} A_i \epsilon^{p_i t}\right) + \int_{-\tau}^{0}\left[A_0 + \sum_{i=1}^{n} A_i \epsilon^{p_i(t-\lambda)}\right] K d\lambda$$

$$= \sum_{i=1}^{n} K A_i \epsilon^{p_i t}\left(\frac{1}{p_i}(\epsilon^{p_i \tau} - 1) - \tau\right). \tag{11}$$

From (7) the steady-state response from $t=0$ to $t=\tau$ is given by

$$R(t) = A_0 K t + \sum_{i=1}^{n} \frac{KA_i}{p_i}(\epsilon^{p_i t} - 1)$$
$$+ \sum_{i=1}^{n} KA_i\left[\frac{1}{p_i}(\epsilon^{p_i \tau} - 1) - \tau\right](\epsilon^{p_i t} + \epsilon^{p_i(t+\tau)}$$
$$+ \epsilon^{p_i(t+2\tau)} + \cdots). \tag{12}$$

Since,

$$\epsilon^{p_i t} + \epsilon^{p_i(t+\tau)} + \epsilon^{p_i(t+2\tau)} + \cdots = \frac{\epsilon^{p_i t}}{1 - \epsilon^{p_i \tau}} \tag{13}$$

equation (12) becomes

$$R(t) = A_0 K t - K\sum_{i=1}^{n} \frac{A_i}{p_i} - K\sum_{i=1}^{n} \frac{A_i \tau}{1 - \epsilon^{p_i \tau}} \epsilon^{p_i t}. \tag{14}$$

Equation (14) gives in closed form the steady-state response to a saw-tooth voltage of any network whose indicial response is of the form given by (9). Examples of the application of this equation will be given below. For the present, we turn to another type of electromotive force commonly encountered in television work.

Response of the Generalized Network to the Pulse Electromotive Force

The pulse electromotive force shown in Fig. 3 is slightly more complicated, in that each period contains two parts of interest. This function is described by

$$\left.\begin{aligned} E_0(t) &= E,\ 0 < t < \tau_1 \\ E_0(t) &= 0,\ \tau_1 < t < \tau \\ E_1(t) &= E,\ -\tau < t < -\tau + \tau_1 \\ E_1(t) &= 0,\ -\tau + \tau_1 < t < 0 \end{aligned}\right\}. \tag{15}$$

From inspection, it may be seen that the steady-state response of the generalized network of (9) to the pulse electromotive force is, from $t=0$ to $t=\tau_1$,

$$R_{01}(t) = EA_0 + E\sum_{i=1}^{n} A_i\epsilon^{p_i t} - EA_0 - E\sum_{i=1}^{n} A_i\epsilon^{p_i(t+\tau-\tau_1)}$$

$$+ EA_0 + E\sum_{i=1}^{n} A_i\epsilon^{p_i(t+\tau)}$$

$$- EA_0 - E\sum_{i=1}^{n} A_i\epsilon^{p_i(t+2\tau-\tau_1)} + \cdots$$

and, by (13),

$$R_{01}(t) = EA_0 + E\sum_{i=1}^{n} \frac{A_i(1 - \epsilon^{p_i(\tau-\tau_1)})}{1 - \epsilon^{p_i\tau}} \epsilon^{p_i t} \tag{16}$$

From $t=\tau_1$ to $t=\tau$ the steady-state response is

$$R_{02}(t) = R_{01}(t) - EA_0 - E\sum_{i=1}^{n} A_i\epsilon^{p_i(\tau-\tau_1)}$$

$$= E\sum_{i=1}^{n} A_i \frac{1 - \epsilon^{-p_i\tau_1}}{1 - \epsilon^{p_i\tau}} \epsilon^{p_i t}. \tag{17}$$

Examples of Network Response to These Driving Forces

In magnetic deflection of cathode-ray beams for television scanning a saw-tooth current through a suitable deflection coil is used. A first approximation to this saw-tooth current may be produced by applying a saw-tooth voltage to the grid of an amplifying tube, with the de-

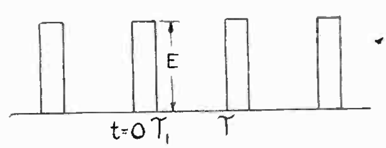

Fig. 3

flection coil in the plate circuit. The simplest circuit to consider in this connection is a series R and L arrangement, where R is the sum of the internal and external resistances in the plate circuit, and L the inductance of the deflection coil. For this circuit, the indicial current, for a suddenly applied unit voltage on the grid, is

$$I(t) = \frac{\mu}{R}(1 - \epsilon^{-Rt/L}). \tag{18}$$

This is of the form of (9), so we may substitute in (14) and obtain the plate current variation for a saw-tooth grid voltage $e_g = Kt$. The steady-state variation will then be given by

$$i(t) = \frac{\mu K}{R}\left(t - \frac{L}{R} + \frac{\tau}{1 - \epsilon^{-R\tau/L}}\,\epsilon^{-Rt/L}\right). \tag{19}$$

Introducing $f = 1/\tau$, the frequency of the saw-tooth voltage, and $\alpha = R/L$, equation (19) may be written

$$i(t) = \frac{\mu K}{R}\left[t - \frac{1}{\alpha} + \frac{1}{f}\cdot\frac{1}{1 - \epsilon^{-\alpha/f}}\,\epsilon^{-\alpha t}\right]. \tag{20}$$

Fig. 4 shows equation (20) plotted for a typical amplifying tube and deflection coil system. It will be observed that the major distortion in this case is due to the exponential term of (20), which

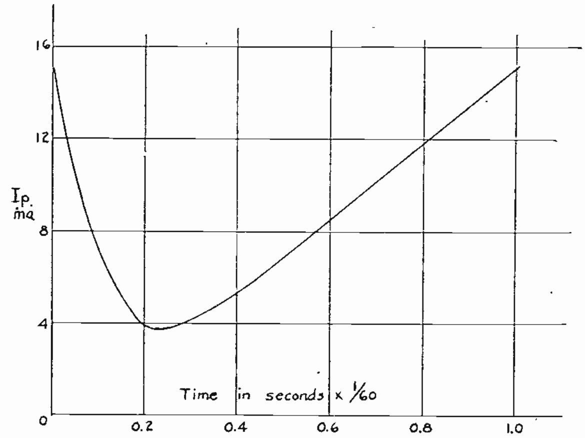

Fig. 4—Variation of plate current through a deflection coil of inductance $L = 100$ henrys, with a type 47 tube, $\mu = 150$, $R = 60{,}000$ ohms, due to a 60-cycle saw-tooth voltage, $e_g = 360t$, applied to the grid.

produces a relatively long retrace time, corresponding to the steep part of the voltage wave. If $\epsilon^{-\alpha/f}$ is small compared with unity, the retrace time, from $t = 0$ to the instant when the current is a minimum, is given by the approximate relation

$$t_{\text{retrace}} = -\frac{2.30}{\alpha}\log_{10}\frac{}{\alpha}.$$

It is also seen that a large value of α is necessary for a short retrace time. For a given inductance of the deflection coil, which is determined by the magnetic deflection sensitivity of the deflection coil, and the maximum plate current variation allowed, the tube having the greatest plate resistance will give the shortest retrace time. For this reason, tetrodes or pentodes are almost a necessity, if a saw-tooth voltage is used on the grid, and a short retrace time is desired. This is, of course, in line with the steady-state harmonic analysis of the situation.

Fortunately, the retrace time for magnetic deflection may be made as small as we wish, within limits, by the device of combining the saw-tooth electromotive force with a pulse at the end of each period, to take care of the retrace. A few words will serve to make this clear. Suppose that the saw-tooth grid voltage is increasing in the positive direction. Then the plate-current variation in the inductance will produce a voltage which lowers the plate voltage of the tube. When the grid voltage suddenly decreases at the end of the period, the rapid decrease of plate current causes the plate potential to rise to a high value. Thus the plate current cannot decrease as rapidly as the grid voltage. If, however, a large negative pulse is supplied at the end of the period, the plate current can be made to reach its minimum value within the time of duration of the pulse. Thus, we may employ an input voltage of the form shown in Fig. 5. This voltage may be considered as the

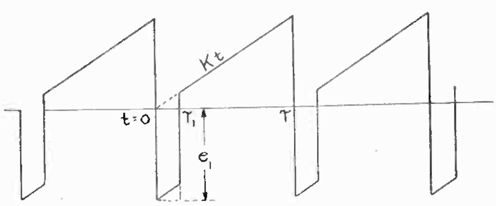

Fig. 5—Wave form of the grid voltage necessary to produce a saw-tooth plate current variation through an inductive load.

sum of the preceding saw tooth, and a periodic rectangular pulse of magnitude $-e_1$ and duration τ_1. Computing the response due to the pulses by (17), it is found that they introduce a component in the steady-state response which is given by

$$i_1(t) = \frac{\mu e_1}{R} \frac{1 - \epsilon^{\alpha\tau_1}}{1 - \epsilon^{-\alpha\tau}} \epsilon^{-\alpha t} \tag{21}$$

when $t > \tau_1$. The distortion term of (19) is

$$i_2(t) = \frac{\mu K}{R} \frac{\tau}{1 - \epsilon^{-\alpha\tau}} \epsilon^{-\alpha t}. \tag{22}$$

Adding these terms, it is seen that they cancel if

$$e_1 = \frac{K\tau}{\epsilon^{\alpha\tau_1} - 1} \tag{23}$$

or, if $\alpha\tau_1 \ll 1$, as it is in practical cases,

$$e_1 = K\tau/\alpha\tau_1. \tag{24}$$

Thus, by adding to the saw tooth a pulse of the proper magnitude, the retrace time may be limited to the duration of the pulse, and distortion is absent over the rest of the cycle. In practicable circuits, the limiting duration of the retrace time is usually determined by the magnitude of the very high plate voltage produced during the retrace time. It is also difficult to produce rectangular pulses but this is not

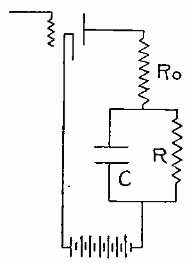

Fig. 6—Circuit by which the wave form of Fig. 5 is obtained from a series of voltage pulses applied to the grid.

of great importance. Distributed capacity in the deflection coils may be sufficient to produce oscillations during the retrace time, but these can usually be eliminated by a shunt resistance of high value across the coils.

It is of interest here to describe one method by which this combination of saw tooth and pulse may be produced. Referring to Fig. 7

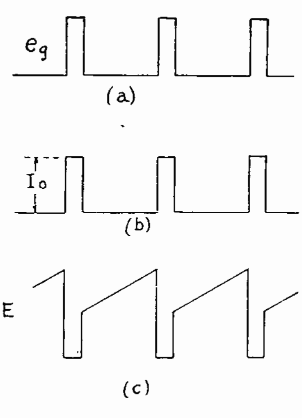

Fig. 7—Wave forms of the grid-voltage—plate-current, and plate-voltage variations occurring in the circuit of Fig. 6.

(a) we may introduce a periodic pulse on the grid of the tube, the pulse coming from any suitable relaxation oscillator. We will suppose that, without the pulse, the tetrode of Fig. 6 is biased below cutoff. The introduction of a positive rectangular pulse will cause a corresponding pulse of plate current, as shown in Fig. 7 (b). Let us now determine the variation voltage across the plate impedance. The indicial voltage produced by a suddenly applied unit current flowing through this impedance will be given operationally by

$$E_1(t) = R_0 + R/(pRC + 1) \tag{25}$$

which is equivalent to

$$E_1(t) = R_0 + R(1 - \epsilon^{-t/RC}). \tag{26}$$

Assuming a series of current pulses of magnitude I_0, the steady-state voltage across the impedance is found from (16) and (17) to be

$$E = I_0\left[R_0 + R - \frac{R(1 - \epsilon^{-(\tau-\tau_1)/RC})}{1 - \epsilon^{-\tau/RC}}\,\epsilon^{-t/RC}\right] \tag{27}$$

$$0 < t < \tau_1$$

and,

$$E = -R\frac{1 - \epsilon^{\tau_1/RC}}{1 - \epsilon^{-\tau/RC}}\,\epsilon^{-t/RC} \tag{28}$$

$$\tau_1 < t < \tau.$$

If $\tau_1/RC \ll 1$ and $\tau_1 \ll \tau$ equations (27) and (28) may be approximated by

$$E = I_0R_0, \qquad 0 < t < \tau_1 \tag{29}$$

$$E = \frac{I_0\tau_1}{C}\,\frac{\epsilon^{-t/RC}}{1 - \epsilon^{-\tau/RC}}, \qquad \tau_1 < t < \tau. \tag{30}$$

The voltage applied to the grid of the amplifying tube then has the form shown in Fig. 7 (c) and is the desired combination of pulse and saw tooth. The value of RC is adjustable, of course, so that only the first, and nearly linear, portion of the exponential is used. It is interesting to note that, as long as the approximations hold, the magnitude of the pulse component is, from (27), independent of R and C, and further, that the peak value of the saw-tooth component, from equation (30), is independent of R_0, and is proportional to $I_0\tau_1$, the magnitude of the charge delivered to the RC combination during each pulse.

Other Causes of Distortion in Magnetic Deflection Circuits

The simple R, L series circuit treated above has the disadvantage of a constant direct-current component of current in the deflection coils, which produces a constant deflection of the cathode-ray beam. This deflection may be compensated by another opposing coil carrying a constant current, but since this complicates the deflecting system, and means a greater drain on the power supply, it is advantageous to use a circuit such as is shown in Fig. 8, where the direct-current component is blocked by the condenser C and supplied by the shunt inductance L_1.

It is obvious from harmonic analysis that C should offer negligible impedance to the fundamental component of the saw-tooth current wave, and that L_1 should offer a high impedance. However, for picture frequency scanning currents, between twenty and forty cycles, these conditions are difficult to fulfill, and it is of interest to examine the amount of distortion produced by departures from the ideal.

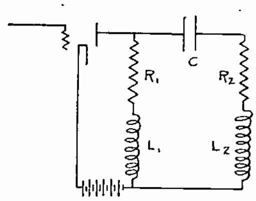

Fig. 8—Circuit used for blocking the direct-current component of the plate current.

Considering the shunt inductance first, let us assume that C is so large that negligible distortion is caused by it. The equivalent circuit is then that of Fig. 9 (a), where R_0 is the plate resistance of the tube. In this circuit, we have the operational relations

$$i_2(R_2 + pL_2) = i_1(R_1 + pL_1) \tag{31}$$

and,

$$i_2 = i_0 \frac{1}{1 + \frac{L_2}{L_1} \cdot \frac{R_2/L_2 + p}{R_1/L_1 + p}}. \tag{32}$$

Now, if

$$\frac{R_1}{L_1} = \frac{R_2}{L_2}, \tag{33}$$

the circuit of Fig. 9(a) is equivalent to that of 9(b), a series R and L circuit, which we have already treated, and in which i_0 is linear if the combination of saw-tooth and pulse voltage is applied. Also from (32),

$$i_2 = \frac{L_1}{L_1 + L_2} \cdot i_0 \tag{34}$$

and we see that i_2 is linear with time. Hence if the resistance and inductance of the shunt arm are adjusted according to (33), no distortion will be produced by this arm. If (33) is not fulfilled, the time constants of the two shunt arms will be different, and a distortion term is present, Since this case may be treated in the manner shown in the next paragraph, it will not be developed here.

To examine the effect of the condenser, let us suppose that the effect of the shunt inductance is made negligible by a large value of L_1. Then the equivalent circuit becomes that of Fig. 9(c). The current i_2 is given operationally by

$$i_2 = pC\mu e_g/(p^2L_2C + pC(R_0 + R_2) + 1). \tag{35}$$

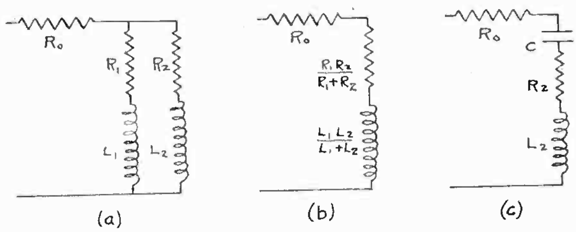

Fig. 9—Equivalent circuit elements of Fig. 8.

We now make use of a well-known operational formula.[1] This is

$$\frac{p}{ap^2 + bp + c} = \frac{1}{2a\beta}\left[\epsilon^{(-\alpha+\beta)t} - \epsilon^{-(\alpha+\beta)t}\right] \tag{36}$$

where,

and,

$$\left.\begin{aligned}\alpha &= b/2a\\ \beta &= \sqrt{\alpha^2 - c/a}\end{aligned}\right\}. \tag{37}$$

Applying this to (35), we obtain for the indicial current

$$A(t) = \frac{\mu e_g}{2L_2\beta}\left[\epsilon^{(-\alpha+\beta)t} - \epsilon^{-(\alpha+\beta)t}\right] \tag{38}$$

where,

and,

$$\left.\begin{aligned}\alpha &= \frac{R_0 + R_2}{2L_2}\\ \beta &= \sqrt{\alpha^2 - 1/L_2C}\end{aligned}\right\}. \tag{39}$$

Using (14) we have for the steady-state response of this circuit to a saw-tooth voltage in terms of the current through the deflection coil

$$i_2 = \frac{\mu K}{2L_2\beta}\left[\frac{\tau}{1 - \epsilon^{-(\alpha+\beta)\tau}}\cdot\epsilon^{-(\alpha+\beta)t} - \frac{\tau}{1 - \epsilon^{(-\alpha+\beta)\tau}}\cdot\epsilon^{(-\alpha+\beta)t} - 2\beta L_2C\right]. \tag{40}$$

There is no term in (40) which is proportional to t, as there is in

[1] *Loc. cit.*, p. 133.

(20). However, if β is made nearly equal to α, so that $(-\alpha+\beta)$ is small, $\epsilon^{(-\alpha+\beta)t}$ may be replaced by the first three terms of its series expansion,

$$\epsilon^{(-\alpha+\beta)t} = 1 - (\alpha - \beta)t + \frac{(\alpha - \beta)^2}{2} t^2 + \cdots. \tag{41}$$

Thus, $\epsilon^{(-\alpha+\beta)t}$ in (40) yields a term proportional to t, plus a term in t^2 which causes a flattening of the slope of the response towards the end of the voltage period. The term containing $\epsilon^{-(\alpha+\beta)t}$ has already been encountered in the series R and L circuit; this term causes a slow retrace time, and may be removed as before by adding a pulse to the saw tooth.

From (41) we may estimate the amount of distortion caused by a finite $(-\alpha+\beta)$. Equation (41) may be written as

$$\epsilon^{(-\alpha+\beta)t} = 1 - (\alpha - \beta)t\left[1 - \frac{(\alpha - \beta)}{2} t\right]. \tag{42}$$

At the end of the period, then, the response has departed from the ideal by the fraction

$$s = \frac{\alpha - \beta}{2} \tau$$

and, from (39), since $\alpha-\beta$ is a small quantity,

$$\begin{aligned} \alpha - \beta &= \alpha(1 - \sqrt{1 - 1/\alpha^2 L_2 C}) \\ &= 1/(R_0 + R_2)C \end{aligned} \tag{43}$$

and the fractional departure from linearity at the end of the period is

$$s = \frac{\tau}{2(R_0 + R_2)C} = \frac{1}{2fC(R_0 + R_2)} \tag{44}$$

where, as before, $f = 1/\tau$ is the frequency of the driving voltage. The distortion in this circuit is then inversely proportional to the frequency, coupling capacity C, and total plate resistance.

It may be objected that separating the elements of the circuit of Fig. 9(a) in the way we have done yields results which are different from the behavior of the complete circuit. The above results, however, are accurate to the first order of approximation. The circuit of Fig. 9(a) may be analyzed in its entirety for particular values of the parameter without trouble. The general treatment of the complete circuit is too complicated to have much meaning, since the determination of its indicial response involves the solution of a cubic equation.

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DISCUSSION ON "ECHOES OF RADIO WAVES"*

N. Janco

G. Builder:[1] The author has put forward a theory of long delay echoes based on the magnetoionic theory. In doing so, he has completely misinterpreted the theory as expounded by Appleton and Mary Taylor, to whose original papers he refers. The difficulty has clearly arisen through neglecting the exact significance of the accepted definition of the sense of rotation of an elliptically polarized wave. Appleton defines the sense of rotation with reference *to an observer looking in the direction of propagation of the wave.*

Now, in the magnetoionic theory, it generally happens that a wave traveling in the ionized medium is split up into two components having different phase and group velocities and suffering different attenuations. For simplicity, consider propagation in the direction of the magnetic field. The two components are then circularly polarized and have opposite senses of rotation. Of the two components one is termed the ordinary and the other the extraordinary. The ordinary wave is *that which is the less affected by the earth's field.* That is to say, its propagation differs less from that of propagation in the absence of the impressed magnetic field.

Consider a wave incident vertically on a horizontal layer of ionized gas, such as exists in the upper atmosphere. In general, the wave is split into two component waves having elliptic polarizations. In the northern hemisphere the ordinary component has a right-handed sense of rotation when the wave is traveling upwards. At reflection there is a reversal in the sense of rotation and the ordinary component has a left-handed sense of rotation when the wave is traveling downwards. Thus this component will be affected in the same way by a layer of ionization whether it is traveling upwards or downwards, since the sense of rotation with respect to the earth's magnetic field is the same in the two cases. Thus, if the ionization of the lower layer is such that the extraordinary wave is reflected and the ordinary wave is not, the ordinary wave continues upward and is reflected at the upper layer. When it meets the lower layer as it travels downward, it again penetrates this layer and reaches the ground.

There is abundant experimental support for this interpretation of the theory. Observations, by Appleton and Builder, Ratcliffe and White, and others, are available. For the present purpose, it is only necessary to point out frequent occurrence of simultaneous reflections, from the E and F layers, of the extraordinary and ordinary wave respectively, which has been observed on wavelengths less than about 200 meters. In such cases the ordinary wave, which requires the greater electron density for reflection, penetrates the lower layer and is reflected at the upper, while the extraordinary continues to be reflected by the smaller ionization density in the lower layer. Now, if the author's interpretation of the theory were correct, this would clearly not give rise to simultaneous echoes from the two layers, since the ordinary wave, after reflection from the upper layer, would become extraordinary and could not penetrate the lower layer and so reach the ground.

Thus this single experimental observation invalidates the author's theory. It appears to be unfortunate that, in the current literature, the extraordinary

* Proc. I.R.E., vol. 22, p. 923; July, (1934).
[1] University of Sydney, Sydney, Australia.

wave is loosely referred to, for conditions existing in the northern hemisphere, as having a left-handed sense of rotation. This practice is, of course, due to the truth of the statement for reception at the ground of the downcoming waves. In the same way, in the southern hemisphere the ordinary wave has a right-handed sense of rotation in the downcoming ray. Thus the definition is in some respects unsatisfactory but appears to be universally accepted, often without due regard to its significance. It may be noted that, as Ratcliffe and White have already pointed out,[2] the sense of rotation of each component is unchanged at reflection if the sense is referred to the direction of the earth's magnetic field. Moreover, with respect to the earth's field the sense of rotation for each magneto-ionic component would be the same for workers in both hemispheres. It is perhaps worth considering the advisability of a revised definition.

[2] J. A. Ratcliffe and E. L. C .White, *Phil. Mag.*, vol. 16, p. 138; July, (1933).

Proceedings of the Institute of Radio Engineers
Volume 23, Number 11 *November, 1935*

BOOK REVIEWS

Information for the Amateur Designer of Transformers for 25 to 60 Cycle Circuits, by H. B. Brooks, National Bureau of Standards, Circular No. C 408, June, 1935. Price five cents.

This circular describes practical methods for designing and constructing power transformers with good efficiency and reasonable cost, using home facilities. Extensive details are given on various problems, such as the selection of core materials and size, wire sizes, and insulation methods. It will enable the home constructor to go about the job without tedious preliminary mathematics. In the matter of insulation however, it is this reviewer's opinion that less transformer failures will result by providing an insulating paper layer between successive layers of wire in all cases, irrespective of potential gradient per layer, since amateur winding methods do not insure that the wire keeps in layers. Details concerning the insulation requirements for high voltage windings (several kilovolts) are less complete, as are also the methods for providing electrostatic shields between windings.

*RALPH R. BATCHER

The National Physical Laboratory Report for the Year 1934. Published by His Majesty's Stationery Office for the Department of Scientific and Industrial Research, 1935. 260 pages. Price 13s.

This report describes the activities of the National Physical Laboratory of England which include a wide range of scientific, technical, and engineering subjects. The report is divided into a number of sections in which the work of the following departments are treated: Physics, Electricity, Radio, Metrology, Engineering, Metallurgy, Aerodynamics, and William Froude Laboratory. Lists of the personnel and of the publications of the Laboratory for 1933 are included.

The radio engineer will be particularly interested in the report on work of the Radio Department. This is divided into sections dealing with propagation of waves, direction finding, atmospherics, and radio apparatus and materials. Among the interesting subjects treated may be mentioned ionosphere measurements, angle of incidence of downcoming signals, electronic oscillations, and effect of temperature variation on coils and condensers in transmitters and means of compensation.

The radio engineer will find material of equal interest in the section on work of the Electricity Department, particularly that on electrical standards and measurements. Here such subjects as standards of radio frequency, international comparisons of frequency, attenuation and impedance measurements and standard condensers are briefly treated. Other electrical measurements, photometry, and photoelectric cells are also treated.

Results of the comparison of two piezoelectric oscillators as time-rate standards with a Shortt clock are briefly given in the report of the Metrology Department.

Engineers having to do with acoustics will be interested in the description of the new acoustics laboratory found in the report of the Physics Department. One section of the report treats on sound with brief summaries of the work on sound transmission, reverberation, and noise measurements and the calibration of microphones and loud speakers.

†E. L. HALL

* Hollis, Long Island, N.Y.
† National Bureau of Standards, Washington, D.C.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained without charge by addressing the publishers.

Circular R-1002-C issued by Weston Electrical Instrument Corporation of Newark, N.J., covers power level indicators, high speed movements, and DB meters.

Catalog No. 18 which is of a general nature has recently been issued by Federated Purchaser, Inc., of 25 Park Place, New York; 1331 South Michigan Avenue, Chicago; and 546 Spring Street, N.W., Atlanta, Ga.

The Ward Leonard Electric Company of Mount Veron, N.Y., has issued a series of new bulletins. Bulletin 501, covers D.C. manual starters; 701 D.C. compound manual starters; 1201, D.C. manual speed regulators; 1301 and 1302, D.C. compound manual speed regulators; 8601, controlled rectifiers; and 8651, magnetic chuck rectifiers.

Bulletin No. 250 of the National Company, 61 Sherman Street, Malden, Mass., covers various components and receivers for high-frequency services.

The National Union Laboratories of 365 Ogden Street, Newark, N.J., has issued data sheets on the Type 43 power output pentode, the 84/6Z4 full wave rectifier, the 1B4 voltage amplifier, and the 1A4 voltage amplifier.

Communication Products, Inc., of 245 Custer Avenue, Jersey City, N.J. has recently issued data on "Properties and Uses of the New Q-MAX Radio-Frequency Coating Products."

The RCA Manufacturing Company, Radiotron Division of Harrison, N.J., has issued Application Note No. 51 on the 6F5, a high-mu triode. Data sheets have been issued on the 1B5/25S duodiode triode, the 6E5 electron-ray tube, and the 898 high power radio-frequency amplifier, oscillator, and class B modulator.

Cornish Wire Company of 30 Church Street, New York City has issued a brochure on all wave antennas.

A general catalog No. 188 has been issued by the Insuline Corporation of America, 23 Park Place, New York City.

A leaflet covering a new phonograph for sixteen-inch records has been issued by the Ansley Radio Corporation of 240 West 23rd Street, New York City.

Leaflets have been issued by Shure Brothers Company of 215 West Huron Street, Chicago, on their new "spheroid" nondirectional crystal microphone.

Output switching panel, 271A, for radiotelephone broadcast systems is described in a booklet issued by the Western Electric Company of 195 Broadway, New York City.

Proceedings of the Institute of Radio Engineers
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CONTRIBUTORS TO THIS ISSUE

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Schafer, J. Peter: See Proceedings for June, 1935.

Schlaack, N. F.: See Proceedings for July, 1935.

Skellett, A. M.: See Proceedings for February, 1935.

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APPLICATION FOR ASSOCIATE MEMBERSHIP

(Application forms for other grades of membership are obtainable from the Institute)

To the Board of Directors

Gentlemen:

I hereby make application for Associate membership in the Institute of Radio Engineers on the basis of my training and professional experience given herewith, and refer to the members named below who are personally familiar with my work.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I will be governed by the constitution of the Institute as long as I continue a member. Furthermore I agree to promote the objects of the Institute so far as shall be in my power, and if my membership shall be discontinued will return my membership badge.

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ARTICLE II—MEMBERSHIP

Sec. 1: The membership of the Institute shall consist of: * * * (c) Associates, who shall be entitled to all the rights and privileges of the Institute except the right to hold any elective office specified in Article V. * * *

Sec. 4: An Associate shall be not less than twenty-one years of age and shall be a person who is interested in and connected with the study or application of radio science or the radio arts.

ARTICLE III—ADMISSION AND EXPULSIONS

Sec. 2: * * * Applicants shall give references to members of the Institute as follows: * * * for the grade of Associate, to three Fellows, Members, or Associates; * * * Each application for admission * * * shall embody a full record of the general technical education of the applicant and of his professional career.

ARTICLE IV—ENTRANCE FEE AND DUES

Sec. 1: * * * Entrance fee for the Associate grade of membership is $3.00 and annual dues are $6.00.

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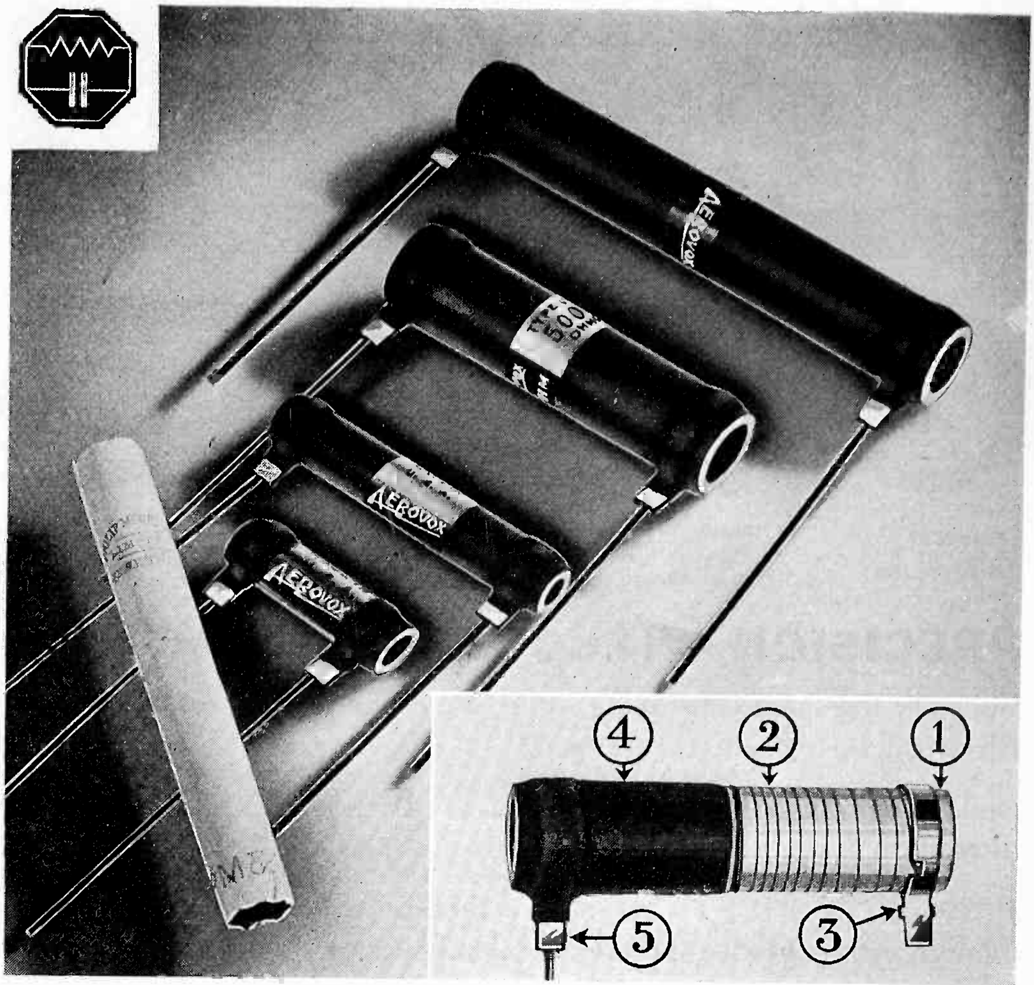
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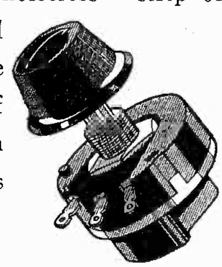

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