Proceedings of the I·R·E

MAY 1939

VOLUME 27 NUMBER 5

Institute of Radio Engineers

National Convention

San Francisco, California, June 27–30, 1939

Fourteenth Annual Convention

New York, N. Y., September 20–23, 1929

Rochester Fall Meeting

November 13, 14, and 15, 1939

New York Meeting—Engineering Societies Building—June 7

33 West 39th Street, New York, N. Y.

SECTION MEETINGS

CINCINNATI
May 21

CLEVELAND
May 25

DETROIT
May 19

LOS ANGELES
May 16

PHILADELPHIA
June 1

PITTSBURGH
May 16

PORTLAND
May, 15

WASHINGTON
May 8

SECTIONS

ATLANTA—Chairman, Ben Akerman; Secretary, J. G. Preston, 230 Ansley St., Decatur, Ga.
BOSTON—Chairman, H. W. Lamson; Secretary, E. B. Dallin, 64 Oakland Ave., Arlington, Mass.
BUFFALO-NIAGARA—Chairman, H. C. Tittle; Secretary, E. C. Waud, 235 Huntington Ave., Buffalo, N. Y.
CHICAGO—Chairman, V. J. Andrew; Secretary, G. I. Martin, RCA Institutes, 1154 Merchandise Mart, Chicago, Ill.
CINCINNATI—Chairman, H. J. Tyzzer; Secretary, J. M. McDonald, Crosley Radio Corp., 1329 Arlington, Cincinnati, Ohio.
CLEVELAND—Chairman, S. E. Leonard; Secretary, Hollie C. Williams, Rm. 1932, 750 Huron Rd., Cleveland, Ohio.
CONNECTICUT VALLEY—Chairman, E. R. Sanders; Secretary, W. R. G. Baker, General Electric Co., Bridgeport, Conn.
DETROIT—Chairman, L. C. Smeby; Secretary, R. J. Schaefer, 9753 N. Martindale, Detroit, Mich.
EMPORIUM—Chairman, R. K. McClintock; Secretary, D. R. Kiser, Hygrade Sylvania Corp., Emporium, Penna.
INDIANAPOLIS—Chairman, I. M. Slater; Secretary, B. V. K. French, P. R. Mallory & Co., E. Washington, St., Indianapolis, Ind.
LOS ANGELES—Chairman, F. G. Albin; Secretary, M. T. Smith, General Radio Co., Seward & Romaine Sts., Hollywood, Calif.
MONTREAL—Chairman, Sydney Sillitoe; Secretary, R. E. Hammond, 1261 Shearer St., Montreal, Que.
NEW ORLEANS—Chairman, G. H. Pierce; Secretary, D. W. Bowman, 8327 Sycamore St., New Orleans, La.
PHILADELPHIA—Chairman, H. J. Schrader; Secertary, R. L. Snyder, 103 Franklin Rd., Glassboro, N. J.
PITTSBURGH—Chairman, W. P. Place; Secretary, R. E. Stark, 90 Pilgrim Rd., Rosslyn Farms, Carnegie, Penna.
PORTLAND—Chairman, H. C. Singleton; Secretary, E. R. Meissner, United Radio Supply, Inc., 203 S. W. Ninth Ave., Portland, Ore.
ROCHESTER—Chairman, H. J. Klumb; Secretary, H. C. Sheve, Stromberg-Carlson Telephone Manufacturing Co., Rochester, N. Y.
SAN FRANCISCO—Chairman, F. E. Terman; Secretary, L. J. Black, 243-30th St., Oakland, Calif.
SEATTLE—Chairman, R. O. Bach; Secretary, Karl Ellerbeck, Pacific Telephone and Telegraph Co., 612 Northern Life Tower, Seattle, Wash.
TORONTO—Chairman, R. C. Poulter; Secretary, N. Potter, Canadian National Carbon Co., Ltd., 805 Davenport Rd., Toronto, Ont.
WASHINGTON—Chairman, Gerald C. Gross; Secretary, M. H. Biser, Capitol Radio Engineering Institute, Riggs Bank Bldg., Washington, D. C.

Proceedings of the I·R·E

Published Monthly by
The Institute of Radio Engineers, Inc.

VOLUME 27 — *May, 1939* — NUMBER 5

Entered as second-class matter October 26, 1927, at the post office at Menasha, Wisconsin, under the Act of February 28, 1925, embodied in Paragraph 4, Section 412 of the Postal Laws and Regulations. Publication office, 450 Ahnaip Street, Menasha, Wisconsin. Subscription, $10.00 per year; foreign, $11.00.

THE INSTITUTE

The Institute of Radio Engineers serves those interested in radio and allied electrical-communication fields through the presentation and publication of technical material. In 1913 the first issue of the PROCEEDINGS appeared; it has been published uninterruptedly since then. Over 1500 technical papers have been included in its pages and portray a currently written history of developments in both theory and practice.

STANDARDS

In addition to the publication of submitted papers, many thousands of man-hours have been devoted to the preparation of standards useful to engineers. These comprise the general fields of terminology, graphical and literal symbols, and methods of testing and rating apparatus. Members receive a copy of each report. A list of the current issues of these reports follows:

Standards on Electroacoustics, 1938
Standards on Electronics, 1938
Standards on Radio Receivers, 1938
Standards on Radio Transmitters and Antennas, 1938.

MEETINGS

Meetings at which technical papers are presented are held in the twenty-one cities in the United States and Canada listed on the inside front cover of this issue. A number of special meetings are held annually and include one in Washington, D. C., in co-operation with the American Section of the International Scientific Radio Union (U.R.S.I.) in April, which is devoted to the general problems of wave propagation and measurement technique, the Rochester Fall Meeting in co-operation with the Radio Manufacturers Association in November, which is devoted chiefly to the problems of broadcast-receiver design, and the Annual Convention, the location and date of which is not fixed.

MEMBERSHIP

Membership has grown from a few dozen in 1912 to more than five thousand. Practically every country in the world in which radio engineers may be found is represented in our membership roster. Approximately a quarter of the membership is located outside of the United States. There are several grades of membership, depending on the qualifications of the applicant. Dues range between $3.00 per year for Students and $10.00 per year for Members. PROCEEDINGS are sent to each member without further payment.

PROCEEDINGS

The contents of each paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Material appearing in the PROCEEDINGS may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety may not be reproduced, however, without specific permission from the Institute.

Papers submitted to the Institute for publication shall be regarded as no longer confidential. They will be examined by the Papers Committee and Board of Editors to determine their suitability for publication. Suggestions on the mechanical form in which manuscripts should be prepared may be obtained from the Secretary.

SUBSCRIPTIONS

Annual subscription rates for the United States of America, its possessions, and Canada, $10.00; to college and public libraries when ordering direct, $5.00. Other countries, $1.00 additional.

The Institute of Radio Engineers, Inc.
Harold P. Westman, Secretary
330 West 42nd Street
New York, N.Y.

Precipitation-Static Interference on Aircraft and at Ground Stations*

HERBERT M. HUCKE†, ASSOCIATE MEMBER, I.R.E.

Summary—The effect of precipitation static on aircraft reception is described and a history of previous work on the subject given. The results of an expedition assembled by the United Air Lines for a study of the subject are described under a series of chapter headings. The meteorological conditions producing the static areas were explored by flights through them and the theory of their formation is discussed. Flight tests of all known types of antistatic antennas were made in bad-weather areas and their effectiveness compared. A theory that the interference results from corona produced on the plane structure was developed and tested by mounting the plane on insulators and charging it to 100,000 volts with all radio equipment and personnel on board. A method for reducing the corona was developed as a result and proved by flight tests. A study of the electromagnetic radiation from corona discharges was made with a synthetic static generator and the operation of the metallically shielded antistatic loop antennas explained and its limitations established. A commercial form of plane discharge system was developed and its advantages and limitations are described. The application of the aircraft results to ground radio reception is discussed.

I. Introduction

TWO types of atmospheric static interference are normally experienced in radio reception on aircraft. The first consists of the usual short, intermittent crashes which result from lightning discharges. The second is a continuous type of interference which may completely blanket reception for periods as long as several hours. It is usually recognized as a combination of noises containing frying sounds, intermittent or regular crackling, and a characteristic musical "crying" during which the noises run up and down the musical scale.

This second type has been classified under the general heading of Precipitation Static,[1] since its effects are usually experienced when flying through rain, snow, hail, ice crystals, and dust clouds. It may also be experienced when flying near, but not actually in, precipitating particles, but this is an exception which occurs only in unusual cases and then only for periods of a few minutes.

Precipitation static has been experienced in ground station reception for many years as indicated in a paper by Marriott[2] on reception conditions at Denver, Colorado, in 1906. Curtis[3] in 1914, discussing steamship antennas, states, "This shows that the particular sort of discharge here considered is caused by charged particles of air or water being driven against the antenna by the wind, the group frequency and also the intensity depending on the number of particles striking the antenna per second." Kennelly,[4] in his comments on the Marriott paper states, "Mr. Marriott points out that electrified water droplets impinging on the receiving aerial, could account for at least a part of the disturbance." A review of available publications on the subject shows many other references as will be brought out later.

Since the phenomena occurred only occasionally at ground stations its effects were not serious and an incentive for intense research on this type of static was lacking. Therefore, it remained largely in the realm of discussion until it became of importance to the air-transport industry almost twenty years after Marriott's paper.

From 1928 to 1934 all transport aircraft had been equipped with radio. Radio beacons had been installed and pilots had begun to depend upon them as a means of navigation. The reliability of radio navigation was seriously impaired, however, by occasional complete loss of reception for long periods while flying in precipitation. This loss of reception shortly became a contributing cause in several serious accidents and the incentive for research was established.

Various theories as to the cause of this form of static were advanced. One of these followed Curtis' idea that "... the intensity depended upon the number of particles striking the antenna per second." It was assumed that the antenna on a plane traveling at 200 miles per hour would strike a large number of charged particles per second and thus deliver a correspondingly large amount of noise to the receiver. This was substantiated in practice since pilots had been able to reduce the static intensity by throttling motors and thus decreasing the antenna velocity.

Following this line of reasoning a National Air Transport plane, equipped with an antenna covered with friction tape, was tested in 1930. In 1932 all United Air Lines planes were equipped with antennas having an extra thick rubber covering. These meth-

* Decimal classification: R521×R114. Original manuscript received by the Institute, December 1, 1938.

† Formerly, United Air Lines Transport Corporation, Chicago, Ill. Now, Acting Communications Engineer, Air Safety Board, Civil Aeronautics Authority, Washington, D. C.

[1] The term Precipitation Static was agreed upon by Subcommittee Number One of the Radio Technical Committee for Aeronautics at a meeting in Washington, D.C., on December 15, 1937.

[2] R. H. Marriot, "Radio range variation," Proc. I.R.E., vol. 2, pp. 37–52; March, (1914).

[3] Austin Curtis, Discussion on "Radio range variation," by R. H. Marriott, Proc. I.R.E., vol. 2, p. 55; March, (1914).

[4] A. E. Kennelly, Discussion on "Radio range variation," by R. H. Marriott, Proc. I.R.E., vol. 2, p. 57; March, (1914).

ods were supposed to prevent charged particles from striking the antenna wire and thus reduce the noise. Neither proved of value and other solutions were sought.

Pilots' reports indicated that the intensity of the static increased at some rate considerably greater than the first power of the plane velocity. They also indicated that the occurrence of St. Elmo's fire on propeller tips was in some manner associated with the radio interference. No practical solutions were indicated, however, until early in 1935 when D. S. Little of RCA Aviation Division recalled that the metallically covered loop direction finders used on Great Lakes boats gave reception in snowstorms when open-wire antennas were useless.

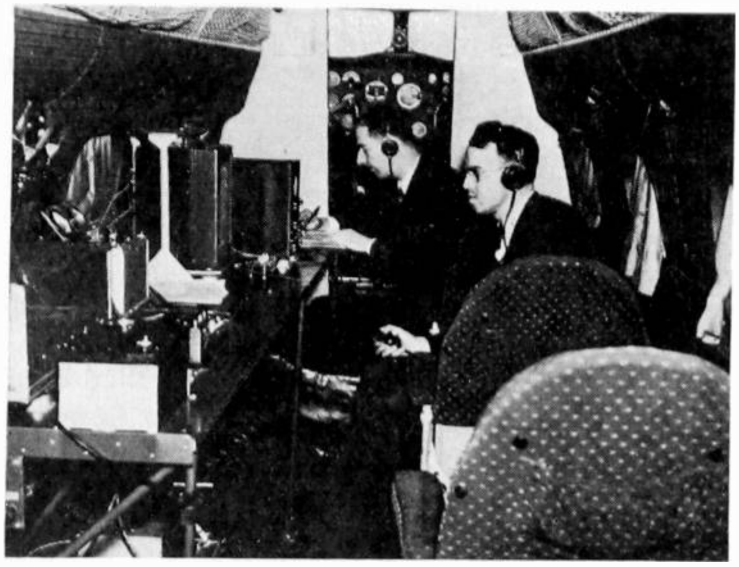


Fig. 1—Interior arrangement of laboratory plane.

Transcontinental and Western Air and United Air Lines immediately constructed metallically shielded loops suitable for aircraft and tested them through the winter of 1935 and 1936.

In October, 1935, the Snow Static Subcommittee of the Radio Technical Committee for Aeronautics met in Chicago and reviewed the available information on the subject. A study of a model plane in a wind tunnel filled with charged particles was suggested and in December, 1935, Homer Dana succeeded in reproducing the phenomena in a mild form in a small wind tunnel at Washington State College.

In March, 1936, Marcus O'Day of Reed College, Portland, Oregon, connected a recording electrometer to a sheet of copper foil attached to the inside of the windshield of a United Air Lines plane flying between Portland and Salt Lake City. The recordings proved that the rain and snow particles striking the windshield were electrostatically charged and that the magnitude of the charge bore a relationship to the noise heard in the headphones.

In July, 1936, a paper by Morgan[6] gave the results of the Transcontinental and Western Air experiments with the metallically shielded loop. Its improvement over open wire antennas was so great that Transcontinental and Western Air began equipping their entire fleet with loops in October of that year and all air-transport lines were required by a Department of Commerce order to equip their planes by October, 1937.

Morgan's paper suggests that the metallic shielding of the loop "... prevents the charged particles from striking the antenna structure" and thus allowed normal reception. He also commented that "... under some conditions of flight, the plane will build up sufficient charge to show a plainly visible corona discharge (St. Elmo's fire) and yet radio reception may be affected very slightly. When the corona discharge from parts of the plane becomes very pronounced, there will be interference, but this may actually be electromagnetic in origin from the heavy corona current flow." As will be shown later, this theory subsequently became the key to an entirely new explanation of the precipitation static phenomenon.

The United Air Lines' metallically shielded loop tests in large Pacific Northwest storm areas had shown periods as long as 30 minutes during which the loop was unable to prevent static from blocking radio beacon reception. Since the metallic shielding unquestionably prevented any particles from striking the antenna, it was apparent that some additional source of noise must be accounted for or some other explanation given. The loop was a great improvement over the old antenna but the possibility of its failure for as long as thirty minutes justified additional research toward a still better solution.

A program for flight investigation was accordingly set up in November, 1936, and continued until June, 1937. This work was carried out in a standard ten-passenger, twin-engine Boeing 247-D all-metal transport plane. As shown in Fig. 1 a workbench was installed with electrometers, oscillograph, three tandem ink recorders, and several radio receivers. A telephone system between pilots and all seats was necessary so that engineers could converse without leaving their positions in front of instruments. Numerous antennas were installed from time to time as will be shown later.

A crew consisting of engineers and physicists from Reed College, Purdue University, Oregon State College, Bendix Radio Corporation, and Bell Telephone Laboratories and pilots, engineers, and a

[6] H. W. Morgan, "Rain static," Proc. I.R.E., vol. 24, pp. 959–963; July, (1936).

meteorologist from United Air Lines manned the plane from time to time during the flights.

The problem was attacked simultaneously on three general fronts to determine, first, the meteorological aspects of static formation and its avoidance; second, the value of special antennas for reducing the interference; and third, the static-generating effect of the plane and its reduction. These three classifications of the problem are covered in the sections which follow.

II. Meteorological Aspects

All atmospheric static is caused primarily by disturbances in the electrostatic field which surrounds the earth. While the theories regarding these disturbances are not entirely complete, the following viewpoint will be of assistance in understanding the problem.

The earth is a huge ball floating in free space, and as such, it gathers an electric charge which is stored in the atmosphere which surrounds it. The potential gradient near the surface is normally about 35 volts per foot of altitude. This tapers off as the atmosphere becomes thinner with altitude until at 20,000 feet it is about 3 volts per foot. The total voltage between the earth and the outermost reaches of the atmosphere has been estimated[6] at about 1,000,000 volts. As long as this charged atmosphere remains evenly distributed, a plane may fly through it without suffering from static interference.

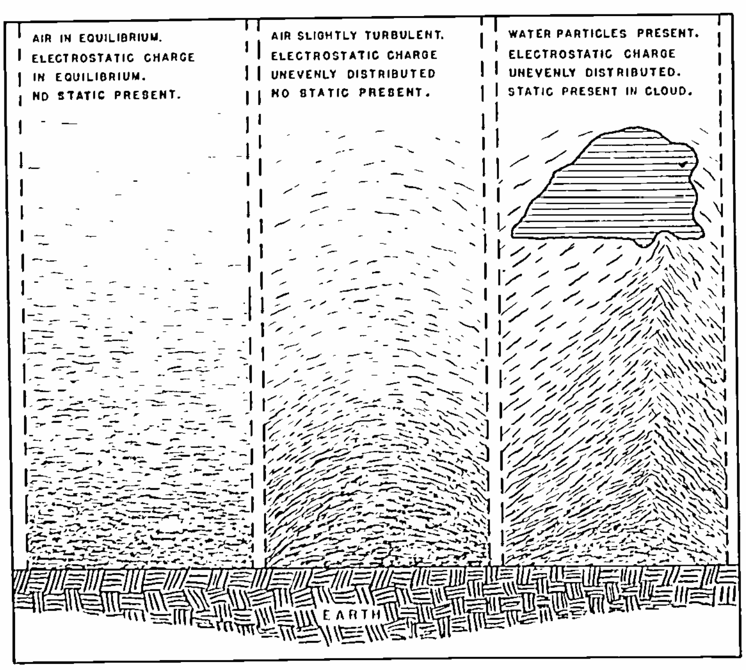


Fig. 2—Air warmed by the sun rising to form a thundercloud.

The air surrounding the earth is not normally in equilibrium due to the sun's rays heating portions of it unequally and causing it to rise (see Fig. 2). As it rises, the electrostatic field becomes distorted, but under average conditions readjusts itself and retains reasonable equilibrium. The rising air, however,

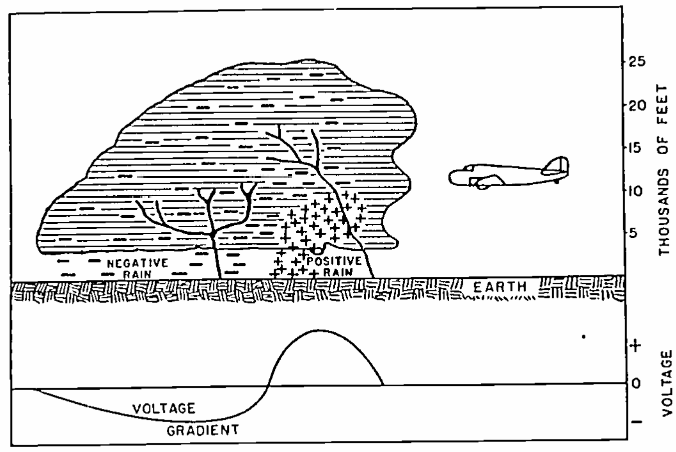


Fig. 3—Distribution of electrical charges in a thundercloud of the heat type.

brings up moisture which condenses to fog and forms clouds.

If the fog forms slowly the electrostatic charges on its particles retain an even distribution. If it forms rapidly with turbulent air currents, the water droplets are churned about and the cloud becomes unstable electrostatically. Lightning will occur if sufficient electrostatic instability results, thus forming the common thunderstorm.

When water droplets are carried about by the wind in a cloud they are usually split up into larger and smaller units. Tests indicate that the large droplet is usually positive while the fine spray which is separated from it by the wind is usually charged negatively. There is considerable question as to whether the mechanical action of splitting the droplets produces the difference in electric charge or whether the difference in charge results from the fine spray being carried upward while the heavier droplets are carried downward due to gravity. In the second case it would seem that their position with respect to the earth's electrostatic field could produce a difference in charge. There is good possibility that differently polarized droplets result from the combination of the several actions rather than by any single method.

The general theory of thunderstorm formation as first described by Simpson[7] in England gives a good picture of the situation. In Fig. 3 the warm air enters at the bottom of the cloud and rises toward its top. Actually, the cloud is the result of this rising air entering a cooler portion of the earth's atmosphere. Considerable turbulence results in the rising portion of the cloud and water particles are often carried

[6] W. J. Humphries, "Physics of the Air," pp. 382–393. McGraw-Hill Book Company, New York, N.Y., (1929).

[7] W. J. Humphries, "Physics of the Air," pp. 300–305. McGraw-Hill Book Company, New York, N.Y., (1929).

upward and fall back into the rising current a number of times. If the churning action goes on for a sufficient length of time the water droplets may be built up into large hailstones. When their weight becomes great enough so that the rising air can no longer support them, they fall out of the cloud. The distribution of electric charges on the droplets in the cloud is usually as indicated in Fig. 3. If we fly through such a cloud

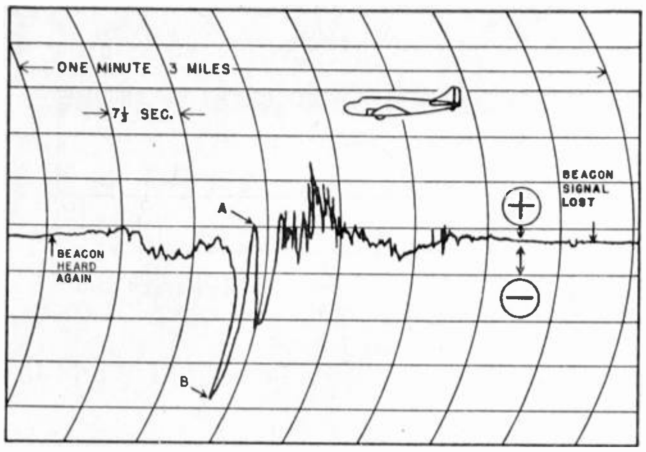


Fig. 4—Recording of electrostatic charge on a plane flying through a very small heat-type thundercloud. Point B is estimated to be several hundred thousand volts negative.

(from right to left) we should theoretically record a gradual change from positive to negative as the two types of charged areas are passed.

Actually, such a simple cloud does not exist in nature. In the course of our work we flew through and recorded the voltages of about fifty clouds in a number of different directions and at different altitudes. Most cloud records are too long to reproduce in a single illustration but Fig. 4 shows the condition found in a very small cloud. It is interesting to note that the distribution of charged droplets in a cloud is vastly more random than the theoretical Simpson cloud indicates. Our flights indicate that the interior of a cloud is in constant motion and consequently the charged droplets are undergoing continuous change of position. The flight tests also indicated that if any comparison of antennas is to be of value it must be made by having the antennas on the same plane and switching from one antenna to another in a second or less. Switching back and forth between two antennas must be repeated many times before comparisons can be trusted. For example, if one antenna were tested at point A and another at point B about one second later, the conclusions would be erroneous. The cloud shown in the illustration was too small for practical antenna comparisons and most useful results were obtained in much larger clouds.

From a meteorological standpoint there are roughly two types of clouds in which static will be formed. The first is the simple thundercloud, formed by the rising of moisture on a hot summer day. The second is formed by two air masses of different temperatures (for example, from over the Japan current and from the Arctic regions) coming together and forming an air-mass "front." Either type varies widely in its degree of turbulence and lightning does not necessarily result in every case. Even though lightning does not occur, the turbulent mixture of plus- and minus-charged snow, rain, or ice particles is present.

The warm-air type of thundercloud usually reaches its maximum early in the afternoon and begins to subside as the sun goes down. The cloud usually begins to spread out in the evening and the updrafts are less violent. When in this condition it is satisfactory for flying and will give good precipitation-static areas for test purposes. The static usually disappears before midnight.

The "front" type of clouds persist on through the night and will show static areas at practically any time. In winter, the "fronts" cover areas hundreds of miles long and are most troublesome from an airplane-radio standpoint. Summer thunderstorms may be avoided by flying around them but this is impractical for the long winter air-mass fronts. In mountainous country, air-mass movements are usually broken up into secondary turbulent areas over the crests of the mountain ranges. These turbulent areas also usually contain charged moisture particles.

Whenever the line of the air-mass front lies at right angles to the line of flight, the static area is traversed in a short time. When the front is parallel

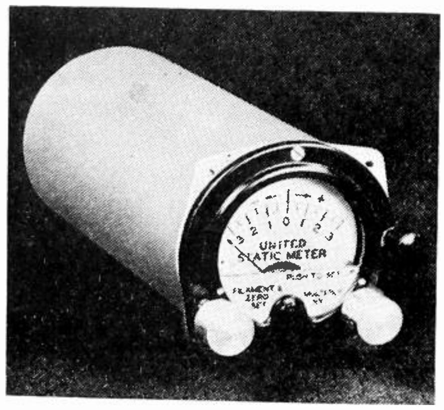


Fig. 5—Vacuum-tube electrometer mounted to fit standard aircraft instrument panel. A 9-inch by 3/16-inch metal rod projected from the plane into the air stream serves as a pickup.

to the line of flight the plane may be in the static area for several hours. This is the condition which constitutes the greatest hazard to radio navigation. Since the air-mass front is usually not perpendicular to the face of the earth and may lie in a horizontal plane, it is possible to avoid the static area by changing altitude.

Airline meteorologists are constantly cognizant of the location of all air-mass fronts and can, to a limited extent, advise pilots as to the probable location of static areas so that they may be avoided. In an effort to increase the reliability of meteorologists' predictions, three planes were equipped with vacuum-tube electrometers (Fig. 5) and flight experience correlated with weather information during the winter of 1937–1938. Results so far have not added very much to our present knowledge. They bear out Dean's[8] conclusion that "The southeasterly edge of low-pressure areas was found to be the most common source of disturbances." This may also be concluded from air-movement studies as explained in any modern meteorological text.[9]

In general, our meteorological work did not point directly toward a solution of the problem of reducing static but rather, furnished the necessary background which was essential for the work described in the sections which follow.

III. Antistatic Antennas

At the time the expedition was assembled every known type of antistatic antenna was installed on the plane. The object was to evaluate the relative merits of each and, if possible, to determine whether their operation could be improved. This investigation was

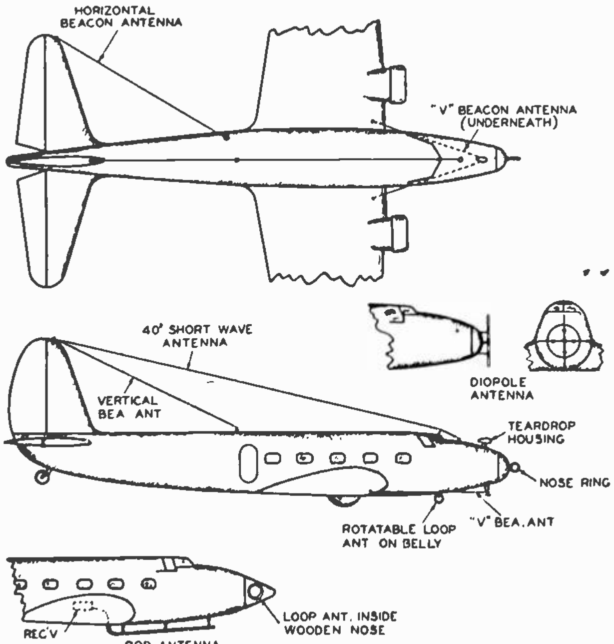


Fig. 6—Location of test antennas on laboratory plane.

carried on simultaneously with the meteorological and plane static-generation work.

In the paragraphs which follow, reference will be made to a number of antennas, shown in Fig. 6, which were tested during the flights. United States Department of Commerce radio beacons with a power of about one kilowatt in the 200- to 400-kilocycle band and airline ground telephone transmitters of 400 watts power in the 3000- to 6000-kilocycle band were used for comparison of reception. Fig. 7 shows the forward baggage pit equipped for antenna tests.

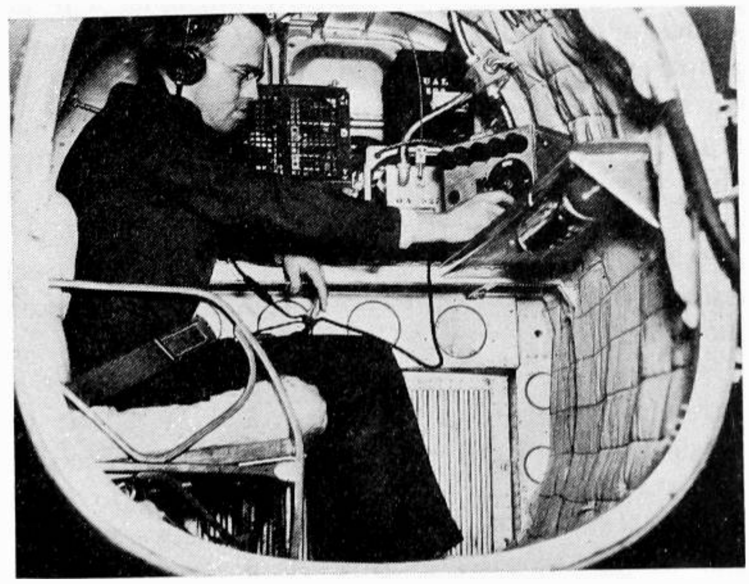


Fig. 7—Forward compartment for antenna comparisons. The switch on the wall above the engineer's hand allowed instant switching of three different loops to the receiver *OA*.

Antenna comparison tests indicate that precipitation-static interference is stronger at the rear than at the front of the plane. When static was of average strength, the loop located in the teardrop housing and the loop on the belly were both rotated and indicated that the source of maximum disturbance was toward the rear of the plane. When the disturbance became extreme, rotating the loops indicated static in all directions.

In mild precipitation static, when beacon reception on the V antenna was normal, the two rear beacon antennas were so noisy that no reception was possible. The vertical rear antenna had a 25-to-1 better signal pickup due to similar polarization of the radiation from the ground transmitter, but the static pickup was about the same on either. Both rear beacon antennas were about the same length and spacing from the fuselage. We concluded that precipitation-static-interference radiations did not have a definite polarization.

Although the 40-foot top antenna was far superior to the lower V antenna from a signal pickup standpoint, in precipitation static the V antenna would pick up 6000-kilocycle short-wave stations 1000 miles away when they were unreadable on the top antenna, thus further supporting the theory that the static was worse at the rear of the plane.

[8] S. W. Dean "Correlation of directional observations of atmospherics with weather phenomena," Proc. I.R.E., vol. 17, pp. 1185–1191; July, (1929).

[9] W. J. Humphries, "Physics of the Air," pp. 317–333. McGraw-Hill Book Company, New York, N.Y., (1929).

Early theories held that the grid of the first tube in a receiver was overloaded by static and was, therefore, blocked. To check this, microammeters were connected in receiver grid-return circuits and in series with all open-wire plane antennas. Grid-return meters would indicate only alternating voltages since the grids were inductively or capacitively coupled to the antennas. No grid-return currents were observed in severe static conditions. The antenna meters indicated currents of from a few microamperes to two milliamperes. Currents up to ten microamperes were also noted, however, when no static could be heard in the receivers. It appeared that the static heard in the receivers increased as this current increased but not always in direct proportion. Apparently the current had a large direct-current component with a smaller alternating-current modulation superimposed upon it.

Although a trailing wire was not tested as an antenna, it appeared that it would be about the noisiest form of antenna for reception in precipitation static. A trailing 150-foot steel cable carried as high as two milliamperes in vigorous "warm-front" conditions. The static leak connected across the input of the average aircraft receiver is about one-half megohm and with a two-milliampere peak current the voltage drop across the antenna input of the receiver could be 1000 volts. The noise modulation on this direct-current potential appeared to be less than one per cent, or only a few volts of random alternating voltage.

During the tests we reeled out the 150 feet of steel No. 14 B & S stranded aircraft cable in an effort to tune the entire plane to a frequency other than that used for receiving. It did not increase or decrease the static on the beacon frequencies. The short-wave receiver was however, tuned to a 60-meter wavelength; hence, the 150-foot cable plus the 65-foot plane length was more than one wavelength long. Reeling the cable in and out gave two antinodes of maximum static and two nodes of minimum static. The minimum, however, was not sufficiently low to aid reception materially.

At the time the tests were begun there were some precipitation-static theories which assumed that the noise was due to charged particles striking the antenna. To check this, a special pair of rod antennas were constructed. These were hollow, one-inch-diameter tubes; one of bakelite and the other of aluminum. A single-wire antenna was held in the center of these tubes by insulating disks and the lead-in completely enclosed in metal tubing. With this arrangement no particle of any kind could strike the antenna itself. The aluminum tube was grounded to the plane at three points with one-tenth-megohm resistors. The bakelite tube was painted with a solution of airplane dope and graphite so that its entire surface had a 10,000-ohm resistance to the plane. These antennas were somewhat similar in principle to those described by Dieckman[10] and Pickard[11] in 1912 and 1920. Good beacon reception was obtained on either antenna, but they gave no advantage over the regular No. 14 bare copper-wire antenna exposed to the snow and rain particles. We concluded that the impinging particles were not sources of noise, or that other noises were so great that the impinging particle noise was obscured.

The metallically covered loop antennas gave the following average advantage over the regular bare-wire beacon antennas during seven flights through precipitation static.

1. The advantage varied with the intensity of the static encountered.
2. In mild precipitation static the advantage as measured by the static output of a receiver in root-mean-square volts was 20 or 30 to 1.
3. In heavy precipitation static the advantage dropped to 5 or 10 to 1.
4. In extremely heavy precipitation static no beacon reception could be heard on any loop antenna even when flying within two or three miles of the beacon station.

On one test trip in a Pacific tropical-marine warm-air-mass front, no beacon reception was possible for a period of 25 minutes when any of the antistatic loops were used. Had the plane remained in this air-mass layer it would have been without beacon reception for several hours, since the front was parallel to the airway. With the assistance of the meteorologists such conditions normally can be avoided, and this particular flight represents an extreme case. It did appear, however, that the antistatic loop should be supplemented with some other system and meteorological guidance if a complete solution were to be obtained.

To determine whether the advantages of the metallically shielded loops lay in their metal covering, an experimental wooden nose was installed on the plane, and covered with copper foil. (Fig. 8.) The foil was cut at suitable points to make a Faraday shield. An unshielded loop in this nose gave practically the same results as the loops with the metal immediately surrounding the wire. A metallically shielded loop in the teardrop housing gave practically the same re-

[10] M. Dieckman, "Luftfahrt und Wissenschaft," part 1, (1912).

[11] G. W. Pickard, U. S. Patent No. 842910, February, 1907. See, "Static elimination by directional reception," Proc. I.R.E., vol. 8, pp. 358–395; October, (1920).

sults as the metallically shielded nose-ring loop or metallically shielded loop on the plane belly.

The nose-ring loop was usually about 5 per cent better than the loop on the plane belly; probably because it was farther forward. A ring-type low-impedance metallically shielded loop with an impedance-matching network gave the same results as a high-impedance loop of the same metallically covered construction.

The wooden nose without copper foil was painted with a mixture of dope and graphite so that it had an average resistance of 20,000 ohms to the plane structure. Signal pickup dropped about 15 per cent for loops inside this nose. No change in precipitation-static advantage occurred. An unshielded loop in this nose suffered from static, while a metallically covered loop in the same place gave the usual advantage. Position about the nose of the plane seemed to have very little bearing on the static effects.

These tests indicated clearly that the precipitation-static-reducing ability of the metallically shielded loop did not result from preventing the particles from striking the wire within it. It was apparent that the metal had the property of shielding against the static while letting the signal through and that copper was as effective as aluminum.

Previously pilots had reported that insulated surfaces such as windshield, de-icers, and nonmetallic loop housings charged up with respect to the plane. When the charge on them became high and the plane suddenly flew into a higher- or lower-charged cloud area, these insulated surfaces sparked over to the plane structure. Noise occurred in the receivers at each spark-over.

Painting the nonmetallic loop housing with dope and graphite stopped this source of noise by allowing the charges to leak off and equalize with the plane structure. If, however, the plane flew through an icing area, an ice cap formed on top of the graphite paint. This ice cap is an insulator which charges up and sparks over in the same manner as the insulated surface. Thus, in ice, the special paint did not accomplish its purpose. A partial solution was achieved by streamlining loop housings so that ice formation was reduced. Fig. 8 shows two pointed noses on housings for reducing this condition. Flight and ice-tunnel tests show that these pointed housings provide some reduction in dry-ice conditions but are not effective in wet ice.

During the course of the flights it was found that the bonding on one of the ring cowls had broken. This ring cowl, insulated from the plane by leather pads, charged up in precipitation and sparked over at regular intervals. In average charged clouds, this sparking caused a headphone noise sounding like pebbles falling in a metal pail. Any other exposed metal parts on a plane which are not bonded would cause a similar noise. The first steps toward improving plane reception in precipitation static should include a thorough inspection of all bonding.

Flight experience in dust storms was limited to only a few minutes and no useful tests could be completed. From pilots' reports and ground-station experience it appears that dust static acts in the same manner as snow or rain static and should be reducible by similar means.

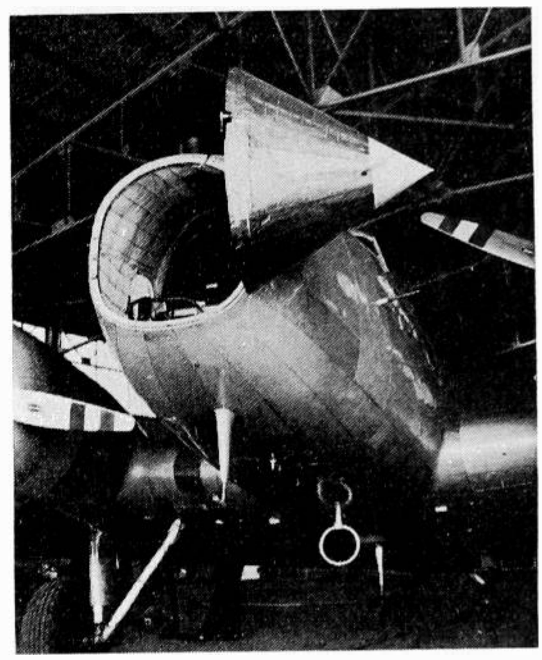


Fig. 8—Copper-covered wooden nose for testing loop antennas. Sharp point (white fabric) at end of nose used to reduce ice accumulation.

While the work on antistatic antennas did not provide a better solution for precipitation static, it did establish the advantages and some of the limitations of metallically shielded loops and provided substantiating evidence for the theories which are developed in section IV.

IV. The Generation of Static on the Plane

If a plane slowly climbs up through a cloud in which the charged droplets are uniformly distributed and in electrostatic equilibrium, no precipitation-static disturbance is heard in the radio. As it climbs, however, it must rise from an area of one charge to an area of different charge with respect to the earth. The voltage between earth and plane must, therefore, gradually increase as it climbs upward and decrease when it glides downward.

With a sharp two-foot steel point (Fig. 9) projecting from the rear of the plane, discharges of 10 to 15 microamperes or more have been measured while ascending or descending through charged fog particles. No static was heard in this condition. As long as the charging and discharging of the plane does not exceed a certain rate no static is heard. Since there must be a difference of potential between the plane

Fig. 9—Discharge point on tail of laboratory plane.

and surrounding atmosphere before discharge can take place, it is apparent that the plane must exceed a certain potential before static is heard.

In such a condition the 40-foot short-wave antenna on top of the plane gave no noticeable static in the short-wave receiver until a continous current flow of two microamperes was exceeded. The receiver had a one-half-megohm resistor across the antenna input through which this current flowed. The direct-voltage drop across this resistor was therefore one volt but the ripple superimposed on this direct voltage must have been only a few microvolts and of a random nature. Thus the noise in the receiver is caused by the random variation of the discharge flow rather than the total discharge.

Whenever the discharge from the steel point or antenna exceeded a certain rather broad value, the characteristic precipitation-static sounds were heard in the headphones.

A number of short metal rods were installed on the plane to learn the distribution of this discharge. These were arranged on the nose, tail, each wing, behind the exhaust outlet, behind the propeller, and at four points along the plane belly. The points were connected in a number of group arrangements to vacuum-tube electrometers which were in turn connected to paper recorders. Recordings were made in a variety of cloud formations over a period of eight weeks. Since, at a speed of 180 miles an hour the variation in charge on the plane was so rapid that it was impossible to make a record of it with the usual notebook and pencil, automatic recording devices were the only means by which a usable record could be obtained.

The grouping of these points which gave the most orderly results consisted of a pointed two-foot rod in the disturbed air at the tail, a pointed two-foot rod on the nose projecting into the undisturbed air ahead of the plane, and a plate on the nose to record the impacting moisture particles. Although conditions varied widely and the tests are lacking in completeness, a rough summarization of the conditions encountered is given in Fig. 10.

A study of data on all the points resulted in the following conclusions:

1. The plane may be either positive or negative with respect to the surrounding cloud.
2. At any instant one wing may be in positive cloud particles while the other is in negative.
3. At any instant the nose of the plane may be in positive particles while the tail is in negative or vice versa.

The maximum cross flow measured from wing to wing was about 500 microamperes though undoubtedly larger flows are possible. The maximum would occur with a stroke of lightning. There are records of

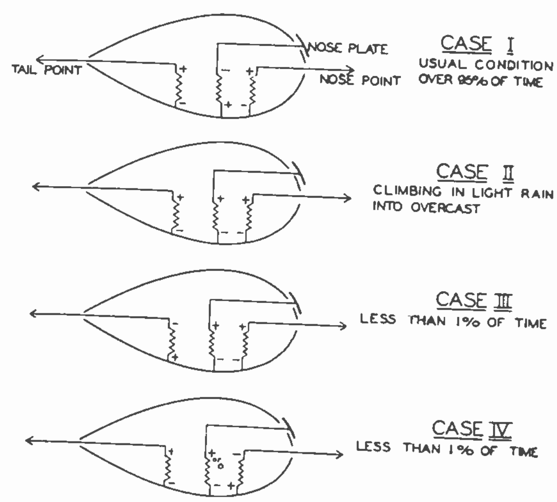


Fig. 10—Electrostatic-charge distribution on laboratory plane while in flight through a representative group of charged-cloud formations.

harmless lightning strikes on all-metal planes which indicate wing-to-wing flows of several thousand amperes.

During one of our flights we encountered a condition in a thundercloud in which the plane's magnetic compass moved 10 degrees with respect to the gyrocompass for a period of several minutes. This may have been due to a strong magnetic field in the cloud or to a cross flow of current in the plane structure. Ground tests with a storage battery connected from wing to wing required a current of about 45 amperes direct current to produce the same compass deviation. A nose-to-tail current of 125 amperes produced the same effect. This would vary with the position of the plane with respect to the earth's magnetic field.

It is known that a negatively charged point will go into corona about 50 per cent more readily than a positively charged point. It is also known that the action of the propeller in cutting up water particles at a tip speed of 800 feet per second will produce an electric charge. It is reasonable to believe that the wing of a plane moving at 260 feet per second will break up water particles and produce a charge. The electric-discharge recordings are, therefore, the result of at least seven factors:

1. The plus or minus charges of the water particles in the cloud which are collected by the wings.
2. The generation of charge due to the wing sections breaking up water particles.
3. The generation of charge due to the propeller breaking up water particles.

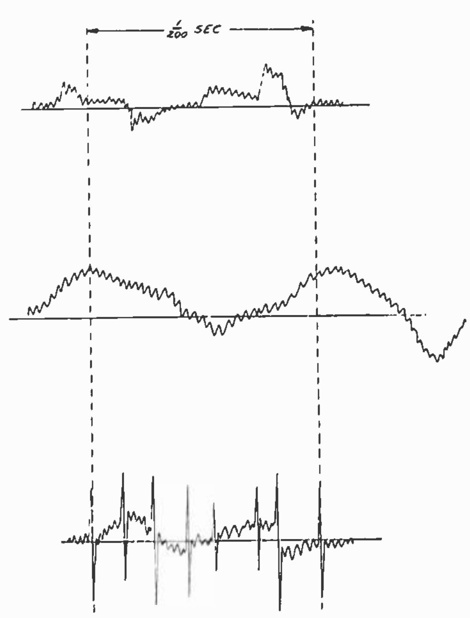


Fig. 11—Cathode-ray oscillograph patterns of voltage fluctuations across a 100,000-ohm resistor connected in series with the pointed steel rod of Fig. 9. These were taken through a three-stage resistance-coupled amplifier with an upper cutoff of 57,000 cycles.

4. Foreign matter in the water particles (Portland, Oregon, tap water split by the rotating propeller gave a positive charge while Cheyenne, Wyoming, tap water gave a negative charge).
5. The rectification action of the test points with different polarity of the plane charge.
6. Piezoelectric voltages due to squeezing ice and snow crystals by impact of wing or propeller.
7. Cross-current flows due to the plane short-circuiting sections of the cloud having different potentials.

From the above it is obvious that the mechanism by which the plane gathers an electrostatic charge is quite complex. Many test flights and much research

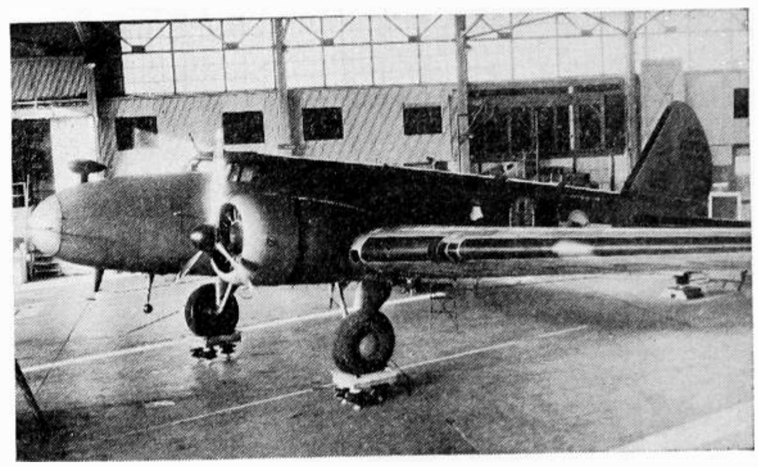


Fig. 12—Laboratory plane mounted on insulators during charging tests.

would be necessary to evaluate each of this group of variables. Such work did not appear justifiable since, in any case, it appeared probable that whether the plane became plus or minus it eventually reached a sufficiently high potential for corona discharges to appear on the wing tips or any sharp projecting points. As a check on this assumption, a cathode-ray oscillograph connected to any of the test points gave typical corona-discharge tracings whenever the characteristic sounds were heard in the radio. Typical tracings are shown in Fig. 11.

To test this theory further the plane was charged up by a small Wimshurst machine while standing on the ground and, by bringing a pointed ground wire near its structure, the characteristic precipitation-static sounds could be duplicated. Since this experiment was limited by the insulation of the rubber tires, the alternating-current modulation of the Wimshurst disk and the general variability of such a generator, a 100,000-volt direct-current X-ray power supply was assembled for additional tests. The plane was set up on high-voltage insulators in a large metal hangar (Fig. 12) and charged up to either plus or minus 100,000 volts.

With this arrangement, engineers could remain inside the all-metal plane with all radio equipment operating and use the test equipment in the same manner as was possible in flight. The tests further substantiated the corona-discharge theory. The voltage was sufficient to make antistatic loops and open-wire antennas inoperative in the same general ratio as had been observed on the test flights. The characteristic precipitation-static sounds were produced without the presence of precipitation.

The hissing noise in the receivers with regular antennas began as low as 30,000 volts, depending upon local humidity and the proximity of the artificial ground plane to the various points on the plane. The "crying" precipitation-static sounds usually began at about 55,000 volts and occurred more readily when the plane was positive with respect to ground. This crying phenomena was readily traced to a corona discharge from some point on the plane. Artificial points were set up for its study and we concluded that the ionized air around the point breaks down at an audio-frequency as well as a radio-frequency rate. This rate varied with the amount of moisture in the air, the shape of the point, and the voltage gradient at the point. Under controlled conditions it would produce any audio-frequency note. For example, on one test the noise was produced in the following order:

55,000 volts	frying noise (radio-frequency disturbance)
60,000 volts	frying noise begins to pulse at about 10 cycles per second
62,000 volts	frying noise pulsed at 100 cycles
64,000 volts	frying noise pulsed at 500 cycles
66,000 volts	frying noise pulsed at 2000 cycles
68,000 volts	frying noise pulsed at 8000 cycles
70,000 volts	frying noise pulsed at 15,000 cycles
72,000 volts	frying noise pulsed at above audibility

At about this voltage some other point on the plane began the same sounds and went on up through the musical scale.

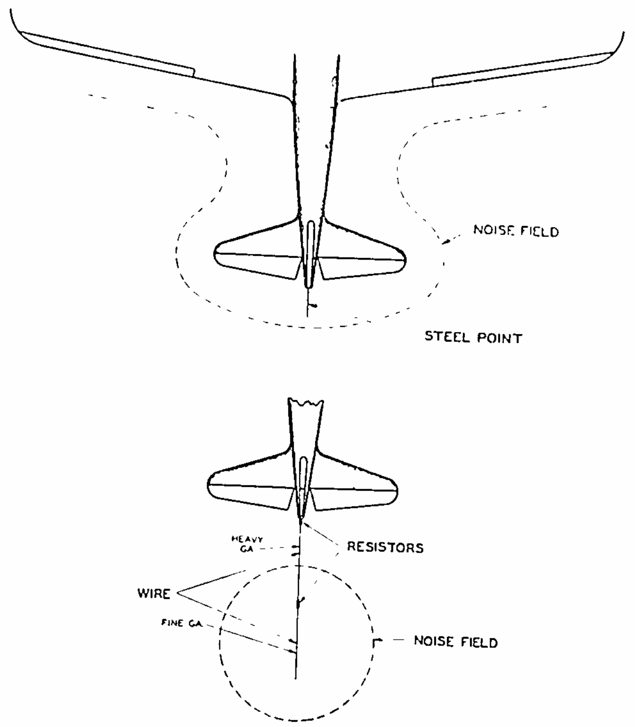


Fig. 13—Noise field patterns from corona discharge on laboratory plane showing effect of suppressor resistance and isolation of discharge points.

At any one time a number of points on the plane produced this musical corona in any order. This, then, was the cause of the characteristic precipitation-static sound.

A study of the plane structure indicated that antenna masts, rivet heads, cotter keys on aileron hinges and tail wheels, the antennas themselves, and any sharp points on the plane were the focal points of the corona discharges and consequently the source of precipitation-static interference while in flight.

Unless these discharge points are quieted, precipitation static could not be eliminated. Covering them with an insulator, reducing their sharpness, or covering them with a well-rounded corona shield would only allow the plane to build up to a still higher potential until some other point started corona.

There were two possible solutions; first, reduce the ability of the plane to gather or generate charges, and second, admit that the plane could not be prevented from gathering charges and work out a means for discharging it, which would not cause radio interference.

The second solution offered the best possibilities although means for accomplishing the first are within the bounds of reason. It is possible that a partial solution of both may be used eventually.

A study of the corona-discharge noise from test points indicated that it had a very short wavelength and that its attenuation with distance was rapid. The field pattern caused by a point in corona at the rear of the plane is shown in Fig. 13. When a resistor was added in series with the point, its interference-producing ability was materially reduced. Curves of resistance values versus receiver noise for constant current discharge, show 85 per cent noise reduction at 50,000 ohms, 92 per cent at 100,000 ohms, 98 per cent at 1/2 megohm, and 99 per cent at 1½ megohms. These values, of course, only apply to this special case but later tests on planes in flight gave similar results. Moving the corona point away from the plane took advantage of the rapid attenuation of the radiation and gave a better pattern. A single three-thousandths-inch-diameter wire which was, in effect, an infinite series of sharp points, gave a still better pattern as shown in Fig. 13.

At 100,000 volts, a discharge of one milliampere from a fine wire, isolated 50 feet from the plane via a large-diameter wire, gave no disturbance in the radio if a 100,000-ohm resistor was connected in series. A fine wire without resistors two feet from the plane prevented all radio reception with one fortieth as much current.

This indicated that a fine-wire discharger with suitable suppressor resistor, located as far from the plane as possible, might be a solution to our problem.

Reeling in a seven-thousandths-inch-diameter wire while in flight is almost impossible mechanically. In order to avoid this, an arrangement of 17 short, fine wires and one-watt metalized resistors was flight-tested first. It had the mechanical advantage that the wires were so short that they did not drag on the ground on take-off, and reeling them in was, therefore, not necessary. This system proved unsuccessful in flight although its failure was later found to be the result of improperly constructed suppressors. When the 17-wire system failed, a single long wire and reel arrangement was constructed. After much difficulty with wire breakages, a single wire finally held on the plane long enough to complete one test flight.

This test proved successful. At fifty miles from the Rock Springs radio beacon, the signal was completely obscurred by precipitation static. When the fine wire and suppressor were reeled out, the static dropped to almost zero and the beacon could be heard clearly (see Fig. 14). The wire was reeled in and out several times before it broke and its static-reducing properties held true each time. The corona-discharge theory had been proved in flight and a cure had been attained but a method for using it in practical daily operations remained to be worked out.

V. Corona-Discharge Radiation

The remainder of 1937 and the spring of 1938 were devoted to the development of a practical fine wire and suppressor system. In addition, a research into the nature of the corona radiation was carried out, in order to determine theoretically the distance at which the discharge wires should be trailed behind the plane. A study of the effectiveness of shielded-loop antennas was carried on as part of the radiation research in an effort to improve their antistatic properties, if possible.

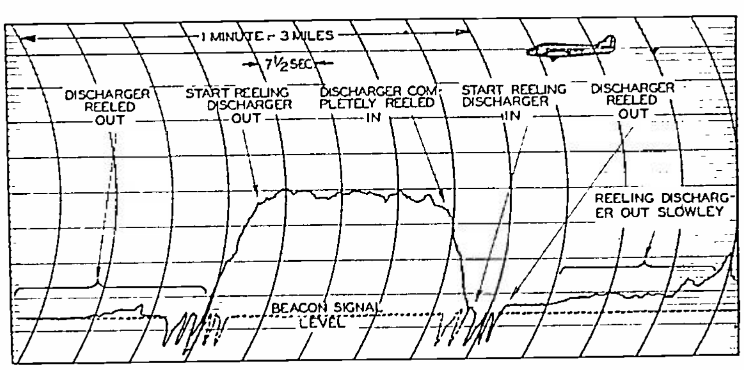


Fig. 14—Effect on reeling out experimental discharge wire and suppressor on reception of Rock Springs, Wyoming, radio beacon previously obscured by precipitation static. Beacon-keying characteristic is not recorded when wire is reeled in.

This work was carried on with a synthetic precipitation-static generator especially developed as a reference noise source. The generator consisted of a conventional 15,000-volt direct-current rectifier operating from 110-volt, 60-cycle power. The negative terminal was connected to a steel phonograph needle through a 10-megohm resistor and the positive terminal was connected to a piece of fine copper screen about as large as a silver dollar. The entire assembly was mounted in a shielded box to prevent stray radiations. When the phonograph needle is brought within one-half inch of the screen, a vigorous corona discharge begins. The radiation from this discharge as picked up by a radio receiver sounds exactly like the precipitation static heard in flight. The radiation field was increased or decreased by attaching an antenna to the negative (needle) terminal of the gap. A 42-inch vertical brass rod with an 8-inch diameter "burr" at its end, made by cutting up strips of copper into a multitude of sharp points, gave a reliable radiation field.

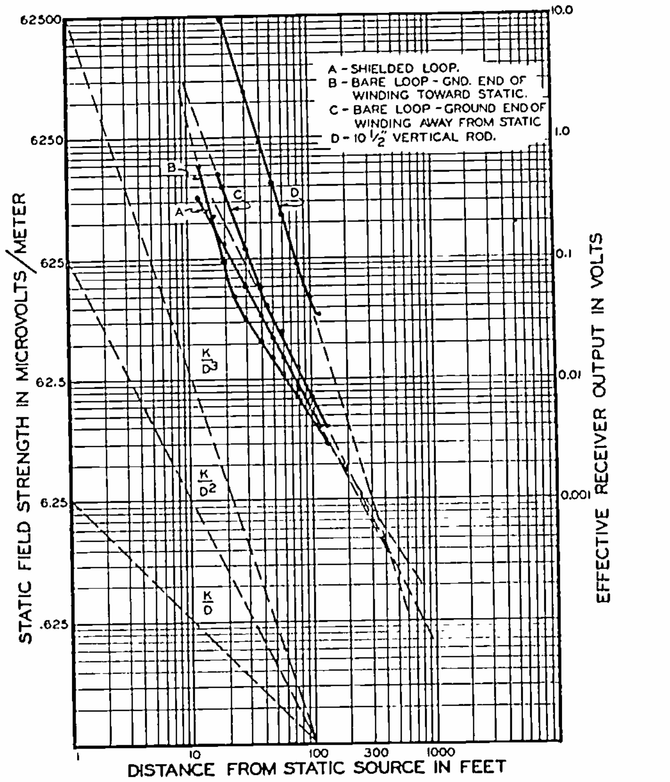


Fig. 15—Attenuation of electromagnetic radiation from corona-static generator with different types of receiving antennas. The three dashed lines are typical slopes for reference purposes.

A movable receiving equipment in a completely shielded metal box tuned to the 200- to 400-kilocycle band was mounted on a small wagon. An arm on this box was arranged for mounting various types of antennas for test. Suitable means for comparing field strengths were worked out. The static generator was located out in the open to prevent reflections from buildings and the receiving system rolled toward and away from it to measure the contours of the radiation fields emitted from it. All antennas were adjusted to have identical pickup from a 350-kilocycle radio beacon before data were taken. Four curves on the following types of antennas are shown in Fig. 15:

1. A $10\frac{1}{2}$-inch vertical rod.
2. A 12-inch mean diameter metallically shielded low-impedance loop having one side of the winding grounded and the other connected to the grid of the first receiver tube.
3. The same loop with the shielding removed and the grounded edge of the winding toward the generator.
4. The same loop with the shielding removed and the grounded edge of the winding away from the generator.

Many other curves were drawn to check apparatus reliability, consistency of results, high-impedance loops, position of the insulated gap in the shielding and the effect of isolated shields. The group on Fig. 15 serves as a typical example and may be interpreted in the following manner:

1. The radiation from the synthetic precipitation-static generator consists of two components:
   a. The usual electromagnetic radio wave (commonly called the "radiation field") which attenuates inversely as the distance. The energy in this component is quite small and its effects on the curves can be neglected.
   b. The nonradio "induction field" immediately surrounding the radiator, which is usually considered as returning to its source during each half cycle.
2. That the measurable effects of the "induction field" of this generator do not extend more than several hundred feet from it and that this "induction field" consists of two components:
   a. An electrostatic field of high intensity.
   b. An electromagnetic field of lesser intensity.
3. The attenuation of the electrostatic and electromagnetic components of the "induction field" is normally as the inverse square of the distance. In this case, however, the ratio of the intensity of the two components varies with the construction of the generator radiator and also with the intensity of the corona discharge from the sharp points of the "burr."
4. A rod-type open antenna picks up a large amount of the electrostatic component and a small amount of the electromagnetic component. Its curve should fall off as the inverse square of the distance but, because of the proximity of the ground plane, falls off as about the inverse cube of the distance.
5. The completely shielded loop picks up practically none of the electrostatic component due to its shielding. Its curve represents only the electromagnetic component which falls off as the inverse square of the distance. The ground plane is nonmagnetic and does not affect this component as it does the electrostatic component.
6. The unshielded unbalanced-winding loops pick up all of the electromagnetic component but only part of the electrostatic component, depending upon the end that is turned toward the generator. The grounded-turn end of the winding partially shields the turns on one side with the resultant unsymmetrical pattern. The "knees" in curves *B* and *C* represent the point where the inverse-cube electrostatic component crosses the inverse-square electromagnetic component for the particular loop. They do not, of course, represent the relative strengths of the two components in space, since each type of antenna "sees" the induction field in a different proportion. The upper half of curve *B* represents the summation of both components and falls off at some power greater than the inverse cube until the knee is reached, after which the electromagnetic predominates.
7. An unshielded balanced loop cancels out practically all of the electrostatic component and picks up all of the electromagnetic component. Its curve is identical with the completely shielded loop, curve *A*.

While these curves were an interesting demonstration of the radiation phenomena, they gave us no clue as to how the operation of the antistatic loops might be improved. Instead they established the limitations of the device.

As long as the corona radiation from sharp points on the plane is largely electrostatic in nature, the loops have an advantage over an open-wire antenna. Whenever the corona becomes severe and ceases to be a soft, blue glow, it breaks up into a number of irregular fingers. These fingers increase the electromagnetic component and the loop shielding is ineffective against it.

The curves also show that the ratio of advantage of the loop over the ordinary antenna changes with the distance of the antennas from the noise source and the intensity of the noise source. This accounts for the many conflicting results observed by pilots and investigators of the phenomena.

Curves have been drawn with the loops and antennas actually mounted on the airplane. In this case the synthetic generator was placed on rollers and moved with respect to the plane. Results correlate with the other curves except for reflection troubles due to the large metal wing or fuselage areas. This work also substantiated the desirability of removing the discharge points at a distance from the plane. When the static generator was connected to the plane structure, instead of a few inches from it, the noise increased as much as five to one.

During the previous work with the plane charged to 100,000 volts a pair of horizontal and vertical dipole antennas were tested. They were tuned and coupled to the receiver by means of a balanced line and an electrostatically shielded antenna transformer. Although they gave a definite improvement over corona static as compared to a single bare wire, their signal pickup at 200 to 400 kilocycles was too poor for practical aircraft use. The test results, however, parallel the results with the unshielded balanced loop and indicate the value of balanced antennas in reducing the electrostatic component of corona radiation. An electrostatically shielded input transformer alone, connected to an unbalanced antenna, gave no improvement over unshielded coupling methods.

The radiation studies did not disclose any new means for improving the present types of metallically shielded antistatic loop antennas. Precise electrostatic balancing of the loop windings with respect to the metallic covering might give a small improvement. Such balancing is difficult to maintain in commercial practice and the advantage gained does not seem to justify the maintenance costs.

The improved knowledge of corona-radiation attenuation was of value in understanding the proper length of suppressor resistors covered in section VI.

VI. A Practical Discharge System

A successful trailing-wire discharge system is predicated on the ability of the fine wire to go into corona before other sharp points on the plane. It must also have sufficient area to dissipate continuously the charge picked up and generated by the plane. This dissipation must take the form of a "soft" corona so that no strong electrostatic or electromagnetic disturbances are created. The fine wire must be sufficiently far from the plane to allow good attentuation of a certain amount of unavoidable disturbances before they reach the plane antennas.

The diameters of the sharpest points of the average transport-plane structures are as follows:

1. Trailing edge of propeller 0.047 inch minimum.
2. Antenna wires 0.063 inch minimum.
3. Sheet-metal edges on tail structures 0.057 inch.
4. Miscellaneous cotter pins or sharp metal edges 0.005 to 0.030 inch.

The discharge wires must be smaller in diameter than these items so that the discharge will occur on them rather than on the plane. The suppressor resistor diameter must be greater than these items so that no discharge will take place on its surface. Sharp points such as cotter keys must be rounded off and sharp edges of sheet metal smoothed over so discharges will take place only on the fine wire. It is particularly important to reduce sharp points on antenna structures and near-by surfaces.

During the Oakland tests when the plane was charged to 100,000 volts, the V beacon antenna was replaced with wires ranging from 0.003 inch in diameter to tubing one inch in diameter. As the diameter increased, reception improved and outgoing corona current decreased. Receiver noise decreased in proportion. The data indicated, however, that practically no improvement could be made by increasing the present No. 14 B & S gauge wires to No. 10. Wires larger than No. 10 are usually impractical on aircraft because of weight and drag considerations.

The search for suitable discharge wires produced a seven-strand steel aircraft cable of 0.031-inch diameter which seemed mechanically strong enough to withstand whipping in flight. This diameter was roughly one half the propeller-edge and antenna-wire diameters and the wire should therefore go into corona at about one fourth the voltage. One-eighth-inch-diameter rope was impregnated with graphite and given a rubber coating for weathering protection. By proper control of impregnation the rubber-covered rope could be produced with a resistance value of 500 ohms to several megohms per foot. The combination of wire and rope provided a flexible suppressor and discharger which would withstand flight conditions.

The Boeing laboratory plane was equipped with one-half-inch metal tubes running from the cabin to each wing tip, each stabilizer tip, and the tail. Reels of fine wires and rope suppressors were provided so that they could be released and retracted via the tubing from any of the five points. Flights with this arrangement were made in the fall of 1937 to determine the length of wire, length of suppressor, location of discharger, and number of dischargers required to provide relief from precipitation static.

A study of the data indicated that a fine-wire length of five feet, suppressor length of five feet, and resistance of 100,000 ohms were sufficient for practical purposes. Dischargers located at either stabilizer tip or the tail gave relief while those on the wing tips were not effective. The wing-tip locations are closer and parallel the plane antennas.

Flight tests were then repeated on a Douglas DC-3 plane whose over-all size is approximately 50 per cent greater than the Boeing. The 5-foot by 5-foot by 100,000-ohm discharger also proved sufficient for this plane. Calculation and estimates based on experimental data indicate that dischargers on planes of this size must dissipate from five to fifty watts of energy during vigorous storm conditions. It seems reasonable to expect that still larger aircraft must have several dischargers or a larger single discharger to provide relief.

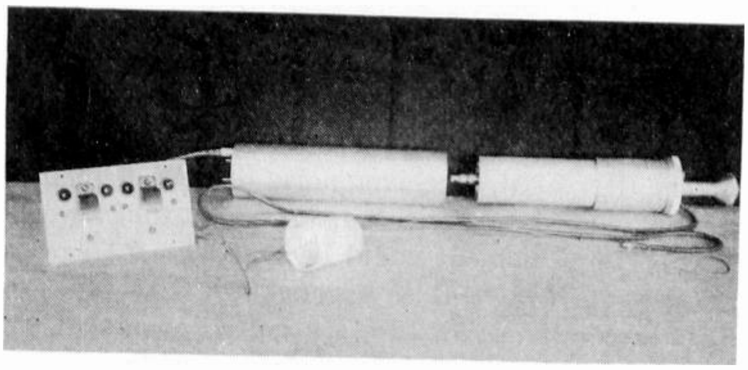


Fig. 16—Plane discharge-cartridge assembly. From left to right are shown the pilot's control panel for two cartridges, the socket for mounting on rear of the plane, and the cartridge with its plunger extended. The black flexible resistor lies in front of the socket and cartridge as does the fine wire which terminates at the white-paper wind cone in the foreground.

Having determined the discharger size, the next step required consideration of a practical mechanical means for using it. Three methods were tested:

1. Permanent attachment of the discharger to the plane tail so that it could drag until replacement was necessary.
2. A remotely controlled reel was installed in the plane tail so that the pilot could roll it out and retract it when in precipitation static.
3. A discharger stored in a small container which could be ejected when necessary. The container was made plug-in so that it might be quickly replaced at the end of the flight and cheaply refilled at leisure.

System 1, failed due to whipping, which usually destroyed the discharger on take-off or in flight. Systems 2 and 3 permit the use of a small cheap paper wind cone at the end of the fine wire and thus stop the whipping. Systems 2 and 3 were both successful but 3 weighed about one half as much, cost less, and allowed quicker replacement, hence was adopted.

Fig. 16 shows the static-discharge cartridge with accessories. Usually two sockets and cartridges are installed for safety purposes. The cartridge is spring-loaded and contains the wind cone (a standard paper cup with bottom removed), suppressor rope, and fine wire. The spring is released electrically from the panel in the pilot's cab and ejects the wind cone and discharger through the weatherproof end of the cartridge. Test buttons with indicator lights are provided on the pilot's panel so that he may check the circuits before departure and also flash back when the discharger has been successfully ejected. A 0.016-inch-diameter stainless-steel wire has been manufactured for the commercial models, thus providing a 9-to-1 discharge factor over the propeller's sharpest edge.

Flight-test data indicate that the static-discharge-cartridge system will give marked relief from precipitation static on both the 200- to 400-kilocycle and 2000- to 6000-kilocycle aircraft frequencies. Some extreme conditions were found in which relief was not complete in either band and additional relief was obtained by using a metallically shielded loop for the beacon frequencies. In conditions where the loop alone would normally be inoperative for as long as 30 minutes, the combination of discharger and loop reduced the inoperative period to approximately two minutes. Comparative data of this kind are, of course, subject to wide variation, but two years of test flying, through the worst conditions that could be found, indicates that for all practical purposes, beacon reception should always be possible except for a two- or three-minute period when the crest of the precipitation-static area is traversed.

Metallically shielded loops for the 2000- to 6000-kilocycle band are not yet developed and the two-way voice communication may therefore be inoperative for a longer period than the beacon channel. This condition, while undesirable, is not serious from a safety standpoint.

V. Ground-Station and Marine Static Reduction

An understanding of precipitation static on aircraft readily points to the application of similiar principles to ground-station reception. The "crying" sounds heard on aircraft have often been heard at ground stations in snow storms, thunderstorms, and dust storms. Stations at high altitudes such as Cheyenne, Wyoming, seem to suffer from this condition more often than stations at low altitudes. There are, however, many reports from low-altitude stations such as Cleveland, Ohio, and Dallas, Texas. Dust storms are often the cause in the Mississippi Valley.

Nakai[12] explains the ground-station phenomena on the basis of the charge on each raindrop. The charge per drop is assumed and effect of a number of drops striking the antenna summated to account for a

[12] Tomozo Nakai, "Some notes on rain static in Japan," Proc. I.R.E., vol. 25, pp. 1375–1380; November, (1937).

current of 20 microamperes measured in the antenna lead-in. The effective interference is given as equivalent to a field strength of about 5000 microvolts per meter. He suggests that each drop striking the antenna sets up a short damped oscillation as it gives up its charge. The summation of these individual oscillations produces a hissing or frying background of interference.

The creation of charged raindrops has previously been explained by Simpson in his work on thunderstorms. A static machine using water drops was devised by Kelvin in 1867. The rain-static phenomena described by Nakai therefore follows well-known electrostatic principles. Our tests on aircraft indicate that the 20 microamperes probably consisted of a large direct-current component with a superimposed alternating-current component equivalent to 5000 microvolts of noise.

In contrast with these conditions, the writer and others have often observed ground-station static of a severe "crying" type when no particles of moisture or dust were striking the antenna. In one case the beginning of rainfall stopped the interference instead of starting it. De Groot[13] describes similar effects observed in the Dutch East Indies in 1916. It is thus apparent that the charged-raindrop theory does not hold for all cases and some other explanation is necessary.

The presence of the characteristic musical "crying" readily points to the same oscillating corona from sharp points as occurred on aircraft.

Fig. 17 illustrates a possible explanation. During thunderstorms, a high potential builds up between clouds and earth. When it becomes high enough, lightning occurs and the tension is temporarily relieved. After the lightning flash, the building-up process repeats until another breakdown point is reached.

In between recurring flashes, buildings, trees and radio antennas are alternately subjected to a high electrostatic stress since they are directly between the clouds and earth. Observations show that a blue corona (St. Elmo's fire) occurs from sharp points on such projections during these periods. A three-inch diameter soft, blue corona has been observed on a receiving antenna wire in daylight during such conditions. Instantly after a near-by lightning flash, the blue glow disappears, only to build up again until the next flash. The increase in interference in a radio receiver follows the increase and decrease in visible corona on the antenna.

The same effects have been observed on aircraft in flight. A blue corona builds up on the propeller tips and antenna wires until a near-by lightning flash occurs. Instantly thereafter the interference ceases, the glow dissappears and the radio-beacon signals are heard again for a few seconds until the cycle is repeated.

Lightning studies show that most discharges are of a multiple nature caused by energy in clouds discharging horizontally from one cloud to another as the main stroke goes to earth at one point. Thus a lightning stroke occurring a number of miles away may relieve the tension and stop the corona from an antenna at a local point. It is not necessary that visible clouds exist over the local station since the electrostatic stress seems to extend for some miles from the center of disturbance.

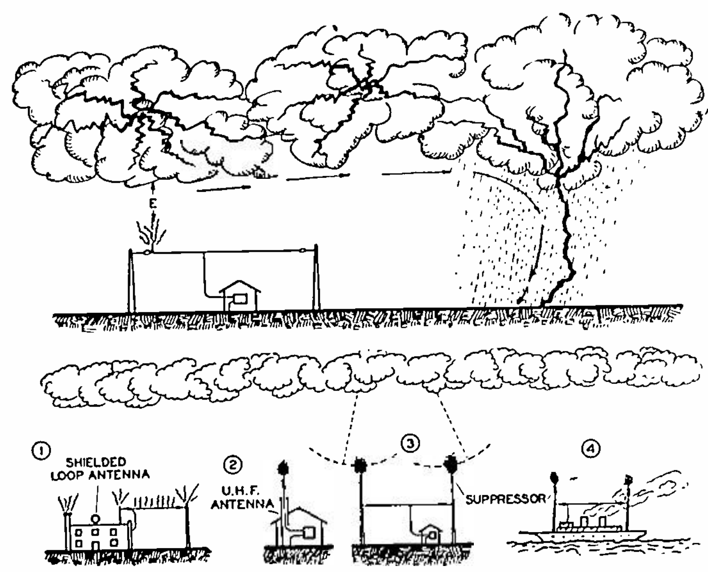


Fig. 17—Corona-static production and suggested means for improving reception.

If the corona on the receiving station antenna is a mild, blue glow, reception may not be seriously disturbed. If the antenna has sharp edges on its structure the corona may concentrate at such points and produce the characteristic "crying" sounds. Several methods for improving reception are possible as follows:

1. Remove all sharp edges from the antenna and immediately adjacent structures. Also increase the diameter of the antenna wire.
2. Install the antenna below the highest surrounding structures so that lines of stress will terminate on them instead of the antenna.
3. Install suppressor resistors at the highest points on the antenna and attach some sharp discharge points to and above the suppressors.
4. Since discharges from near-by structures may reach the antenna by radiation, it may be necessary to install suppressors and sharp points on them also.

[13] Cornelius J. de Groot, "On the nature and elimination of strays," Proc. I.R.E., vol. 5, pp. 75–120; April, (1917).

Suppressors and dischargers installed on top of ultra-high-frequency vertical receiving antennas have given marked improvement in actual tests. Similiar systems installed on marine vessels should result in material improvement of reception in a large number of tropical static conditions. Metallically shielded loops should give added relief.

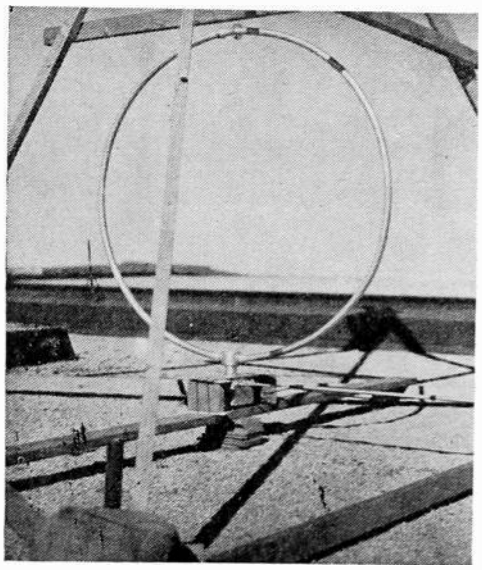


Fig. 18—Six-megacycle shielded-loop antenna for static reduction at ground stations.

Much additional work must be done before commercial forms of ground-station suppressors can be constructed. Resistors from one tenth to one megohm, having a low distributed capacitance, should prove satisfactory. Tests of a ground-station metallically shielded loop are in progress (see Fig. 18).

A study of the raindrop-charging process is being made at Purdue University along with studies of other types of aircraft dischargers. Work done on loop phenomena and discharge-point shapes at Oregon State College will be published in the near future. A study of the plane-charging process is being made at Reed College. Funds have recently been provided by the Civil Aeronautics Authority for extending this work and it is expected that some of the phenomena which are only briefly discussed in this paper eventually will be more thoroughly treated.

Acknowledgment

Many of the theories and practices discussed in the preceding pages were contributed by members of the flight party. The fine-wire and suppressor system of plane discharge resulted from a suggestion of R. H. George of Purdue University. Many of the test procedures were suggested by Drs. Marcus O'Day and A. A. Knowlton of Reed College, who assisted with the plane-charging analysis. Professor E. C. Starr assisted materially with the high-voltage aspects of the problem and has continued his investigations at Oregon State College. The Bendix Radio Corporation and Bell Telephone Laboratories loaned test equipment and personnel.

H. W. DeWeese, N. E. Klein, and Cullen Moore of the United Air Lines laboratory staff carried out the bulk of the equipment construction. Captain A. C. Ball located the static areas and piloted the plane through them.

Space does not permit mention of many others who made important contributions and for whose assistance and advice the writer is greatly indebted.

The Application of Negative Feedback to Frequency-Modulation Systems*

J. G. CHAFFEE†, ASSOCIATE MEMBER, I.R.E.

Summary—Negative feedback can be applied to a frequency-modulation receiver of superheterodyne type by causing a portion of the output voltage to frequency-modulate the local oscillator in such phase as to reduce the output signal. As a consequence of this arrangement the effective frequency modulation of the intermediate wave is diminished by the feedback factor. This reduction is accompanied by a decrease in noise and distortion. Restoration of the original signal level by increasing the degree of modulation at the transmitter brings about a corresponding increase in signal-to-noise ratio provided the disturbance is not too great, while distortion ratios are improved to about the same extent. These effects are treated analytically for the case where the disturbance level is sufficiently low to permit simplifying assumptions to be made. The results are in general agreement with observations made on an experimental laboratory system.

Comparing the feedback system with a frequency-modulation system using amplitude limitation, the ratio of signal level to noise level in the absence of modulation is identical in two systems. During modulation the noise level increases in the feedback system by an amount depending upon the ratio of the effective frequency shift of the intermediate-frequency wave to the signal band width. By keeping this ratio small, the increase in noise during modulation can be made relatively unimportant.

In cases where the disturbance level is high, phenomena have been observed which are very similar to those encountered when amplitude limitation is used.

Introduction

THIS paper describes a method for improving the performance of receivers designed to receive frequency-modulated waves. In its broader aspects this method can be described as the application of the principle of negative feedback to a superheterodyne frequency-modulation receiver. In its details the application of the feedback principle necessitates the use of a rather unusual circuit arrangement. This circuit differs from that of the simple feedback system in that the voltages fed back are not of the same frequency as those applied to the input of the receiver, and are caused to influence the response of the system by modifying the performance of the modulator.

In the ordinary feedback amplifier a part of the output voltage is carried back to the input and there combined with the applied voltage. The result is to modify the output and if the gain of the system is thereby reduced the feedback is said to be negative. The many advantages which result from negative feedback have been described by Black[1] and are coming to be more generally appreciated. The present paper deals with a method for adapting this principle to a frequency-modulation receiver and will show an example of its application to an experimental system in the laboratory.

General Discussion

Method of Applying Feedback

Consider a frequency-modulation receiver in which the incoming wave is combined with the output of a local oscillator in a modulator to produce a wave of intermediate frequency. This is then amplified, converted into an amplitude-modulated wave, and finally detected. The frequency of the intermediate wave is equal to the instantaneous difference in the frequencies of the incoming carrier and the local oscillator. So long as the frequency of this oscillator remains fixed the intermediate wave will be frequency-modulated in exact correspondence with the incoming wave. Suppose now that the local oscillator is frequency-modulated from a source of the same frequency and phase as that applied to the transmitter. As the index of modulation at the local oscillator is increased from zero the extent to which the intermediate wave is modulated will diminish since its instantaneous frequency is equal to the difference in the frequencies of the two sources. It then follows that if these two devices are modulated to the same extent the difference frequency will become constant and the output of the system will be zero. Finally a further increase in modulation of the local oscillator will cause the intermediate wave to be modulated with a 180-degree phase reversal.

This process can be readily analyzed as follows: Assume the oscillator at the transmitter to have been frequency-modulated by the signal wave

$$e = E_1 \cos pt. \tag{1}$$

The voltage delivered to the modulator by the incoming wave will be

$$A \cos (\omega_1 t + x_1 \sin pt + \phi_1) \tag{2}$$

where $x_1 = \Delta\omega_1 \div p = \rho E_1 \div p$, and $\Delta\omega_1$ is 2π times the maximum frequency shift. The local oscillation impresses the wave

$$B \cos (\omega_2 t + x_2 \sin pt + \phi_2) \tag{3}$$

where

$$x_2 = \Delta\omega_2 \div p.$$

* Decimal classification: R361.2×R261. Original manuscript received by the Institute, November 10, 1938.

† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] H. S. Black, "Stabilized feedback amplifiers," *Elec. Eng.*, vol. 53, pp. 114–120; January, (1934).

Application of these waves to a square-law modulator will yield a difference frequency wave proportional to

$$AB\cos\left[(\omega_1-\omega_2)t+(x_1-x_2)\sin pt+\phi_1-\phi_2\right] \quad (4)$$

for the case where $\omega_1>\omega_2$, or when the reverse is true

$$AB\cos\left[(\omega_2-\omega_1)t-(x_1-x_2)\sin pt+\phi_2-\varphi_1\right]. \quad (5)$$

In either case the resultant modulation index of the intermediate wave is the numerical difference of the original indexes, the difference in sign in the two cases signifying that the detected outputs will be of opposite phase. If $x_1=x_2$ the modulation is reduced to zero, and if $x_2>x_1$ modulation reappears with a phase reversal. It might be noted that if x_2 were originally made negative, thus causing the two oscillators to be frequency-modulated in opposite phase, the apparent modulation of the incoming wave could be increased indefinitely.

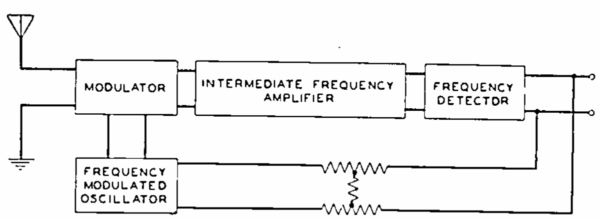


Fig. 1—Basic feedback circuit.

Suppose, now, that instead of frequency-modulating the local oscillator from an independent source, the equivalent is accomplished in a practical way. For this purpose a voltage from the output of the receiver is impressed upon the local oscillator as shown in Fig. 1. The transmitted wave will then have a modulation index $x_1=\rho_1E_1\div p$, while the local oscillator, being acted upon by a portion of the output voltage E_0, will have an index $x_2=k\rho_2E_0\div p$. If the frequency detector[2] is assumed to be linear the amplitude of the detected output will be proportional to the product of the amplitudes A and B of the incoming and local oscillator waves, the resultant index of the intermediate wave, and the slope factor a_1. Thus we can write the output voltage amplitude

$$\begin{aligned} E_0 &= \alpha a_1AB(x_1-x_2)p \\ &= \alpha a_1AB(\rho_1E_1-k\rho_2E_0). \end{aligned} \quad (6)$$

Therefore,

$$E_0=\frac{\alpha a_1AB\rho_1E_1}{1+\alpha k a_1AB\rho_2}\cdot \quad (7)$$

Setting $\alpha a_1AB=\mu$ and $k\rho_2=-\beta$ we obtain the familiar form encountered in the analysis of feedback amplifiers

$$E_0=\frac{\mu(\rho_1E_1)}{1-\mu\beta}\cdot \quad (8)$$

Without feedback the output of the system is merely $\mu(\rho_1E_1)$. The feedback factor $1+\alpha a_1kAB\rho_2$ is a measure of the extent to which the over-all gain of the system has been modified by feedback. If this factor is greater than unity the feedback is negative, while if k is made negative by reversing the feedback connections the effect is regenerative, and instability is encountered when the factor becomes zero.

It will be noted that when $\alpha a_1kAB\rho_2\gg1$, (7) becomes

$$E_0=\frac{\rho_1E_1}{k\rho_2}\cdot \quad (9)$$

Thus for large amounts of feedback, the output signal becomes independent of such factors as fading of the incoming wave, variations in the local oscillator voltage, or changes in detector efficiency. Hence automatic gain control is secured. This feature is equivalent to that found in ordinary feedback amplifiers in that for large amounts of feedback the over-all gain becomes independent of variations in the performance of the amplifier proper.

Reduction of Noise

The application of negative feedback in the manner described brings about a reduction in signal level by decreasing the effective modulation of the received wave. It then becomes possible to increase the modulation level at the transmitter to a corresponding degree and thus to restore the output signal to its former value. This process is made possible through the use of frequency rather than amplitude modulation since the permissible degree of modulation is then determined by the receiver characteristics. It will be shown that feedback also reduces the noise level at the output of the receiver, provided that the disturbance is not too great. Thus when the modulation level is raised to offset the effect of feedback an improvement in signal-to-noise ratio is realized.

The mechanism by which noise is reduced can be described qualitatively as follows: Noise at the output terminals of the receiver is caused to frequency-modulate the intermediate wave in such fashion as to produce, upon detection, a component which tends to cancel that which would exist in the absence of feedback. An analysis of this process for the case where the carrier is large compared with the disturb-

[2] The term frequency detector is used in this paper to designate the combination of conversion circuit and amplitude detector. A more extended discussion of modulation and detection is given in Appendix A.

ance responsible for the noise is developed[3] in Appendix B. It is assumed that the disturbance can be represented by a continuous spectrum of sinusoidal voltages of equal amplitude but phased at random. Impressed along with the disturbance is the signaling wave (2). Then if N^2 is the mean disturbing power per unit of band width in the vicinity of the carrier frequency, and r_1 is the resistance of the input circuit, it is shown that the output noise power is[4]

$$P_N = \frac{2N^2 r_1}{F^2}\left[a_0^2 + \frac{a_1^2\Delta\omega^2}{2F^2} + \frac{a_1^2 q_a^2}{3}\right] q_a \quad (10)$$

where a_0 and a_1 are, respectively, the gain and slope factor of the intermediate amplifier and conversion system as defined by (47), and q_a represents the upper limit of frequency response of the output circuit, or the upper limit of audibility as the case may be. F is the feedback factor $(1-\mu\beta)$. The corresponding signal power is

$$P_s = \frac{A^2 a_1^2 \Delta\omega^2}{2F^2}\cdot \quad (11)$$

The reduction in signal level occasioned by feedback can be offset by increasing the frequency shift of the transmitted wave. If it is increased so as to have the value $\Delta\Omega = F\Delta\omega$ then the shift of the intermediate-frequency wave will be restored to its original value $\Delta\omega$ and the signal level will remain unchanged. Then the noise power becomes

$$P_N = \frac{2N^2 r_1}{F^2}\left[a_0^2 + \frac{a_1^2\Delta\Omega^2}{2F^2} + \frac{a_1^2 q_a^2}{3}\right] q_a \quad (12)$$

which can be written

$$P_N = \frac{2N^2 r_1}{F^2}\left[a_0^2 + \frac{a_1^2\Delta\omega^2}{2} + \frac{a_1^2 q_a^2}{3}\right] q_a. \quad (12a)$$

The noise-to-signal power ratio is improved by the factor F^2, since

$$\frac{P_N}{P_s} = \frac{1}{F^2}\,\frac{4N^2 r_1}{A^2}\left[\frac{a_0^2}{a_1^2\Delta\omega^2} + \frac{1}{2} + \frac{q_a^2}{3\Delta\omega^2}\right] q_a. \quad (13)$$

Of the factors in (12) the first is the result of modifications of the amplitude of the incoming wave by the disturbance. Although subject to reduction by feedback it can be balanced out completely by the use of differentially connected frequency detectors having slope factors a_1 and $-a_1$. The second term is dependent upon the degree of modulation of the intermediate wave. It is usually of less consequence in its effect upon the listener. The remaining term is the result of phase modulation of the signal wave by the disturbance. Under the condition that the output signal is held constant by increasing the transmitted band, all terms which contribute to the noise level in a given case are reduced alike by feedback.

If differential-frequency detection is employed (12a) becomes

$$\frac{2N^2 r_1 a_1^2}{F^2}\left[\frac{\Delta\omega^2}{2} + \frac{q_a^2}{3}\right] q_a. \quad (14)$$

During nonsignaling periods the first term becomes

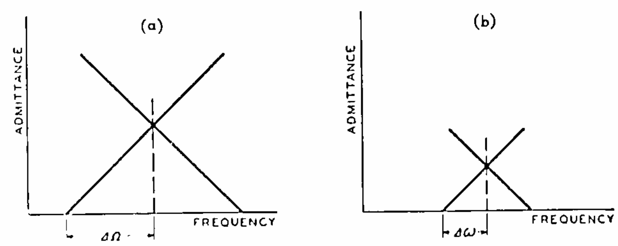


Fig. 2—Idealized conversion-system characteristics for (a) limiter system, (b) feedback system.

zero. Hence during periods of modulation the background noise power is increased by the factor

$$1 + \frac{3}{2}\,\frac{\Delta\Omega^2}{F^2 q_a^2} = 1 + \frac{3}{2}\,\frac{\Delta\omega^2}{q_a^2}\cdot \quad (15)$$

If conditions are such that the maximum shift experienced by the intermediate-frequency wave is numerically equal to q_a, then the noise level will be increased by 4 decibels during periods of full modulation. In the experimental system to be described the ratio of $\Delta\omega$ to q_a was allowed to attain a value of 1.75, resulting in a maximum increase of 7.5 decibels.

In order to secure large noise reduction it is necessary to produce a frequency shift in the transmitted wave much greater than the signal band width. Thus, in common with frequency-modulation systems employing amplitude limiters,[5] this advantage is secured at the expense of band width. In this connection it is of interest to compare amplitude limitation and feedback systems on the basis of equal width of transmitted band. Hence it will be assumed that in

[3] An analysis of the effect of feedback upon noise in this system was first developed by J. R. Carson by methods similar to those used in "Variable frequency electric circuit theory with application to the theory of frequency modulation," Carson and Fry, *Bell Sys. Tech. Jour.*, vol. 16, pp. 513–540; October, (1937). This has been embodied in a paper by Carson entitled, "Frequency modulation; theory of the feedback receiving circuit," publication of which is planned in the July, 1939, issue of the *Bell Sys. Tech. Jour.* Carson's treatment is more general in that an arbitrary signal wave is postulated whereas the analysis given in Appendix B is restricted to a sinusoidal signal wave. The methods used here are more elementary and may therefore appeal to a somewhat wider audience.

[4] The expressions for signal and noise power used in this section are relative. Factors determining their absolute magnitude are given in the Appendix. In all cases the symbol $\Delta\omega^2$ is to be taken as signifying $(\Delta\omega)^2$.

[5] E. H. Armstrong, "A method of reducing disturbances in radio signaling by a system of frequency modulation," Proc. I.R.E., vol. 24, pp. 689–740; May, (1936).

each case the transmitted wave is modulated to the extent of $\pm\Delta\Omega = \pm F\Delta\omega$. In Fig. 2 are shown idealized characteristic of conversion systems which might be used in the two systems. The adjustment shown in Fig. 2(a) is suitable for use with the limiter system. With the feedback system that shown in Fig. 2(b) would be necessary to secure the same percentage of amplitude modulation after conversion. This represents the minimum band width which could be provided in the conversion system with feedback, though several considerations make it desirable to use an adjustment lying somewhere between the two shown. The manner of tuning or the slope factors assumed in each case are immaterial to the present comparison provided that, in either system, the linear portion of the characteristic is of sufficient extent to effect proper conversion of the intermediate-frequency wave.

The noise-to-signal power ratio obtainable with the feedback system will be that given by (13) with the first term omitted since it is balanced out by the push-pull arrangement. Thus

$$\frac{P_N}{P_s} = \frac{4N^2r_1}{F^2A^2}\left[\frac{1}{2} + \frac{q_a^2}{3\Delta\omega^2}\right] q_a. \qquad (16)$$

In the system corresponding to Fig. 2(a), the signal power will be

$$\frac{A^2a_1^2\Delta\Omega^2}{2}. \qquad (17)$$

Equation (10) can be used to determine the noise power level for a frequency-modulation system without amplitude limitation by setting $F=1$. If balanced detection is used the term in a_0 becomes zero. It has been shown by Carson and Fry[3] that the addition of an ideal limiter removes all terms but the third, with either single or balanced detectors. Hence for the limiter system the noise ratio becomes

$$\frac{P_N}{P_s} = \frac{4N^2r_1}{A^2}\left(\frac{q_a^3}{3\Delta\Omega^2}\right). \qquad (18)$$

Since $\Delta\Omega = F\Delta\omega$ this can be put in a form similar to (16)

$$\frac{P_N}{P_s} = \frac{4N^2r_1}{F^2A^2}\left(\frac{q_a^2}{3\Delta\omega^2}\right) q_a. \qquad (19)$$

Comparing (16) and (19) it is seen that the noise ratio in the feedback system is greater than that for the limiter system by the factor (15). This is a consequence of the increase in noise level which occurs during modulation in the former system. The ratio of noise level during nonsignaling periods to signal level is identical in the two systems.

While the noise increment which appears during modulation is usually not of great consequence from a practical standpoint, it can be reduced by increasing the feedback factor beyond the point dictated by the signal band which it is permissible to transmit. In previous discussions it has been assumed that the application of a given amount of negative feedback is to be accompanied by a corresponding increase in modulation level at the transmitter. In this way the modulation of the intermediate-frequency wave is kept constant so as to maintain a fixed signal level as the band width of the transmitted wave is increased. Having arrived at a limiting value of band spread the feedback factor can be increased further. Suppose that modulation of the transmitter is to be limited to a value of $\Delta\Omega = F_1\Delta\omega$, but that the feedback applied to the receiver is made to exceed F_1 by a factor which we will call F_2. Then the actual feedback factor will be F_1F_2 and we will have

$$P_s = \frac{A^2a_1^2\Delta\omega^2}{2F_2^2} \qquad (20)$$

$$P_N = \frac{2N^2r_1}{F_1^2F_2^2}\left[\frac{a_1^2\Delta\omega^2}{2F_2^2} + \frac{a_1^2q_a^2}{3}\right] q_a \qquad (21)$$

giving

$$\frac{P_N}{P_s} = \frac{4N^2r_1}{A^2F_1^2}\left[\frac{1}{2F_2^2} + \frac{q_a^2}{3\Delta\omega^2}\right] q_a. \qquad (22)$$

Thus the additional feedback represented by the factor F_2 is directly effective against the noise increment accompanying modulation. Reduction of this increment brings about a still closer correspondence between the limiter and feedback systems as is seen by setting $F=F_1$ in (19) and comparing with (22).

The above discussion and the analysis given in Appendix B are based upon the assumption that the carrier amplitude is large compared with that of the disturbance. A rigorous analysis, applicable to the case where this ratio is unrestricted, becomes exceedingly involved. However, a rough indication of what is to be expected in the presence of a high level of disturbance can be obtained quite simply from (52) developed in Appendix B. Assuming that modulation is not present this can be put in the simple form

$$\sigma = \frac{1}{F}\,\frac{Q'(a_0 + a_1\omega_n)}{1 + \dfrac{Q'}{A'}\left(\dfrac{F-1}{F}\right)\cos\omega_n t}\cos\omega_n t. \qquad (52a)$$

When $Q' \ll A'$ the wave form of the output noise produced by a single element of disturbance is very closely a sinusoid. However, when Q' and A' become comparable in magnitude the output wave becomes badly peaked when $\omega_n t = n\pi$. While the above expression is only a very rough approximation under these conditions, a plot of the wave form so obtained exhibits all of the esential characteristics of the curves given by Crosby in a recent paper[6] dealing with noise in frequency-modulation systems using amplitude limitation. These curves show a similar peaking of the output-noise wave form when the ratio of carrier to disturbance amplitude is in the vicinity of unity. The description given by Crosby of the manifestations of this phenomenon observed in an experimental system applies rather closely to what has been found in the feedback system. A more detailed account will be found in a later section.

Examination of (52a) shows that the output wave can assume very large and even infinite peak values when Q' and A' are approximately equal. The existence of high peak values of noise implies both a large instantaneous deviation in the frequency of the intermediate wave, and a conversion-circuit characteristic of unlimited extent. The finite limits of the characteristic of the over-all intermediate-frequency system have the effect of holding the maximum peaks of noise to a value equal to the highest signal peaks obtainable in the absence of the disturbance. Furthermore, the existence of high noise peaks in the presence of modulation can result in the momentary assumption by the instantaneous intermediate frequency of values outside of the region to which the system is normally responsive. Thus the output signal will appear to be chopped by the higher noise peaks, and as a consequence its energy content will be considerably reduced.

The above effects are, of course, present in systems using limiters and have already been discussed in greater detail by Crosby.[6]

Distortion Reduction

One of the chief benefits which can be realized through the use of negative feedback is the reduction of nonlinear distortion products generated in the forward branch of the system. While the distortion in properly designed amplifiers is sufficiently low for many purposes, cases frequently arise in which the requirements are much more severe. In an amplifier which is to handle several channels in a high-grade multiplex system, the distortion products should be of the order of 60 decibels below the fundamental of the output. This degree of excellence is most readily obtained by using negative feedback.

In radio systems designed for multiplex service it is of equal importance that the distortion level be kept at a correspondingly low level if cross talk is to be avoided. It is therefore of interest to inquire into the manner in which distortion is modified in the present feed-back system.

An analysis of the effect of feedback upon distortion is given in Appendix A. If the transmitter is modulated with a signal wave $S = S(t)$ so that its instantaneous frequency becomes

$$\omega + \rho_1 S \tag{23}$$

then, in the presence of nonlinearity in the receiver, the output of the system can be written as a power series in the variable frequency term $\rho_1 S$. Thus for the first three orders we shall have

$$\sigma = \alpha AB[b_1\rho_1 S + b_2(\rho_1 S)^2 + b_3(\rho_1 S)^3]. \tag{24}$$

If feedback is applied without altering the modulation level at the transmitter it is shown that the above series becomes

$$\sigma_F = \alpha AB\left(\frac{b_1}{F}\rho_1 S + \frac{b_2}{F^3}(\rho_1 S)^2 + \frac{1}{F^4}\left[b_3 - \frac{2b_2^2}{b_1}\left(\frac{F-1}{F}\right)\right](\rho_1 S)^3\right). \tag{25}$$

When the feedback factor F is large this can be written

$$\sigma_F = \alpha AB\left(\frac{b_1}{F}\rho_1 S + \frac{b_2}{F^3}(\rho_1 S)^2 + \frac{1}{F^4}\left[b_3 - \frac{2b_2^2}{b_1}\right](\rho_1 S)^3\right). \tag{26}$$

Upon increasing the modulation by the factor F so as to restore the original level of the fundamental, the output becomes

$$\sigma_F' = \alpha AB\left(b_1\rho_1 S + \frac{1}{F}\left[b_2(\rho_1 S)^2 + \left(b_3 - \frac{2b_2^2}{b_1}\right)(\rho_1 S)^3\right]\right). \tag{27}$$

Second-order distortion products are reduced with respect to the fundamental level by the feedback factor. Third-(and higher-) order products are modified to an extent depending upon the relative values of the distortion coefficients and the amount of feedback. If, as can readily be the case when a balanced detecting system is used,

$$b_3 \gg \frac{2b_2^2}{b_1} \tag{28}$$

third-order products are reduced in the same manner as those of second order. In any case by applying

[6] Murray G. Crosby, "Frequency modulation noise characteristics," Proc. I.R.E., vol. 25, pp. 472–514; April, (1937). The curves referred to are given in Fig. 4 of the above paper.

sufficient feedback a point will be reached where a given increment in feedback will produce a corresponding reduction in all distortion products.

Equation (25) shows that the greatest improvement in distortion is obtained if the modulation level is not increased when feedback is applied. The large reductions result partly from feedback and in part from the fact that the system is operating at reduced percentage of modulation. Under this condition there is no improvement in background-noise ratio, though the noise increment which takes place during modulation is diminished. (See (10) and (11).) Depression

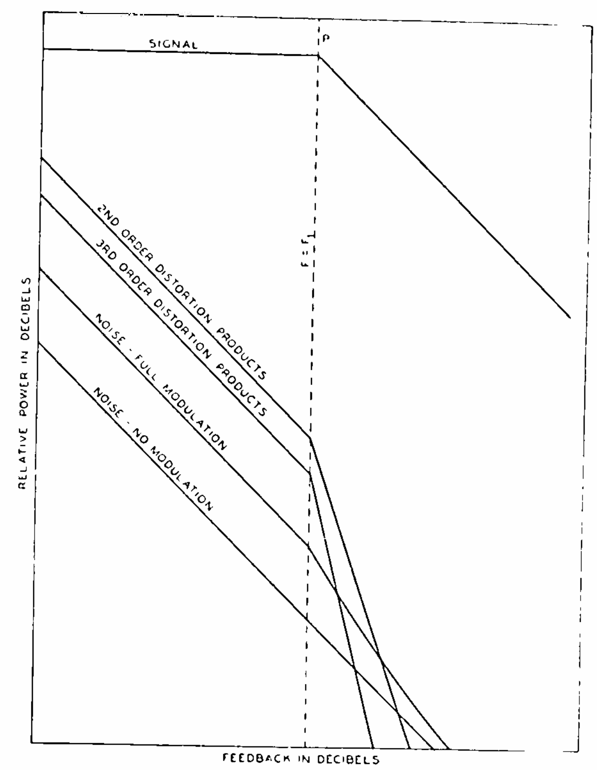


Fig. 3—Theoretical manner in which components of receiver output are modified by feedback. Modulation level at transmitter is assumed to be increased by the feedback factor up to point B, and subsequently to remain fixed.

of both noise and distortion, but with greater emphasis upon the reduction of the latter, can be effected by raising the modulation level by an amount somewhat less than the feedback factor. This procedure has already been discussed in connection with (22) which gives the resulting noise-to-signal power ratio. Under similar conditions we have, from (25),

$$\sigma_F'' = \alpha AB\left(\frac{b_1}{F_2}\rho_1 S + \frac{1}{F_1}\left[\frac{b_2}{F_2{}^3}(\rho_1 S)^2 + \frac{1}{F_2{}^4}\left(b_3 - \frac{2b_2{}^2}{b_1}\right)(\rho_1 S)^3\right]\right) \qquad (29)$$

when the feedback factor is large.

Equations (22) and (29) are most readily interpreted by means of Fig. 3, which illustrates the manner in which the receiver output is modified as the feedback is increased.

It is assumed that a constant signal level is maintained by an increase in modulation level up to a point corresponding to the factor F_1. Beyond this point the modulation remains fixed while the feedback factor is increased to F_1F_2. Since signal and noise levels have been expressed in terms of power, distortion levels are similarly expressed. These levels, to an arbitrary decibel scale, have been plotted against decibels of feedback, given by the expression

$$10 \log F^2 = 20 \log F.$$

Balanced detection and the fulfillment of condition (28) have been assumed.

Over the region in which a constant signal output is maintained by increasing the modulation level, noise and distortion levels decrease in accordance with the feedback. The noise level during modulation continues to exceed the background noise by 4 decibels, assuming an initial frequency shift equal to the highest signal frequency to which the system is responsive.

Beyond the point at which the feedback factor has reached the value F_1, the modulation level at the transmitter is held constant. A further increase in feedback brings about a corresponding decrease in the effective percentage of modulation for the system, causing the signal level to fall in similar fashion. Distortion products now fall off still more rapidly with respect to the signal, so that an increase in feedback amounting to 1 decibel improves the second- and third-order distortion ratios by 2 and 3 decibels, respectively.

The ratio of signal to background or nonsignaling noise remains fixed in this region in spite of the reduction in effective modulation. This ratio is that which would be obtained in a limiter system in which the same high-frequency band is transmitted. The noise increment, however, is diminished by the additional feedback and is made to approach zero.

By suitable choice of the variables F_1 and F_2 it is possible to proportion the benefits of feedback in the most advantageous manner. Thus if noise is of more consequence than distortion, modulation would be increased to the full extent of the feedback; if distortion is of primary concern, as it might well be in a multiplex system, operation as indicated in Fig. 3 would be preferable.

Experimental Results

Description of Equipment

Experimental confirmation of the principles which have been outlined has been obtained with the aid of a laboratory system shown schematically in Fig. 4. This arrangement provided a transmitter, receiver, and source of disturbance all under local control. The transmitter operated at a carried frequency of 20 megacycles. This was frequency-modulated by means of a circuit basically similar to that described by Travis.[7] Tube noise voltage appearing at the output of a high-gain radio-frequency amplifier supplied the high-frequency disturbance.

At the receiver an oscillator similar to that at the transmitter served to beat down the incoming wave to an intermediate frequency of 438 kilocycles. This was applied to a three-stage amplifier having substantially uniform gain over a band of 50 kilocycles, and thence delivered to a balanced frequency detector. In addition to signal voltage, automatic-frequency-control potentials were derived from the detectors. Both were carried back to the local oscillator, but in order to permit independent control of the amount of feedback their respective paths were kept separate. In this way full frequency control could be had even with signal feedback reduced to zero.

Details of the frequency detector and feedback connections are shown in Fig. 5. The conversion system derives its characteristics from antiresonant circuits L_1C_1 and L_2C_2, double-winding coils being used to isolate the rectifier anodes from the plate battery. One circuit is tuned to a frequency 15.4 kilocycles above the intermediate-carrier frequency and the other to a corresponding point below, their characteristics intersecting at a point where the gain is approximately one half of the peak value. Detection takes place in linear rectifiers D_1 and D_2. By means of the arrangement shown, signal potentials are impressed upon the grids of amplifiers A_1 and A_2, while frequency-control voltage appears across condensers C_3, C_4. This voltage becomes zero when

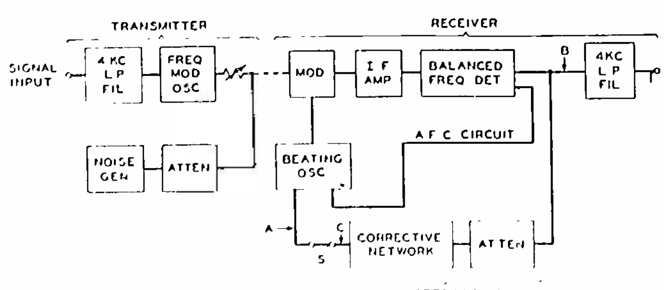


Fig. 4—Schematic of experimental feed-back system.

the receiver is correctly tuned and appears with proper polarity to produce correction of the frequency of the local oscillator in case of slow drifts in the frequency of either oscillator.

The use of a conversion system having peaks separated by an amount considerably exceeding the greatest frequency deviation is the result of a compromise between the readily adjustable and high-impedance antiresonant type of load circuit and others which, though more linear in their characteristics, lead to much lower gain in the conversion

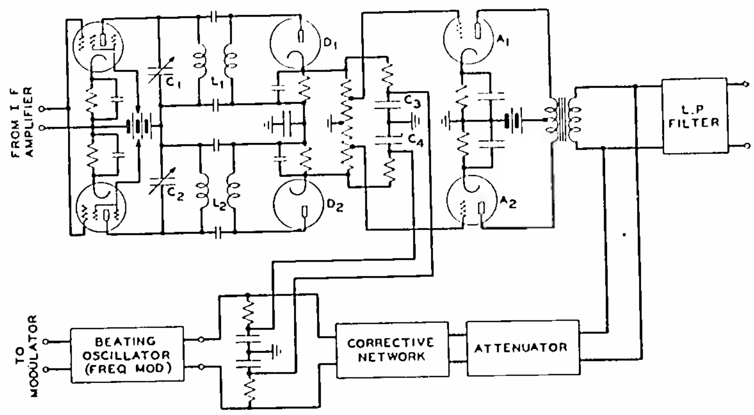


Fig. 5—Details of balanced frequency detector and feedback connections.

stage. While a peak separation of 14 kilocycles would have sufficed in view of the limitations of the transmitter, a considerably greater peak separation without corresponding increase in modulation was used. As a result that portion of the circuit characteristic actually embraced by the modulated intermediate-frequency wave presented a much better approximation of a straight line than would have been possible with minimum peak separation. The penalty for adjusting the circuits in this manner is merely a loss in detecting efficiency and not an impairment of the signal-to-noise ratio. This can readily be overcome by additional audio-frequency amplification.

The signal-frequency feedback path includes an attenuator for adjusting the feedback and a corrective network for preventing singing around the feedback loop. Frequency control and feedback paths are finally combined at the modulation terminals of the local oscillator.

The necessity for the inclusion of a corrective network to modify the transmission characteristics of the feedback path is evident from Fig. 6. This shows the measured gain and phase characteristics of the receiver alone, viewed as a voice-frequency network between points A and B in Fig. 4. This was obtained by applying signal frequencies to the modulation terminals of the beating oscillator and making observations at point B with switch S open, proper termination being provided at both sides of the break. The unmodulated transmitter served, in effect, as the beating oscillator during this measurement. At the lower signal frequencies the phase is practically 180 degrees as indicated by (4) with $x_1 = 0$. As the signal frequency is increased the phase

[7] Charles Travis "Automatic frequency control," Proc. I.R.E., vol. 23, pp. 1125–1141; October, (1935).

is progressively shifted from this value. Except for that produced by the output transformer, the shift takes place within the intermediate-frequency amplifier and conversion circuits. Its magnitude is a measure of the slope of the phase-frequency characteristic of the intermediate-frequency system.

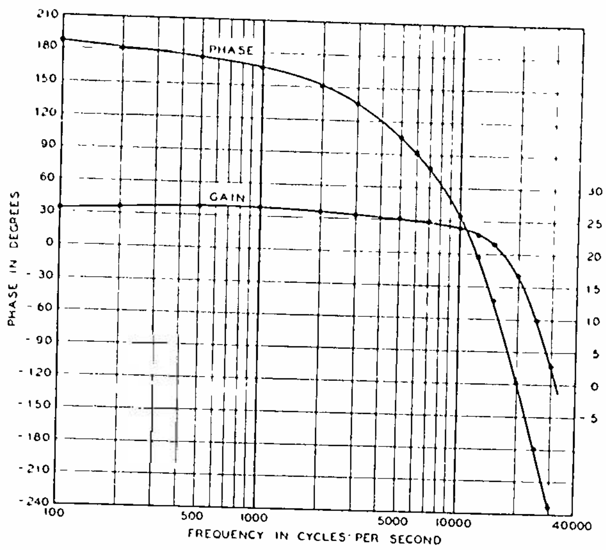


Fig. 6—Phase and gain characteristics of receiver measured between points *A* and *B* of Fig. 4 with switch *S* open. Transmitter in operation but not modulated.

The existence of positive gain at a point of zero phase shows that singing would occur if feedback connections were made directly to the beating oscillator. It was therefore necessary to reduce the gain below unity at the point of zero phase. This was accomplished by including in the feedback path a network designed by R. L. Dietzold. The gain-frequency characteristic of this network is shown in Fig. 7. The modified-loop characteristics as measured between points *A* and *C* with switch *S* open, and with the attenuator set for an 8-decibel loss, are given in Fig. 8. Full feedback is applied only over a band extending to 4 kilocycles, so that the range of frequencies applied to the transmitter and delivered to the listener must be restricted to this figure. The limit of stable feedback which can be realized is indicated by the difference between the loop gain within the useful band and that at the frequency corresponding to zero phase.

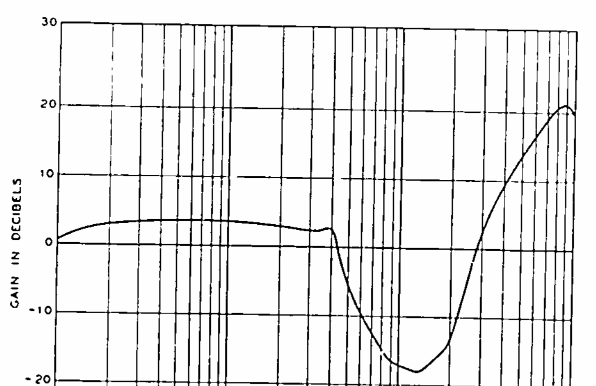


Fig. 7—Gain-frequency characteristic of corrective network inserted in signal feedback path.

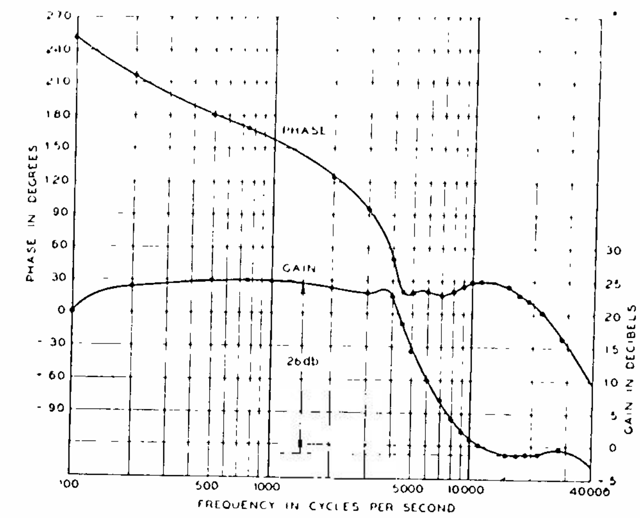


Fig. 8—Gain and phase characteristics of complete feedback loop including corrective network. Measured between points *A* and *C* of Fig. 4 with switch *S* open.

Distortion Measurements

The manner in which distortion levels at the output of the receiver were observed to vary with feedback is depicted in Figs. 9 to 12. In each case the modulation level for zero feedback was such as to shift the frequency of the transmitter ± 7 kilocycles at the rate of 1000 cycles per second. Fig. 9 shows the effect of increasing the modulation in proportion to the feedback so as to maintain a constant output level for the fundamental. Both second and third harmonics tend to be reduced in proportion to the feedback, the improvement in third-harmonic level

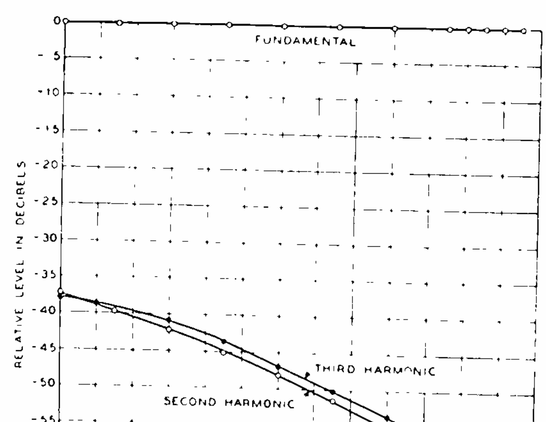


Fig. 9—Effect of feedback upon receiver distortion. Fundamental level kept constant by increasing transmitter modulation in proportion to the feedback. Modulation with no feedback $= \pm 7$ kilocycles at 1000 cycles per second.

being 23.5 decibels for 25-decibel feedback. Failure to realize full reduction of the second harmonic is the result of distortion beginning to manifest itself in one or the other of the modulated oscillators. At the point of 25-decibel feedback the transmitter and beating oscillator were being modulated to the extent of ± 124.5 and ± 117.5 kilocycles, respectively.

The curves of Fig. 10 were obtained by maintaining a constant fundamental level up to the point of 15-decibel feedback and then allowing the modulation level at the transmitter to remain constant thereafter. The results correspond rather closely with the theoretical curves of Fig. 3 and show the very rapid decrease in distortion which takes place

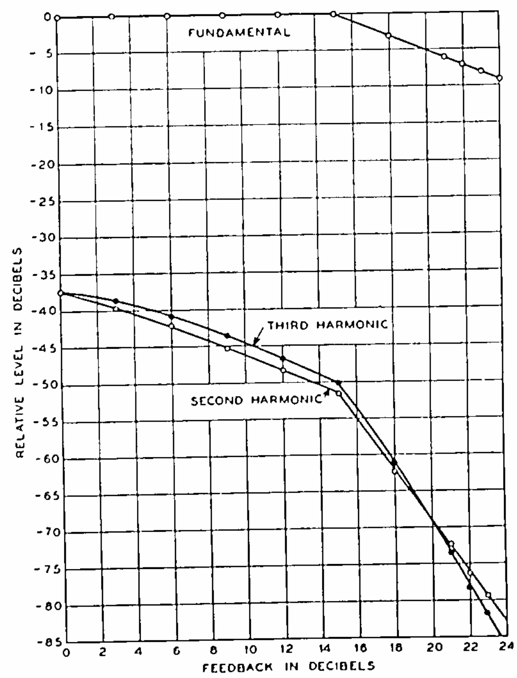


Fig. 10—Effect of feedback upon receiver distortion. Conditions same as indicated for Fig. 9 up to 15-decibel feedback; modulation held constant thereafter.

when the modulation level remains unaltered. (See (25).) A more extreme example of this method of operation is shown in Fig. 11 where modulation was left at its initial value. Harmonic levels soon reached a point beyond which they could not be measured accurately.

In a practical system the loss in signal level resulting from operation in this manner could easily be overcome by the addition of a low-distortion audio-frequency amplifier at the output of the receiver. This amplifier might well embody negative feedback of the more usual type.

A composite of these distortion measurements is given in Fig. 12. Harmonic levels are plotted in decibels below the fundamental and are indicative of the improvements brought about by feedback. If it is assumed that any loss in signal is compensated by additional audio-frequency amplification, the fundamental level would be represented in all cases by the axis of abscissas.

Noise Measurements

In Fig. 13 are given the results of a series of observations of receiver output noise versus amount

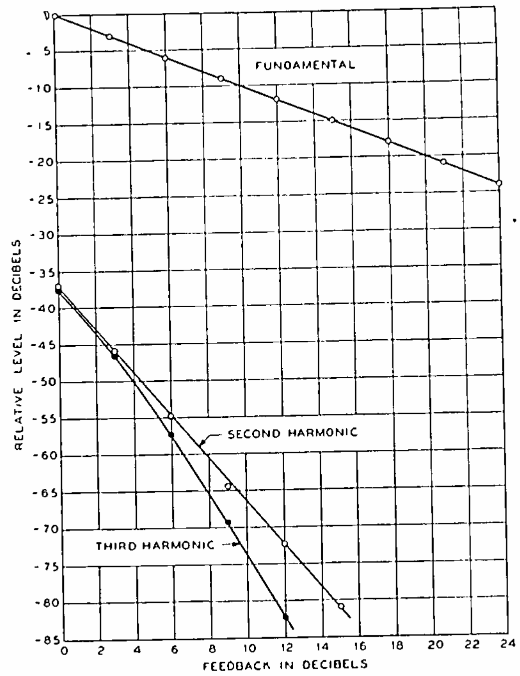


Fig. 11—Effect of feedback upon receiver distortion. Modulation held to a constant value of ± 7 kilocycles.

of feedback for a number of high-frequency disturbance levels. Measurements were made in the absence of modulation and hence are indicative of the manner in which background noise is modified by feedback.

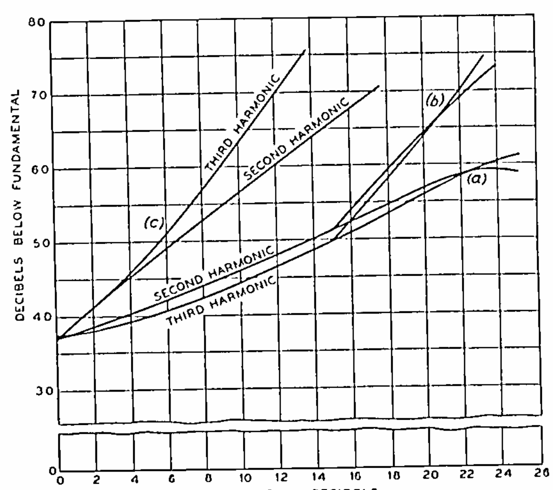


Fig. 12—Composite of data from Figs. 9 to 11 expressing ratios of harmonic levels to fundamental level. Curves (*a*) from Fig. 9, (*b*) from Fig. 10, and (*c*) from Fig. 11.

The signal level indicated is that which could be maintained at low noise levels by increasing the modulation in proportion to the feedback, and is not significant for observations falling within or close to the shaded area, as will be explained subsequently.

The lowest noise level shown is that generated within the receiver while the higher levels were produced by disturbances introduced from the noise generator. The relative magnitude of the effective carrier and disturbing voltages at the grids of the amplitude detector is indicated on each curve. This

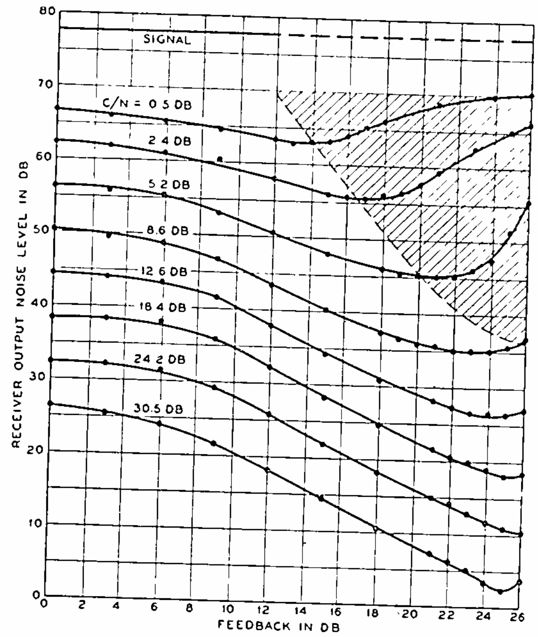


Fig. 13—Effect of feedback upon receiver output noise level for various amounts of high-frequency disturbance. Ratio of root-mean-square values of carrier and disturbance is shown on each curve. Shaded area indicates region of "crackling" in the absence of modulation.

was obtained in the following manner: With the transmitter turned off the noise attenuator was adjusted until the introduced disturbance produced the same value of rectified current as that observed when the carrier alone was applied. This determined the input level from the noise generator which produced equal root-mean-square values of intermediate-frequency carrier and disturbance. Since at very low inputs from the noise generator the net intermediate-frequency disturbance was determined partly by tube noise generated within the receiver, a curve of output noise without feedback versus input from the noise generator was obtained. In the region where the effect of receiver tube noise was evident the assumption of a linear relationship between disturbance level and output noise permitted correction of the curve so that equivalent disturbance levels could be related to any setting of the noise attenuator, or to the receiver output noise level without feedback.

The signal-to-noise ratios at the output of the receiver without feedback are considerably higher than the corresponding ratios of carrier and disturbance levels existing at the amplitude detectors. This is the result of two factors. The intermediate-frequency wave is modulated to the extent of ± 7 kilocycles while the range of frequencies appearing at the output terminals of the receiver is limited to 4 kilocycles. Hence at the output of the balanced detector the noise level in the absence of modulation is 9.6 decibels below that which would be observed at the output of an amplitude-modulation system. Furthermore the admittance characteristic of the complete intermediate-frequency system is such that the effective disturbing voltage delivered to the amplitude detectors is 11 decibels greater than that admitted by an amplitude-modulation system having the minimum intermediate-frequency band width of 8 kilocycles.

Aural observation of the character of the output noise showed that, excluding the shaded area in Fig. 13, the normal characteristics of fluctuation noise are preserved as feedback is applied. Upon crossing the boundary of the shaded area the noise becomes punctuated with intermittent clicks which increase in rapidity and violence as the feedback factor is raised, giving rise to what can be described as "crackling." After passing through a region of maximum turbulence the noise gradually assumes the nature of a much higher level of fluctuation noise.

The region embracing the appearance of the above phenomenon is also characterized by a marked reduction in signal level. At high modulation levels "cracking" begins at a somewhat lower disturbance level than is necessary to initiate it in the absence of modulation. The initial effect is to impart a roughness to tone modulation. Further increase in disturbance level produces a rapid depression of the signal, so

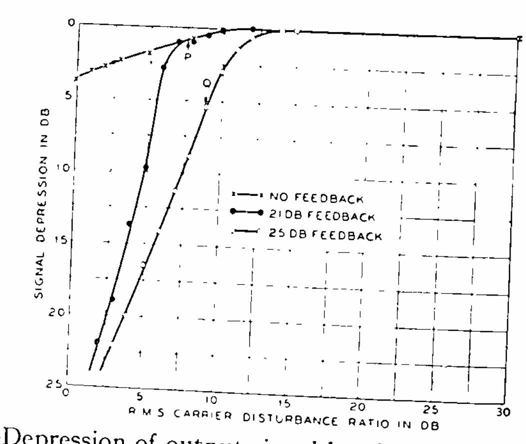


Fig. 14—Depression of output signal by the disturbance. Modulation: ± 7 kilocycles for no feedback, ± 78.5 kilocycles for 21-decibel feedback, and ± 124.5 kilocycles for 25-decibel feedback. Points P and Q denote incidence of crackling in the absence of modulation, for 21- and 25-decibel feedback, respectively.

that it soon becomes submerged in noise. The manner in which this depression takes place is shown in Fig. 14. The signal, produced by 1000-cycle modulation was measured by means of a highly selective analyzer so that observations could be carried well below the general noise level.

The point at which depression of the signal begins coincides with the appearance of roughness in the output tone resulting from the momentary suppression of the signal by the higher noise peaks. A further increase in disturbance level increases the number of peaks per second which rise above the critical value, and the energy content of the signal is rapidly diminished. The point at which faint crackling could first be detected in the absence of modulation is indicated on each curve.

The signal-to-noise ratios obtained with zero and with 25 decibels of feedback are shown in Fig. 15. These are plotted against the ratio of root-mean-square carrier and disturbance levels at the end of the intermediate-frequency channel. Signal levels were measured in the presence of the disturbance so as to take into account the depression of the signal, while noise levels were observed in the absence of modulation.

The improvement resulting from the application of feedback is given by the difference between the two curves and approaches the theoretical improvement at the low noise levels. The curve obtained with feedback exhibits a rather sharp break when the ratio of carrier to effective disturbance is in the vicinity of 10 decibels. Experimental data[6] published by Crosby indicates that in the case of fluctuation noise the ratio of the maximum peak amplitude to the root-mean-square value is about 13 decibels. The corresponding figure in the case of a sine wave is 3 decibels. Hence equality of carrier peak amplitude and the highest peaks of the disturbance obtains when the ratio of their root-mean-square values is 10 decibels. With feedback this condition appears to define a fairly critical disturbance level above which the output signal-to-noise ratio is very rapidly diminished. Crosby has shown[6] that with systems employing amplitude limiters a similar condition marks the point beyond which the noise improvement realized at the lower disturbance levels is soon lost. This point he has termed the "threshold of noise improvement."

A less sharply defined break also occurs in the curve expressing noise conditions in the absence of feedback. This is the result of a progressive destruction, at the higher disturbance levels, of the balancing out of amplitude effects in the push-pull detector which is realized when the noise is low.

While direct comparison of the feedback system with an actual amplitude-modulation system was not possible with the equipment used, it is thought that a comparison based upon theoretical considerations may be of interest. The procedure is as follows: The noise ratios shown in Fig. 15 for the system without feedback are, for disturbances below the threshold value, 9.6 decibels in excess of those which would be realized in a fully modulated amplitude system. A dotted line, displaced from the linear portion of the measured curve by this amount, is shown. The abscissas of the dotted curve do not represent the true carrier-disturbance ratio which would obtain in the amplitude system for the reason that, ideally, the intermediate-frequency amplifying system would have a band width of but 8 kilocycles. In such a system the signal-to-noise ratio would be equal to the carrier-disturbance ratio except at the very high noise levels. Hence the intercept of the dotted line with the axis of abscissas marks the point of equal carrier and disturbance levels in this system.

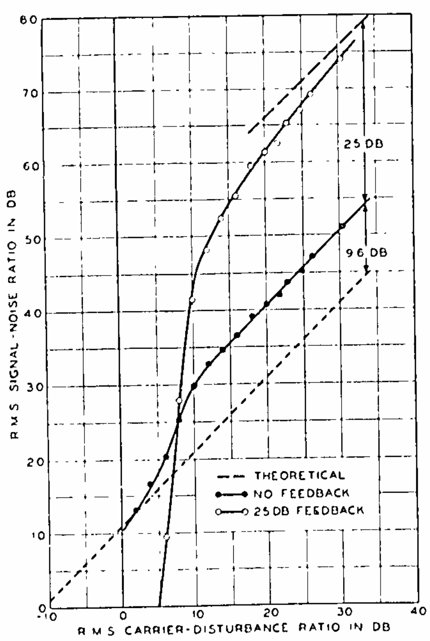


Fig. 15—Output signal-to-noise ratio vs. carrier-disturbance ratio with and without feedback. Modulation ±124.5 kilocycles for 25-decibel feedback and ±7 kilocycles for no feedback.

The difference of 11 decibels between this point and the zero point on the scale as drawn measures the amount by which the disturbance at the rectifiers in the experimental system exceeds that which would be found in the ideal amplitude system.[8] Consequently, if it is desired to relate the data of Fig. 15 to the disturbance ratio which would exist at the input to the detector in the amplitude system, and hence to the signal-to-noise ratio in that system, it is merely necessary to displace the experimental curves to the right by 11 decibels.

Fig. 16 shows such a comparison between the feedback system adjusted to give 25 decibels of feedback, and an ideal amplitude-modulation system. There is also included a curve showing the theoretical

[8] Comparison of the areas under idealized curves representing the square of the transmission through the intermediate-frequency systems in the two cases indicates a difference of 10.1 decibels.

performance which would be approached by a frequency-modulation system using amplitude limitation. Transmitted band width and audio-frequency response equal to that used in the experimental feedback system have been assumed. This corresponds to a deviation ratio of 124.5 kilocycles ÷ 4 kilocycles = 31.1, resulting in a theoretical noise deduction of 34.6 decibels at low disturbance levels. The threshold of noise improvement, indicated at the point x, is located at a point where the peak signal-to-noise ratio in the amplitude system is equal to the square root of the deviation ratio.[6] This factor takes account of the higher disturbance level in the intermediate-frequency channel of the wide-band system. Assuming a factor of 10 decibels between

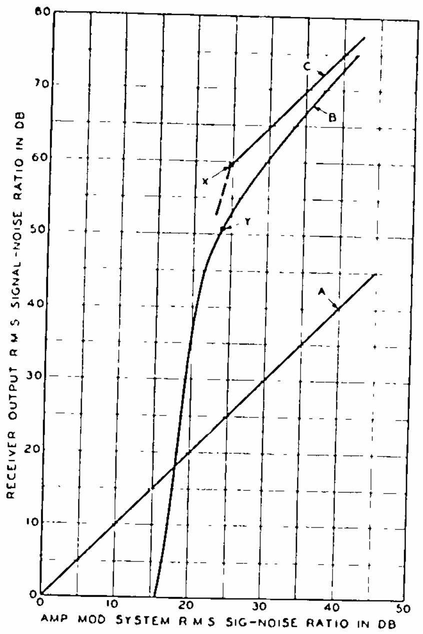


Fig. 16—Theoretical comparison of signal-to-noise ratios obtained at 25-decibel feedback (curve B) with amplitude-modulated system (curve A) and ideal limiter system with deviation ratio = 31.1 (curve C). Point X = threshold of noise improvement for limiter system. Point Y = point where "crackling" first became evident in the presence of ±124.5-kilocycle modulation.

maximum peak and root-mean-square values of fluctuation noise this corresponds to a root-mean-square signal-to-noise ratio of 24.9 decibels in the amplitude system.[9] In the feedback system the point at which crackling was first observed in the presence of modulation is shown at Y. This coincides very closely with the theoretical threshold of noise improvement in the limiter system.

Conclusions

It has been shown that the application of negative feedback to a frequency-modulation system affords a means for effecting large reductions in both noise and receiver distortion. The theoretical analyses of these effects, while they have been simplified to an extent which makes them inadequate to cover all conditions which can be encountered in practice, are adequately substantiated by the observed performance of the experimental system within the limitations of the theory.

Substantial benefits are realized only when the amount of feedback is large, and when the disturbance level is not too great. In common with frequency-modulation systems employing amplitude limitation a large reduction in noise must be paid for by increasing the band width of the transmitted wave. While the principles involved in the two systems are quite different, their performance as regards noise modification, both at high and at low levels of disturbance, exhibits striking similarities. The ability to reduce distortion is, on the other hand, a feature found only in the feedback system.

Acknowledgment

The writer wishes to acknowledge his indebtedness to the following of his colleagues: To Dr. H. W. Bode for his investigations of the problem of stability in feedback systems, and to Mr. R. L. Dietzold for the design of the stabilizing network which was used; to Mr. W. R. Bennett whose unpublished work on basic frequency-modulation problems has been of great value; and to Messrs. E. A. Krauth and O. E. DeLange for assistance in the experimental work. Reference has already been made to the theoretical work of Dr. J. R. Carson.

Appendix A

Analysis of Distortion Reduction

Assume that the transmitter is frequency-modulated with a signal wave which we shall represent by the symbol $S = S(t)$. Then the instantaneous frequency of the transmitter will be

$$\omega_1 + \rho_1 S. \tag{30}$$

The instantaneous phase of the transmitted wave is the integral of this expression and the voltage delivered to the input of receiver can be written

$$A \cos\left(\omega_1 t + \rho_1 \int_0^t S dt + \phi_1\right). \tag{31}$$

Designating the low-frequency voltage delivered at the output of the receiver as $\sigma = \sigma(t)$ the result of feeding back a portion of $k\sigma$ of the output so as to frequency-modulate the local oscillator is the wave

$$B \cos\left(\omega_2 t + \rho_2 \int_0^t k\sigma dt + \phi_2\right). \tag{32}$$

[9] In the absence of more exact information regarding the performance of a system using this large a deviation ratio the portion of this curve below the threshold has been omitted.

Application of these two waves to the modulator produces the intermediate-frequency product[10]

$$\alpha AB \cos \left[\omega_0 t + \rho_1 \int_0^t S dt - \rho_2 \int_0^t k\sigma dt + \phi_0\right] \quad (33)$$

where

$$\omega_0 = \omega_1 - \omega_2$$

$$\phi_0 = \phi_1 - \phi_2.$$

Terms in the above which involve the integral sign represent phase angles which vary with time. Hence we shall rewrite (33) more compactly

$$\alpha AB \cos [\omega_0 t + \theta(t) + \phi_0]. \quad (34)$$

It has been shown by Carson and Fry[3] that the process of detecting a frequency-modulated wave is, in effect, its differentiation. Since the high-frequency wave itself exhibits the integral of the signal wave (see (30)) it can be reasoned that a differentiation process is necessary for the recovery of the signal itself.

Differentiation of the argument of the cosine term in (31) yields the instantaneous frequency (rate of change of phase with respect to time) of the received wave given by (30). Now it can be shown that with a strictly linear frequency detector, the low-frequency output is proportional to the response of the conversion system at the instantaneous frequency. The recovered signal is, therefore, proportional to the variable part of the instantaneous frequency and hence to the time derivative of the variable phase term in the original wave.

In the case of nonlinearity in the characteristic of the frequency detector the output can be expressed, to a sufficiently close degree of approximation, as a power series in the derivative of the phase term $\theta(t)$. Hence the output of the receiver can be written in the form

$$\sigma(t) = \alpha AB \sum_n b_n \left[\frac{d}{dt}\theta(t)\right]^n \quad (35)$$

$$= \alpha AB \sum_n b_n [\rho_1 S - k\rho_2 \sigma]^n \quad (36)$$

where the coefficients b_n are based upon the transfer admittance characteristic of the receiver.

What is now desired is the relationship between σ and S. This can be expressed in the general form

$$\sigma = \alpha AB \sum_n c_n [\rho_1 S]^n. \quad (37)$$

Equations (36) and (37) can now be equated. Replacing σ in the right-hand side of (36) by the series (37) we shall have

$$\begin{aligned} &c_1\rho_1 S + c_2(\rho_1 S)^2 + c_3(\rho_1 S)^3 + \cdots \\ &= b_1[\rho_1 S - k\alpha AB\rho_2(c_1\rho_1 S + c_2(\rho_1 S)^2 + c_3(\rho_1 S)^3 + \cdots)] \\ &+ b_2[\rho_1 S - k\alpha AB\rho_2(c_1\rho_1 S + c_2(\rho_1 S)^2 + c_3(\rho_1 S)^3 + \cdots)]^2 \\ &+ b_3[\rho_1 S - k\alpha AB\rho_2(c_1\rho_1 S + c_2(\rho_1 S)^2 + c_3(\rho_1 S)^3 + \cdots)]^3 \\ &+ \cdots . \end{aligned} \quad (38)$$

After expanding, coefficients of like powers of $\rho_1 S$ can be equated. Then solving for the first three orders of c_n we find

$$c_1 = \frac{b_1}{1 + \alpha b_1 kAB\rho_2} = \frac{b_1}{1 - \mu\beta} \quad (39)$$

$$c_2 = \frac{b_2}{(1 - \mu\beta)^3} \quad (40)$$

$$c_3 = \frac{b_3}{(1 - \mu\beta)^4} - \frac{2\alpha ABk\rho_2 b_2^2}{(1 - \mu\beta)^5} \quad (41)$$

where

$$\mu = \alpha ABb_1 \quad \text{and} \quad \beta = -k\rho_2.$$

Inserting these values in (37) and writing $(1 - \mu\beta) = F$, the receiver output becomes, with feedback,

$$\sigma_F = \alpha AB \left[\frac{b_1}{F}\rho_1 S + \frac{b_2}{F^3}(\rho_1 S)^2 + \left(\frac{b_3}{F^4} - \frac{2b_2^2}{b_1} \cdot \frac{F-1}{F^5}\right)(\rho_1 S)^3\right]. \quad (42)$$

When $F \gg 1$ this can be written

$$\sigma_F = \alpha AB \left[\frac{b_1}{F}\rho_1 S + \frac{b_2}{F^3}(\rho_1 S)^2 + \frac{1}{F^4}\left(b_3 - \frac{2b_2^2}{b_1}\right)(\rho_1 S)^3\right]. \quad (43)$$

Without feedback we have

$$\sigma = \alpha AB[b_1\rho_1 S + b_2(\rho_1 S)^2 + b_3(\rho_1 S)^3]. \quad (44)$$

Appendix B

Analysis of Noise Reduction

In the following analysis it is assumed that the amplitude of the disturbance producing the noise is sufficiently small compared with that of the incoming signal wave so that the principle of superposition will apply. Hence the manner in which the effect of a single disturbing component is modified by feedback will first be developed. Then the effect of a disturbance consisting of a continuous spectrum is derived by direct summation.

[10] This expression constitutes a more general form of (4).

Consider first the case where there are impressed upon the grid of the modulator the incoming wave and the local oscillator voltage[11] as defined by (31) and (32), plus a single disturbing component

$$Q \cos [(\omega_1 + \omega_n)t + \phi_n]. \tag{45}$$

Then the intermediate-frequency product will be

$$\alpha AB \cos\left[\omega_0 t + \rho_1 \int_0^t S dt - \rho_2 \int_0^t k\sigma dt\right]$$
$$+ \alpha BQ \cos\left[(\omega_0 + \omega_n)t - \rho_2 \int_0^t k\sigma dt + \phi_n\right]. \tag{46}$$

For simplicity assume that the intermediate-frequency amplifier and conversion circuit have the ideal transfer admittance characteristic

$$Y(\omega) = a_0 + a_1(\omega - \omega_0). \tag{47}$$

Then all derivatives of Y with respect to ω above the first are zero, and the steady-state response is equal to its response at the instantaneous frequency of the applied wave. Hence after conversion we shall have

$$\alpha AB[a_0 + a_1(\rho_1 S - \rho_2 k\sigma)] \cos\left[\omega_0 t + \int_0^t (\rho_1 S - \rho_2 k\sigma) dt\right]$$
$$+ \alpha BQ[a_0 + a_1(\omega_n - \rho_2 k\sigma)]$$
$$\cos\left[(\omega_0 + \omega_n)t - \int_0^t \rho_2 k\sigma dt + \phi_n\right]. \tag{48}$$

Application of (48) to a linear amplitude detector will yield a low-frequency output proportional to its amplitude. The amplitude factor is readily calculated for the case where $AB \gg BQ$. For if

$$X \cos x + Y \cos y = Z \cos z$$

then

$$Z = \sqrt{X^2 + Y^2 + 2XY \cos(x - y)}$$

and when $X \gg Y$

$$Z \doteq X + Y \cos(x - y). \tag{49}$$

Hence the output of the linear detector will be

$$\gamma\left(\alpha AB[a_0 + a_1(\rho_1 S - \rho_2 k\sigma)] + \alpha BQ[a_0 + a_1(\omega_n - \rho_2 k\sigma)]\right.$$
$$\left.\cos\left[\omega_n t - \rho_1 \int_0^t S dt + \phi_n\right]\right). \tag{50}$$

The term $\alpha\gamma AB\, a_0$ represents direct current. Assuming that this is not fed back to the local oscillator we can then write

$$\sigma = A'[a_1(\rho_1 S - \rho_2 k\sigma)] + Q'[a_0 + a_1(\omega_n - \rho_2 k\sigma)] \cos \xi \tag{51}$$

[11] Arbitrary phase constants will be omitted from these expressions since they do not affect the final result.

where

$$A' = \alpha\gamma AB$$
$$Q' = \alpha\gamma BQ$$
$$\xi = \left(\omega_n t - \int_0^t \rho_1 S dt + \phi_n\right).$$

Solving for σ

$$\sigma = \frac{1}{1 + a_1 A' k\rho_2}\left[1 + \frac{a_1 Q' k\rho_2 \cos \xi}{1 + a_1 A' k\rho_2}\right]^{-1}$$
$$[A' a_1 \rho_1 S + Q'(a_0 + a_1\omega_n) \cos \xi]. \tag{52}$$

If $Q' \ll A'$

$$\sigma \doteq \frac{1}{F}\left[1 - \frac{a_1 Q' k\rho_2 \cos \xi}{F}\right]$$
$$[a_1 A' \rho_1 S + Q'(a_0 + a_1\omega_n) \cos \xi] \tag{53}$$

where $F = 1 + a_1 A' k\rho_2 = 1 - \mu\beta$ as before.

Finally, neglecting terms in Q'^2, we have

$$\sigma = \frac{1}{F}\left[a_1 A' \rho_1 S + Q'\left(a_0 + a_1\omega_n - \frac{A' a_1^2 \rho_1 k\rho_2 S}{F}\right) \cos \xi\right]. \tag{54}$$

The first term is the recovered signal while the remaining terms represent noise. Both signal and noise are modified by feedback. If we let

$$\rho_1 S = \Delta\omega \cos pt \tag{55}$$

then the noise becomes

$$\frac{Q'}{F}\left[(a_0 + a_1\omega_n) - \frac{1}{F}(A' a_1^2 k\rho_2 \Delta\omega \cos pt)\right]$$
$$\cos(\omega_n - x \sin pt + \phi_n). \tag{56}$$

By means of the Jacobi expansions this can be put in the form

$$\frac{Q'}{F}\sum_{m=-\infty}^{\infty}\left[(a_0 + a_1\omega_n) - \frac{A' a_1^2 k\rho_2 mp}{F}\right] J_m(x)$$
$$\cos[(\omega_n - mp)t + \phi_n] \tag{57}$$

where $J_m(x)$ is the Bessel coefficient of the first kind.

Now let it be assumed that the disturbance consists of a very large number of sinusoidal components of like amplitude Q, random phase, and uniformly distributed along the frequency scale. The summation of this series of voltages can be represented by the very general expression

$$f(t) \cos[\omega t + \phi(t)]. \tag{58}$$

So long as $f(t)$, the equivalent amplitude of the high-frequency disturbance, is small compared with the carrier amplitude A, the approximation (49) will be valid and the total output noise can be obtained by summing up the effects of the individual elements which constitute the disturbance.

The effect of a single disturbing element is given by (57). Any term of this expression can be made to have the frequency q if m and ω_n are so chosen that

$$\omega_n = mp \pm q. \tag{59}$$

Then for each value of m in (57) there will be available values of ω_n to satisfy both of the conditions expressed by (59). The total effect is obtained by summing the output power resulting from each contribution since the original elements have random phases. If r_2 is the resistance of the output circuit the total power of frequency q becomes

$$\frac{Q'^2}{2r_2F^2}\left(\sum_{m=-\infty}^{\infty}\left[a_0+a_1(mp+q)-\frac{A'a_1^2kp_2mp}{F}\right]^2 J_m^2(x) + \sum_{m=-\infty}^{\infty}\left[a_0+a_1(mp-q)-\frac{A'a_1^2kp_2mp}{F}\right]^2 J_m^2(x)\right). \tag{60}$$

This is readily evaluated with the aid of tables appended to an earlier paper.[12] The result is

$$\frac{Q'^2}{r_2F^2}\left[a_0^2+\frac{a_1^2\Delta\omega^2}{2F^2}+a_1^2q^2\right]. \tag{61}$$

[12] J. G. Chaffee, "The detection of frequency modulated waves," Proc. I.R.E., vol. 23, pp. 517–540; May, (1935).

The amplitude factor Q remains to be defined. If N^2 is the mean disturbing power per unit band width in the vicinity of the carrier frequency and r_1 the resistance of the input circuit, the peak amplitude of any element is defined by the relation

$$N^2d\omega = \frac{Q^2}{2r_1}. \tag{62}$$

Thus the power associated with each element becomes differentially small, and if the value so obtained is entered into (61) there is obtained the output noise power contained in a band extending from q to $q+dq$. Then we shall have

$$dW = \frac{2N^2r_1(\alpha\gamma B)^2}{r_2F^2}\left[a_0^2+\frac{a_1^2\Delta\omega^2}{2F^2}+a_1^2q^2\right]dq. \tag{63}$$

The total noise power in a band extending to a limiting frequency q_a is

$$P_n = \int_0^{q_a} dW = \frac{2N^2r_1(\alpha\gamma B)^2}{r_2F^2}\left[a_0^2+\frac{a_1^2\Delta\omega^2}{2F_2}+\frac{a_1^2q_a^2}{3}\right]q_a. \tag{64}$$

The corresponding signal power is

$$P_s = \frac{(\alpha\gamma AB)^2}{2r_2F^2}a_1^2\Delta\omega^2. \tag{65}$$

The Relation of Radio Sky-Wave Transmission to Ionosphere Measurements*

NEWBERN SMITH†, NONMEMBER, I.R.E.

Summary—*A simple, rapid, graphical method is given for obtaining maximum usable frequencies and effective reflection heights of radio waves, from vertical-incidence measurements of the critical frequencies and virtual heights of the various layers in the ionosphere. The method consists of the use of "transmission curves," which are superimposed on the curve of frequency against virtual height, observed at vertical incidence. The intersection of the curves gives the level of reflection in the ionosphere.*

The factors considered in deriving the transmission curves are variation of virtual height with frequency, effect of the curvature of the ionosphere and earth, influence of the earth's magnetic field, and absorption by or reflection from lower layers in the ionosphere. A chart is included for rapid calculation of the factor sec ϕ_0, used in plotting the transmission curves.

I. Introduction

It is possible to interpret certain aspects of long-distance radio transmission conditions in terms of measurements of the critical frequencies and virtual heights of the ionosphere made at vertical incidence. The purpose of the present paper is to outline a simple, rapid, graphical method of obtaining, from these vertical-incidence data, the maximum usable frequency over a given path, and the effective heights of reflection of waves incident obliquely upon the ionosphere. Transmission curves of the type to be described have been in use at the National Bureau of Standards since the beginning of 1936, for interpreting radio transmission data and irregularities in terms of the regular ionosphere measurements. More recently they have been used in preparing data for the weekly ionosphere bulletins broadcast over station WWV, and in supplementing local ionosphere measurements by data obtained from observations of the field intensity of distant stations. Part of this work was presented as a paper at the 1936 joint meeting of the Institute of Radio Engineers and the International Scientific Radio Union at Washington.[1]

The treatment of the propagation of radio waves given here is based entirely on the geometrical or ray theory, and considers the ionosphere as nondissipative; i.e., only the refracting strata are considered. It should be remembered that the ray approximation may be subject to a considerable correction in cases where the earth's magnetic field is not negligible. The much-discussed Lorentz polarization term[2,3] has been omitted throughout in the expression for the refractive index.

II. Equivalence Theorem for Plane Ionosphere and Earth

The process of refraction and reflection of radio waves in the ionosphere depends on the fact that the refractive index μ' of the medium decreases with increasing ionic density N. In the absence of a magnetic field, we may write the relation of μ' to N thus

$$\mu' = \sqrt{1 - \frac{f_0^2}{f'^2}} \tag{1}$$

where $f_0 = \sqrt{Ne^2/\pi m}$ and f' is the frequency of the transmitted wave.[4,5]

The propagation of electromagnetic waves in a plane ionosphere, i.e., one whose surfaces of equal ionic density are planes, is illustrated in Fig. 1. By Snell's law, the wave will penetrate the medium until the refractive index is reduced to the value sin ϕ_1, where ϕ_1 is the angle of incidence of the waves upon the ionosphere. At this level the direction of phase propagation is horizontal. This, therefore, is the highest level reached by the wave, or the level of true reflection, at a height z_0 above the lower boundary of the ionosphere. In the case of vertical incidence, reflection occurs where the refractive index is zero.

Let us define the equivalent vertical-incidence frequency f, corresponding to a wave frequency f' over a given transmission path, as the frequency of the wave reflected, at vertical incidence, at the same level as is the actual wave over the given path. It may be easily shown that, in the absence of a magnetic field

$$f = f'\sqrt{1 - \mu_0'^2} \tag{2}$$

where μ_0' is the value of μ' at the level of reflection for the wave of frequency f'.

* Decimal classification: R113.61. Original manuscript received by the Institute, June 27, 1938. Publication approved by the director of the National Bureau of Standards of the U.S. Department of Commerce.

† National Bureau of Standards, Washington, D. C.

[1] N. Smith, S. S. Kirby, and T. R. Gilliland, "Recent correlations between the ionosphere and high-frequency radio transmission." Presented before joint meeting of I.R.E. and U.R.S.I., May 1, 1936.

[2] D. R. Hartree, "Propagation of electromagnetic waves in a stratified medium," *Proc. Camb. Phil. Soc.*, vol. 25, pp. 97–120; January, (1929).

[3] C. G. Darwin, "The refractive index of an ionized medium," *Proc. Roy. Soc.*, vol. 146A, pp. 17–46; August, (1934).

[4] W. H. Eccles, "On the diurnal variations of the electric waves occurring in nature and on the propagation of electric waves around the bend of the earth," *Proc. Roy. Soc.*, vol. 87A, pp. 79–99; August 13, (1912).

[5] J. Larmor, "Why wireless electric rays can bend round the earth," *Phil. Mag.*, vol. 48, pp. 1025–1036; December, (1924).

For the plane ionosphere of Fig. 1, $\mu_0' = \sin\phi_1$ and so

$$f = \frac{f'}{\sec\phi_1},$$

the well-known "secant law." We shall denote by z_v' the height above the lower boundary of the ionosphere of the vertex of the equivalent triangular path (the triangular path having the same base and angle of departure as has the actual path), and by z_v the virtual height above this lower boundary, at vertical incidence, for the equivalent vertical-incidence frequency[6] f. Then

$$z_v' + h = \frac{1}{2} D \cos\phi_1 = \frac{\cos\phi_1}{2} \int_{\text{path}} \frac{ds \sin\phi}{\sin\phi_1}$$

where ϕ is the angle made with the vertical by the element of path ds. Since, by Snell's law, $\mu' \sin\phi = \sin\phi_1$

$$z_v' + h = \frac{\cos\phi_1}{2} \int_{\text{path}} \frac{ds}{\mu'} = \cos\phi_1 \int_0^{z_0} \frac{dz}{\mu' \cos\phi} + h.$$

By Snell's law, also,

$$\mu' \cos\phi = \sqrt{\mu'^2 - \sin^2\phi_1}. \tag{3}$$

Now

$$\mu'^2 = 1 - \frac{f_0^2}{f'^2} = 1 - \frac{f_0^2}{f^2}\cos^2\phi_1$$

so that

$$z_v' = \int_0^{z_0} \frac{dz}{\sqrt{1 - \dfrac{f_0^2}{f^2}}} = z_v. \tag{4}$$

This is the equivalence theorem for the plane earth[7] and is a direct consequence of Breit and Tuve's theorem.[8] It should be noted that ϕ_1 is equal to ϕ_0, the half-vertex angle of the equivalent triangular path.

III. Transmission Curves for Plane Ionosphere and Earth

The virtual height measured at vertical incidence for some frequency f has been shown, on the basis of the simple theory, to be the same as the height of the equivalent triangular path for a higher frequency $f' = f \sec\phi_1$, where ϕ_1 is the angle between the ray entering the ionosphere and the normal to the lower boundary of the ionosphere. This means that for transmission to take place over a given distance D at a given frequency f', the equivalent vertical-incidence frequency f must be returned, at vertical incidence, from a virtual height z_v equal to the height of the equivalent triangular path. In order to determine the z_v and f which will correspond to this transmission, it is necessary to solve simultaneously the vertical-incidence equation,

$$z_v = z_v(f), \tag{5}$$

and the transmission equation,

$$f = \frac{f'}{\sec\phi_1}, \tag{6}$$

where ϕ_1 is determined in terms of z_v by the relation $z_v = z_v'$ and by the geometry of the path. For the case of the plane earth, ϕ_1 is given by the relation

$$D = 2(z_v' + h)\tan\phi_1,$$

as may be seen from Fig. 1.

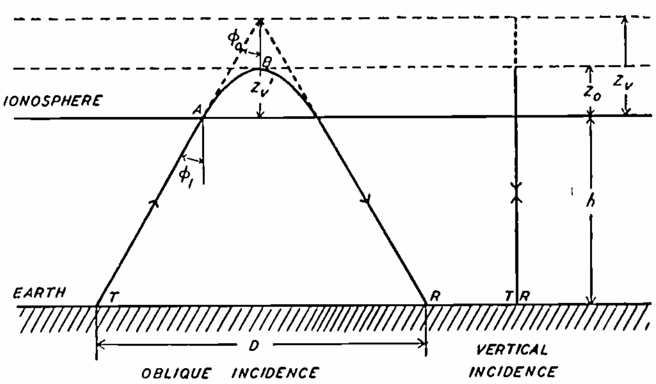


Fig. 1—Vertical-incidence and oblique-incidence transmission over a plane earth. T is the transmitter and R is the receiver. The solid curve is the actual path taken by the waves, and the dotted line shows the equivalent triangular path. z_v is the virtual height at vertical incidence for frequency f; z_0 is the true height of reflection for frequency f at vertical incidence and for frequency $f' = f/\sqrt{1-\mu_0'^2}$ at oblique incidence; z_v' is the height of the equivalent triangular path; and μ_0' is the value of the refractive index at the level z_0 for the frequency f'.

The solution of (5) and (6) is best done graphically. When (5) is plotted there is obtained a curve of virtual height against frequency observed at vertical incidence. This will be called the (z_v, f) curve. Equation (6) gives a family of curves of z_v plotted against f for different values of f' and D. These curves will be called transmission curves. The intersection of a (z_v, f) curve with a transmission curve for a given f' and D gives the height of the equivalent triangular path for transmission of the frequency f' over the distance D, and also gives the equivalent vertical-incidence frequency f for that path.

The solid curve in Fig. 2 is a typical (z_v, f) curve, showing both the E and F_2 layers in the ionosphere. Superimposed on it are the dotted transmission curves for various transmission frequencies and distances. Curves I, II, III, IV, and V are for increasing

[6] The notation on this paper differs somewhat from that in the author's previous papers, in that the primed quantities all refer to oblique-incidence and the unprimed to vertical-incidence transmission.

[7] D. F. Martyn, "The propagation of medium radio waves in the ionosphere," *Proc. Phys. Soc.*, vol. 47, pp. 323–339; March 1, (1935).

[8] G. Breit and M. A. Tuve, "A test of the existence of the conducting layer," *Phys. Rev.*, vol. 28, pp. 554–575; September, (1926).

values of f' and a fairly short distance. For the frequency corresponding to I, transmission is by way of the E layer. For curve II the given distance is within the "skip zone" for E-layer transmission but is transmitted by the F_2 layer. Curve III, which intersects the F_2 curve in two points, shows transmission by

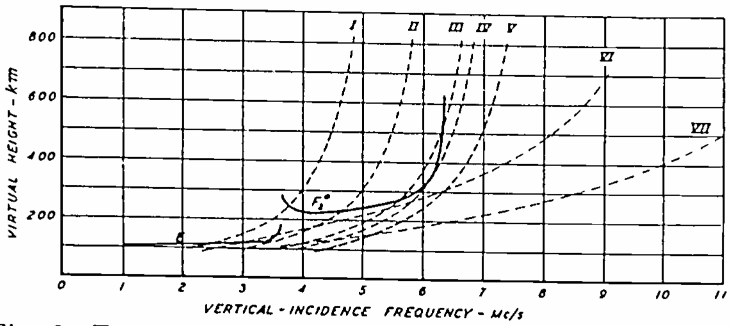


Fig. 2—Transmission curves for various frequencies and distances superimposed on a typical (z_v, f) curve for the *o* component.

Curve	D Kilometers	f' Kilocycles
I	400	5000
II	400	6000
III	400	6900
IV	400	7000
V	400	7600
VI	800	10,000
VII	800	14,000

two paths (high and low values of z_v) as reported by various observers.[9,10] Curve IV corresponds to the maximum possible frequency that can be used over the given distance. As the F_2-layer critical frequency varies, transmission has been observed to begin and fail abruptly at this point. For curve V the frequency f' is too high, and the given distance is within the skip zone. Curves VI and VII are for a longer distance, for which E-layer transmission takes place at higher frequencies than does F-layer transmission. Curve VII shows the skip effect again.

A type of transmission curve based on the principles described thus far was developed independently by L. V. Berkner of the Department of Terrestrial Magnetism, Carnegie Institution of Washington, and described by him at the joint meeting of the I.R.E. and U.R.S.I. at Washington on April 30, 1937.

IV. Logarithmic Transmission Curves

It will be noted that the transmission curve is merely a plot of $f'/\sec\phi_1$ against z_v. If the frequencies be plotted logarithmically on both the (z_v, f) and transmission curves the transmission curve is but a logarithmic $1/\sec\phi_1$ curve, with the abscissa unity at the frequency f'. It is thus possible to use this logarithmic transmission curve to represent any value of f', for any given distance D, by making the abscissa $\sec\phi_1 = 1$ coincide with the given f' on the (z_v, f) curve sheet.

The variation of f and z_v with f' over a given distance may then be determined by sliding the logarithmic $\sec\phi_1$ transmission curve along the frequency scale, and noting its intersection with the (z_v, f) curve, for every value of f'. The maximum usable frequency over the given distance is then the highest frequency for which the two curves have points in common.

An example of this type of logarithmic $\sec\phi_1$ curve is shown in Fig. 3, for the case where f' = maximum usable frequency.

Although we have neglected the magnetic field of the earth thus far in the development, it is necessary to recognize the fact that the vertical incidence (z_v, f) curve actually possesses two branches, one for the so-called *o* component or ordinary ray, and the other for the so-called *x* component, or extraordinary ray. Since the (z_v, f) curve which has been referred to above is for a wave for which $\mu' = 0$ for the equivalent vertical-incidence frequency at the level of reflection, the foregoing development and curves may be applied to the (z_v, f) curve for the *o* component,

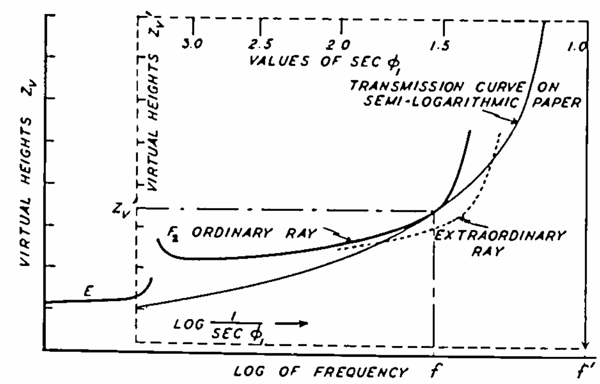


Fig. 3—Logarithmic transmission curve. The transmission curve is plotted as log $1/\sec\phi_1$ against z_v', on transparent paper. The vertical line, $\sec\phi_1 = 1$, is placed to coincide with the wave frequency f' on the semilogarithmic (z_v, f) curve. The intersection of the two curves gives the height z_v' and equivalent vertical-incidence frequency f for the transmission. In this figure f' is the maximum usable frequency for the given distance and ionosphere conditions.

for this component is reflected, at vertical incidence, at the level where $\mu' = 0$, i.e.,

$$N = \pi m f^2/e^2 \text{ at the level of reflection.}$$

V. Transmission Curves for Flat Ionosphere and Curved Earth

If the earth's curvature be not neglected, but the ionosphere be considered flat, the above developments and equivalence theorem are still valid for the part of the wave's path which lies in the ionosphere.

[9] F. T. Farmer and J. A. Ratcliffe, "Wireless waves reflected from the ionosphere at oblique incidence," *Proc. Phys. Soc.*, vol. 48, pp. 839–849; November, (1936).

[10] W. Crone, K. Krüger, G. Goubau, and J. Zenneck, "Echo measurements over long distances," *Hochfrequenz. und Elektroakustic*, vol. 48 pp. 1–7; July, (1936).

Such a case is illustrated in Fig. 4. This is the case where the wave does not travel very far, horizontally, in a part of the ionosphere where μ' differs appreciably from unity, so that $z_v' - z_0$ is very small compared with $(R+h)/\tan^2\phi_1$. It should be noted that, to the first order, ϕ_1 still equals ϕ_0 and $z_v = z_v'$.

In this case the distance of transmission is connected with the angle of incidence by the relation

$$\tan \phi_1 = \frac{\sin \dfrac{D}{2R}}{\dfrac{z_v + h}{R} + 1 - \cos \dfrac{D}{2R}} \tag{7}$$

and this value of ϕ_1 is the one to be used in the transmission equation (6).

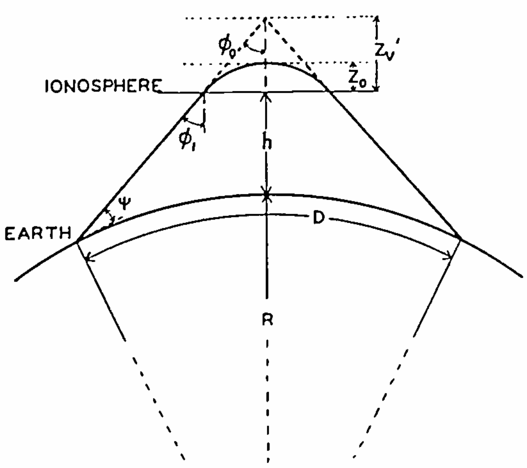


Fig. 4—Geometric relation for determining ϕ_1 in terms of transmission distance D for a curved earth and flat ionosphere. In this and in Figs. 5 and 6 the distance between the ionosphere and the earth is greatly exaggerated for clarity of representation.

The transmission curves can still be plotted logarithmically and applied to the observed (z_v, f) curves to obtain values of z_v' for the oblique transmission, good to the first order of approximation.

VI. Equivalence Relations for Curved Ionosphere and Earth

To obtain the variation, with μ', of ϕ, the angle made by an element of the ray path with the vertical, we must consider first the angle a straight line makes with the vertical at various heights above the earth's surface. The geometry of Fig. 5 leads to the relation

$$\sin \phi' = \sin \phi \left(1 + \frac{dz}{R + h + z}\right). \tag{8}$$

Assuming the validity of Snell's law for a ray traversing an infinitely thin layer of the ionosphere, of thickness dz, we can use (8) to obtain the differential equation

$$\frac{d(\mu' \sin \phi)}{\mu' \sin \phi} = -\frac{dz}{R + h + z}.$$

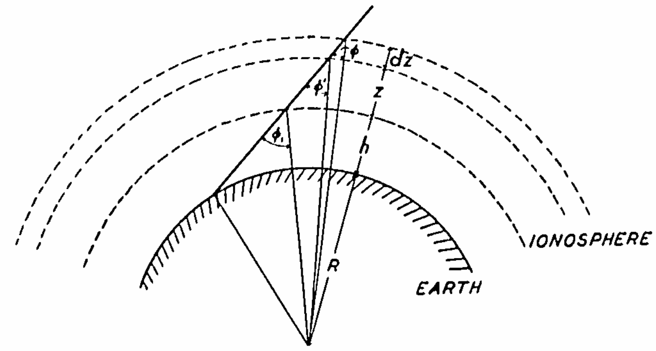


Fig. 5—Variation, with height, of the angle that a straight line makes with the normal to the earth's surface.

Integrating this from the lower boundary of the ionosphere, where $\mu' = 1$, $\phi = \phi_1$, and $z = 0$ (see Fig. 6) up to the level z, we get

$$\mu' \sin \phi = \frac{\sin \phi_1}{1 + \dfrac{z}{R + h}} \tag{9}$$

as the form of Snell's law appropriate to the curved ionosphere.

This means that the wave will penetrate the curved ionosphere until the refractive index is reduced to the value

$$\mu_0' = \frac{\sin \phi_1}{1 + \dfrac{z_0}{R + h}} \tag{10}$$

where ϕ_1 is the angle of incidence of the wave on the lower boundary of the ionosphere, and z_0 is the maximum height of penetration above this lower boundary.

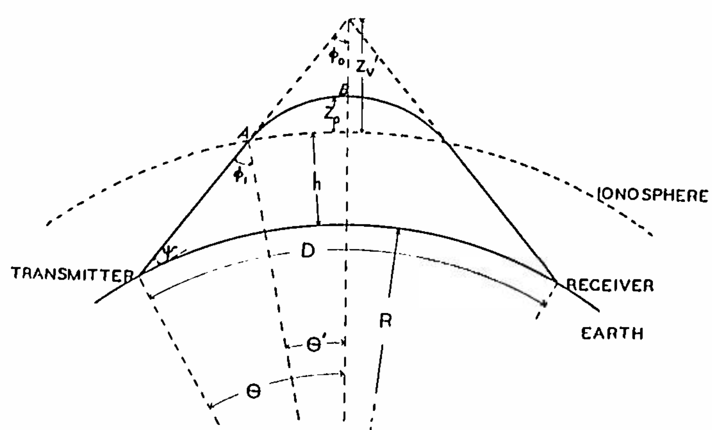


Fig. 6—Transmission through curved ionosphere. z_v' = height of equivalent triangular path, z_0 = true height of reflection, at the level where $\mu' = \sin \phi_1(1 - z_0/(R+h))$, R = radius of earth, D = distance of transmission.

If we express this in terms of ϕ_0, the half-vertex angle of the equivalent triangular path, (10) becomes

$$\mu_0' = \sin\phi_0 \frac{1 + \dfrac{z_v'}{R+h}}{1 + \dfrac{z_0}{R+h}} \tag{10a}$$

or, if $z_0 \ll R+h$, as is usually the case,

$$\mu_0' = \sin\phi_0\left(1 + \frac{z_v' - z_0}{R+h}\right) \tag{10b}$$

instead of the relation $\mu_0' = \sin\phi_0 = \sin\phi_1$, which was true for the flat ionosphere. This means that the level of reflection is lower than would be the case were the ionosphere flat, and a given maximum ionic density will permit a higher frequency to be propagated over a given distance.

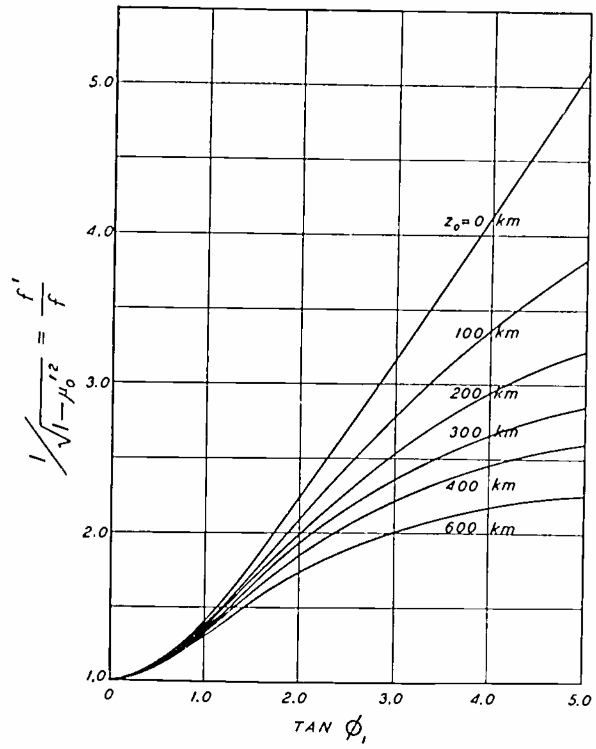


Fig. 7—Variation of $1/\sqrt{1-\mu_0'^2}$ with z_0 and $\tan\phi_1$.

The equivalent vertical-incidence frequency in this case is

$$f = f'\sqrt{1-\mu_0'^2} = f'\sqrt{1 - \frac{\sin^2\phi_1}{\left(1 + \dfrac{z_0}{R+h}\right)^2}}$$

$$= f'\sqrt{1 - \sin^2\phi_0\left[\frac{1 + \dfrac{z_v'}{R+h}}{1 + \dfrac{z_0}{R+h}}\right]^2}. \tag{11}$$

Under practical conditions z_0 is almost always less than 400 kilometers, and values of z_v' greater than 400 kilometers contribute nothing to transmission over appreciable distances, so that we can always consider z_0 and $z_v' \ll R+h$. Using this approximation (11) becomes

$$f = f'\cos\phi_1\sqrt{1 + \frac{2z_0}{R+h}\tan^2\phi_1}$$

$$= f'\cos\phi_0\sqrt{1 - \frac{2(z_v' - z_0)}{R+h}\tan^2\phi_0}. \tag{11a}$$

Fig. 7 gives values of $1/\sqrt{1-\mu_0'^2}$ plotted against $\tan\phi_1$ for various values of z_0, assuming $h = 100$ kilometers (at the bottom of the E layer). These curves may be used to determine the relation of f to f' for various values of z_0 and ϕ_1. By substituting ϕ_0 for ϕ_1 and $z_0 - z_v'$ for z_0, similar curves may also be plotted to determine the relation of f to f' for various values of z_0, z_v', and ϕ_0.

If, further, $z_0 \ll (R+h)\cot^2\phi_1$ or $(z_v' - z_0) \ll (R+h)\cot^2\phi_0$ we may write (11a) approximately,

$$f = f'\sqrt{1 - \mu_0'^2} = f'\cos\phi_1\left[1 + \frac{z_0}{R+h}\tan^2\phi_1\right]$$

$$= f'\cos\phi_0\left[1 - \frac{z_v' - z_0}{R+h}\tan^2\phi_0\right] \tag{11b}$$

a form which it is convenient to use in some discussions. This approximation leads to results good to 1 per cent or better for E-layer transmission, where z_0 is less than 50 kilometers, and for single-reflection F-layer transmission over distances less than 1500 kilometers, or multireflection F-layer transmission where each reflection covers less than 1500 kilometers. For transmission over distances greater than 1500 kilometers for each reflection it is necessary to use the more exact expression.

The correction term, or frequency Δf which it is necessary to subtract from $f'/\sec\phi_0$ in order to obtain f, is the amount by which the equivalent vertical-incidence frequency for the curved ionosphere differs from that for the flat ionosphere. This is

$$\Delta f = f'\left[\cos\phi_0 - \sqrt{1 - \mu_0'^2}\right]$$

which is approximately

$$\Delta f = f'\cos\phi_0\left[1 - \sqrt{1 - 2\frac{z_v' - z_0}{R+h}\tan^2\phi_0}\right]$$

and, if $z_v' - z_0 \ll (R+h)\cot^2\phi_0$

$$\Delta f \doteq f'\left(\frac{z_v' - z_0}{R+h}\right)\frac{\sin^2\phi_0}{\cos\phi_0}.$$

This correction term is zero at vertical incidence and for $z_0 = z_v$, i.e., reflection from a sharp boundary at the height z_0. For $z_0 \neq z_v$ it increases rapidly with distance.

The vertical-incidence frequency f derived above may be used to plot transmission curves of z_v' against f, exactly as in the case of the flat ionosphere. If the assumption be made that z_v' is the same as the z_v measured at the frequency f, then these transmission curves could be superimposed on the (z_v, f) curve, to give directly the z_v' for the oblique-incidence case.

Actually it is more convenient to use the sec ϕ_0 transmission curves, wherein z_v' is plotted against $f'/\sec \phi_0$, and to refer the correction to the (z_v, f) curve. On the assumption that $z_v' = z_v$, Δf can be calculated, for each point on the (z_v, f) curve, for each sec ϕ_0 transmission curve. This transmission curve can then be superimposed on the (z_v, f) curve displaced toward the higher frequencies by the amount Δf, and the oblique-incidence values of z_v' read off.

The calculation of Δf for each distance from z_v and z_0, without any assumption as to z_v', and for a ϕ_0 calculated for an equivalent triangular path of height z_v, is discussed below. This value of Δf can be used in the same manner as was described in the preceding paragraph, and the sec ϕ_0 curves applied, ϕ_0 being calculated for an equivalent triangular path of height z_v instead of z_v'.

If the (z_v, f) curve and the sec ϕ_0 curves are plotted logarithmically, we may calculate a factor $1 + (\Delta f/f)$ as follows:

$$1 + \frac{\Delta f}{f} = \frac{\cos \phi_0}{\sqrt{1 - \mu_0'^2}}.$$

The frequencies on the (z_v, f) curve then can be easily multiplied by this factor by adding the factor logarithmically to the curve, and the log sec ϕ_0 curves can be applied to the resulting corrected (z_v, f) curve. It should be noted that in this expression the ϕ_0 refers only to an equivalent triangular path of height z_v, and not z_v'. The z_v' enters only into the expression for μ_0'.

A typical (z_v, f) curve corrected in this manner for a given distance (2000 kilometers) is shown in Fig. 8. The virtual heights are lower on the corrected curve and the curve extends out to frequencies higher than the critical frequency for the ordinary ray. The correction to the curve is different for different distances, increasing with the distance. This means that the virtual heights of reflection are lower, for greater distances, than if the curvature of the earth were not considered. The limiting frequency of transmission, or maximum usable frequency, over considerable distances is correspondingly increased, the difference becoming as much as 20 per cent at the greater distances, depending of course on the form of the (z_v, f) curve.

When the (z_v, f) curve has been corrected in the above manner it may be plotted logarithmically as described above, and the logarithmic sec ϕ_0 curves may be applied to give directly the maximum usable frequencies and virtual heights over transmission paths of given lengths.

Since the first draft of this paper was written there has come to the author's attention an excellent unpublished paper "Skip Distance Analysis," by T. L. Eckersley and G. Millington, in the form of a contribution to the November, 1937, London meeting of the Special Radio Wave Propagation Committee held in preparation for the Cairo Radio Conference.[11] In this they undertake an analysis of radio transmission over a curved earth and obtain curves for determining the retardation of sky wave over ground wave at a distance, in terms of vertical-incidence

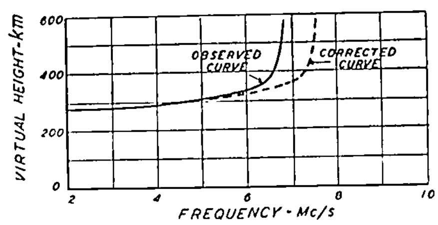


Fig. 8—(z_v, f) curve corrected for a given distance (2000 kilometers) for the effect of the earth's curvature.

measurements. They limit the analysis, however, to the case where z_0 is very small compared with h. The analysis can thus apply only to E-layer transmission, since for F-layer transmission z_0 must be measured from the lower boundary of the ionosphere in order to include the effect of retardation in the E layer.

Their work, however, suggests a method of approximate analysis for larger values of z_0 and a means for easily evaluating the correction factor $1 + \Delta f/f$ without making the assumption that z_v is equal to z_v'.

The development given by Eckersley and Millington begins with the following relation, for the ray path shown in Fig. 6

$$\theta' = \int_A^B d\theta = \int_A^B \frac{ds \sin \phi}{R + h + z}$$

where ds is an element of the ray path making an angle ϕ with the vertical.

Since, from (9)

$$\sin \phi = \frac{1}{\mu'} \frac{\sin \phi_1}{1 + \dfrac{z}{R + h}}$$

[11] *Added in proof:* G. Millington, "The relation between ionospheric transmission phenomena at oblique incidence and those at vertical incidence," *Proc. Phys. Soc.*, vol. 50, pp. 801–825; September 1, (1938).

this becomes

$$\theta' = \frac{\sin\phi_1}{R+h}\int_A^B \frac{ds}{\mu'\left(1+\frac{z}{R+h}\right)^2}. \quad (12)$$

They also used the relation

$$\int_A^B \frac{ds}{\mu'} = \int_0^{z_0} \frac{dz}{\mu'\cos\phi}$$
$$= \int_0^{z_0} \frac{dz}{\sqrt{\mu'^2 - \frac{\sin^2\phi_1}{\left(1+\frac{z}{R+h}\right)^2}}}. \quad (13)$$

For the case where there is no magnetic field

$$\mu' = \sqrt{1-\frac{f_0^2}{f'^2}}, \quad \text{where } f_0 = \sqrt{\frac{Ne^2}{\pi m}}, \quad \text{and so}$$

$$\int_A^B \frac{ds}{\mu'} = \int_0^{z_0} \frac{f'dz}{\sqrt{f'^2 - f_0^2 - f'^2\frac{\sin\phi^2{}_1}{\left(1+\frac{z}{R+h}\right)^2}}}. \quad (14)$$

Putting in this expression the wave frequency f' in terms of the equivalent vertical-incidence frequency given above in (11), (11a), and (11b), and using the value of μ_0' derived in (10),

$$\int_A^B \frac{ds}{\mu'} = \frac{1}{\sqrt{1-\mu_0'^2}}$$
$$\int_0^{z_0} \frac{fdz}{\sqrt{\frac{f^2}{1-\mu_0'^2}\left(1-\frac{\sin^2\phi_1}{\left(1+\frac{z}{R+h}\right)^2}\right)-f_0^2}}. \quad (15)$$

At this point Eckersley and Millington assumed that they were dealing with a thin layer, and that in consequence (1) θ' was only slightly less than

$$\frac{\sin\phi_1}{R+h}\int_A^B \frac{ds}{\mu'},$$

from equation (12), (2)

$$\int_A^B \frac{ds}{\mu'}$$

was only slightly greater than

$$\frac{1}{\sqrt{1-\mu_0'^2}}\int_0^{z_0} \frac{fdz}{\sqrt{f^2-f_0^2}},$$

from equation (15).

This simplification is justified for E-layer transmission, but is unfortunately not justified for F_2-layer transmission, since the lower boundary of the ionosphere must be taken at the beginning of the E layer in each case. We shall not, therefore, confine our discussion to this limited case. It must also be noted that the simplification introduced by assuming $z_0 \ll (R+h)\cot^2\phi_1$ also is not valid except at comparatively short distances, where $\phi_1 = 45$ degrees or less. For practical purposes, it is, however, justifiable to assume $z_0 \ll R+h$. Using this relation then, we obtain

$$\theta' = \frac{\sin\phi_1}{R+h}\int_A^B \frac{ds}{\mu'}\left(1-\frac{2z}{R+h}\right). \quad (16)$$

Eckersley and Millington combined the approximate forms of (12) and (15) and introduced the value of $\sqrt{1-\mu_0^2} = \cos\phi_1$, to obtain

$$\theta' = \frac{\tan\phi_1}{R+h}\int_0^{z_0} \frac{fdz}{\sqrt{f^2-f_0^2}}.$$

We shall, however, combine (15) and (16) to obtain the more exact relation

$$\theta' = \frac{\sin\phi_1}{(R+h)\sqrt{1-\mu_0'^2}}$$
$$\int_0^{z_0} \frac{fdz}{\sqrt{f^2(1-A)-f_0^2}}\left(1-\frac{2z}{R+h}\right) \quad (16a)$$

where

$$1-A = \frac{1-\frac{\sin^2\phi_1}{\left(1+\frac{z}{R+h}\right)^2}}{1-\mu_0'^2} \doteq \frac{1+\frac{2z}{R+h}\tan^2\phi_1}{1+\frac{2z_0}{R+h}\tan^2\phi_1}. \quad (16b)$$

If we let

$$1+B = \frac{1}{\sqrt{1-A\left(\frac{f^2}{f^2-f_0^2}\right)}}, \quad (16c)$$

$$\theta' = \frac{\sin\phi_1}{(R+h)\sqrt{1-\mu_0'^2}}$$
$$\int_0^{z_0} \frac{dz}{\sqrt{1-\frac{f_0^2}{f^2}}}(1+B)\left(1-\frac{2z}{R+h}\right). \quad (17)$$

For values of z_0 considerably less than $(R+h)/2\cot^2\phi_1$, B is usually a small number, and for $z \to z_0$ (near the level of reflection) it is also usually small. It may attain, however, relatively large values.

In general the evaluation of B involves a knowledge of the distribution of ionic density, and a precise evaluation of the quantity θ' involves graphical integration for each case considered. We shall for the present consider two principal cases, (a) $z_0 \rightarrow 0$ and (b) $z_0 = z_v$, the virtual height at vertical incidence for the frequency $f = f'\sqrt{1-\mu_0'^2}$, and discuss some cases of linear distributions of ionic density, where $z/z_0 = f_0^2/f^2$.

(a) $z_0 \ll R+h \cot^2 \phi_1$.

This case applies to E-layer transmission at any distance or to F-layer transmission at short distances (ϕ_1 small).

From (11b) we have

$$f = f' \cos \phi_1 \left[1 + \frac{z_0}{R+h} \tan^2 \phi_1\right], \qquad (18a)$$

and from (16b)

$$1 - A = 1 + 2\left(\frac{z - z_0}{R+h}\right)\tan^2 \phi_1.$$

From (16c)

$$1 + B = 1 + \left(\frac{z_0}{R+h}\right)\frac{1 - \frac{z}{z_0}}{1 - \frac{f_0^2}{f^2}} \tan^2 \phi_1,$$

if we assume that

$$\left(1 - \frac{z}{z_0}\right) \Big/ \left(1 - \frac{f_0^2}{f^2}\right)$$

is nowhere so large as to make B comparable with unity. This is equivalent to assuming that f_0^2/f^2 does not approach 1 much more rapidly than does z/z_0, an assumption that is valid except quite close to a critical frequency. Transmissions involving equivalent vertical-incidence frequencies close to the critical frequency are of interest, however, only when the distance of transmission is short, and in this case ϕ_1 is small, so that B is a small number, anyhow.

Thus

$$\theta' = \frac{\tan \phi_1}{R+h}\left[1 - \frac{z_0}{R+h}\tan^2 \phi_1\right]\int_0^{z_0} \frac{dz}{\sqrt{1 - \frac{f_0^2}{f^2}}}$$

$$\left\{1 + \frac{z_0}{R+h}\tan^2 \phi_1 \frac{\left(1 - \frac{z}{z_0}\right)}{\left(1 - \frac{f_0^2}{f^2}\right)} - \frac{2z}{R+h}\right\}.$$

Now

$$\int_0^{z_0} \frac{dz}{\sqrt{1 - \frac{f_0^2}{f^2}}} = z_v,$$

the virtual height measured at vertical incidence for the equivalent vertical-incidence frequency f. Since the terms involving $z_0/(R+h)$ are small, we may write

$$\theta' = \frac{z_v \tan \phi_1}{R+h}\left\{1 - \frac{z_0}{R+h}\tan^2 \phi_1(1 - C) - C'\right\} \qquad (18b)$$

where

$$C = \frac{1}{z_v}\int_0^{z_0} \frac{dz}{\sqrt{1 - \frac{f_0^2}{f^2}}} \; \frac{1 - \frac{z}{z_0}}{1 - \frac{f_0^2}{f^2}}$$

and

$$C' = \frac{2}{z_v(R+h)}\int_0^{z_0} \frac{z\,dz}{\sqrt{1 - \frac{f_0^2}{f^2}}}.$$

For a linear distribution of ionic density

$$1 - \frac{z}{z_0} = 1 - \frac{f_0^2}{f^2}, \quad z_v = 2z_0, \quad \text{and } C = 1,$$

and

$$\theta' = \frac{\tan \phi_1}{R+h}\int_0^{z_0} \frac{dz}{\sqrt{1 - \frac{z}{z_0}}}\left[1 - \frac{2z}{R+h}\right]$$

$$= \frac{z_v \tan \phi_1}{R+h}\left[1 - \frac{2}{3}\left(\frac{z_v}{R+h}\right)\right]. \qquad (19)$$

For z_0 vanishingly small $C' \rightarrow 0$ and (18) reduces simply to

$$\theta' = \frac{z_v \tan \phi_1}{R+h} \qquad (19a)$$

(b) $z_0 \rightarrow z_v$.

This case applies to reflection from a fairly sharp boundary at the level z_0, the refractive index being nearly unity up to nearly this level. An example of this would be reflection from the sporadic-E region. The angular distance of the part of the ray path where μ departs appreciably from unity is small and so the ionosphere can be considered as essentially flat. This case was treated in section V. The results will be summarized here for completeness. Here $z_v = z_v'$ and

$$\tan \phi_0 = \frac{\sin \theta'}{\dfrac{z_v}{R+h} + 1 - \cos \theta'}. \tag{20}$$

In this case

$$\sqrt{1 - \mu_0'^2} = \cos \phi_0 \left(1 - \frac{z_v - z_0}{R+h} \tan^2 \phi_0\right). \tag{20a}$$

For $z_v - z_0$ vanishingly small compared with $(R+h) \cot^2 \phi_0$, this reduces to the case for the plane ionosphere, where $\sqrt{1-\mu_0'^2} = \cos \phi_0$.

(c) $z_0 = \frac{1}{2} z_v$, i.e., a linear distribution of ionic density.

Here $1 - f_0^2/f^2 = 1 - z/z_0$ and (16a) becomes

$$\theta' = \frac{\sin \phi_1}{(R+h)\sqrt{1-\mu_0'^2}} \int_0^{z_0} \frac{dz \left(1 - \dfrac{2z}{R+h}\right)}{\sqrt{1 - A - \dfrac{z}{z_0}}}.$$

Putting in the value of $1-A$ and integrating

$$\theta' = z_v \frac{\sin \phi_1 \sqrt{1 + \dfrac{2z_0}{R+h} \tan^2 \phi_1}}{(R+h)\sqrt{1-\mu_0'^2}} \left(1 - \frac{2}{3} \frac{z_v}{R+h}\right). \tag{21}$$

Now since we may consider $z_0 \ll R+h$ we can write, as in (11b),

$$\sqrt{1-\mu_0'^2} = \cos \phi_1 \sqrt{1 + \frac{2z_0}{R+h} \tan^2 \phi_1}$$

and thus (21) becomes

$$\theta' = \frac{z_v \tan \phi_1}{R+h} \left[1 - \frac{2}{3} \frac{z_v}{R+h}\right] \tag{19}$$

just as in the case where $z_0 \ll (R+h) \cot^2 \phi_1$. We must, however, write for the equivalent vertical-incidence frequency in this case

$$f = f' \cos \phi_1 \sqrt{1 + \frac{2z_0}{R+h} \tan^2 \phi_1} \tag{21a}$$

instead of the value

$$f = f' \cos \phi_1 \left[1 + \frac{z_0}{R+h} \tan^2 \phi_1\right] \tag{18a}$$

which we could use when $z_0 \ll (R+h) \cot^2 \phi_1$.

The three examples just treated give an idea of what happens on transmission through the ionosphere, and of the angular distance θ' the wave travels in the ionosphere as a function of angle of incidence ϕ_1 (or vertex angle of equivalent triangular path ϕ_0), of true height of reflection z_0, and of virtual height measured at vertical incidence z_v. We must now consider the part of the path from the earth to the lower boundary of the ionosphere. If we consider the incident ray to traverse an angular distance $(\theta - \theta')$ in going from the earth to the lower boundary of the ionosphere, the geometry of Fig. 6 tells us that

$$\tan \phi_1 = \frac{\sin (\theta - \theta')}{\dfrac{h}{R} + 1 - \cos (\theta - \theta')}. \tag{22}$$

For a given ϕ_1, then, we may solve this equation for $(\theta - \theta')$, and, by adding this angle to the angle θ' already computed, we obtain the entire angular distance θ traversed by the wave from the ground to the point of reflection in the ionosphere. For $h = 100$ kilometers, as we are assuming, it is sufficiently accurate to replace $\sin (\theta - \theta')$ by $(\theta - \theta')$, and $1 - \cos (\theta - \theta')$ by $1/2\ (\theta - \theta')^2$ in this equation. We can thus solve for $(\theta - \theta')$, obtaining

$$\theta - \theta' = \cot \phi_1 - \sqrt{\cot^2 \phi_1 - \frac{2h}{R}}. \tag{22a}$$

We can now write the expression for the total distance of transmission D in terms of ϕ_1 (or ϕ_0) z_0 and z_v'

$$D = 2R \left(\cot \phi_1 - \sqrt{\cot^2 \phi_1 - \frac{2h}{R}} + \theta'\right) \tag{23}$$

where θ' is computed as above, from (16a).

The time T required for the sky wave to travel from the transmitter to the receiver is, if c = velocity of the wave in vacuum

$$T = \frac{2}{c} \left(R \frac{\sin (\theta - \theta')}{\sin \phi_1} + \int_A^B \frac{ds}{\mu}\right).$$

If we put in the value of $\int_A^B ds/\mu$ obtained on the basis of the above analysis, and replace $\sin (\theta - \theta')$ by its value in terms of ϕ_1, we may express T in terms of the quantities ϕ_1 (or ϕ_0), z_0, and z_v, as was done with D.

The expression for the height z_v' of the equivalent triangular path can now be written. From the geometry of Fig. 6,

$$\tan \phi_0 = \tan (\phi_1 - \theta') = \frac{\sin \theta'}{\dfrac{z_v'}{R+h} + 1 - \cos \theta'}$$

so that

$$z_v' = (R+h)[\sin \theta' \cot (\phi_1 - \theta') - 1 + \cos \theta']. \tag{24}$$

The relation between z_v, z_v', z_0, ϕ_1, ϕ_0, θ', D, f', and f may be summarized for the cases discussed here.

(a) $z_0 \ll 1/2(R+h)\cot^2\phi_1$.

$$f = f'\cos\phi_1\left[1+\frac{z_0}{R+h}\tan^2\phi_1\right]$$

$$\theta' = \frac{\tan\phi_1}{R+h}\left[1-\frac{z_0}{R+h}\tan^2\phi_1\right]\int_0^{z_0}\frac{dz}{\sqrt{1-\frac{f_0^2}{f^2}}}\left\{1+\frac{z_0}{R+h}\tan^2\phi_1\left[\frac{1-\frac{z}{z_0}}{1-\frac{f_0^2}{f^2}}\right]-\frac{2z}{R+h}\right\}$$

$$T = \frac{2}{c}\left[\frac{R(\theta-\theta')}{\sin\phi_1}+\frac{1}{\cos\phi_1}\int_0^{z_0}\frac{dz}{\sqrt{1-\frac{f_0^2}{f^2}}}\left\{1-\frac{z_0}{R+h}\tan^2\phi_1\left[1-\frac{1-\frac{z}{z_0}}{1-\frac{f_0^2}{f^2}}\right]\right\}\right]$$

$$\theta-\theta' = \cot\phi_1-\sqrt{\cot^2\phi_1-\frac{2h}{R}}$$

$$D = 2R\theta$$

$$z_v' = (R+h)[\sin\theta'\cot(\phi_1-\theta')-1+\cos\theta']$$

$$z_v = \int_0^{z_0}\frac{dz}{\sqrt{1-\frac{f_0^2}{f^2}}}$$

(a′) $z_0=0$.

$$f = f'\cos\phi_1$$

$$\theta' = \frac{z_v\tan\phi_1}{R+h}$$

$$T = \frac{2}{c}\left[\frac{R(\theta-\theta')}{\sin\phi_1}+\frac{z_v}{\cos\phi_1}\right]$$

$$\theta-\theta' = \cot\phi_1-\sqrt{\cot^2\phi_1-\frac{2h}{R}}$$

$$D = 2R\theta$$

$$z_v' = (R+h)[\sin\theta'\cot(\phi_1-\theta')-1+\cos\theta']$$

(b) linear gradient of ionic density ($z_0=1/2\ z_v$).

$$f = f'\cos\phi_1\sqrt{1+\frac{z_v}{R+h}\tan^2\phi_1}$$

$$\theta' = \frac{z_v\tan\phi_1}{R+h}\left[1-\frac{2}{3}\frac{z_v}{R+h}\right]$$

$$T = \frac{2}{c}\left[\frac{R(\theta-\theta')}{\sin\phi_1}+\frac{z_v}{R+h}\right]$$

$$\theta-\theta' = \cot\phi_1-\sqrt{\cot^2\phi_1-\frac{2h}{R}}$$

$$D = 2R\theta$$

$$z_v' = (R+h)[\sin\theta'\cot(\phi_1-\theta')-1+\cos\theta']$$

(c) $z_0=z_v$.

$$f = f'\cos\phi_0$$

$$\tan\phi_0 = \frac{\sin\theta'}{\frac{z_v}{R+h}+1-\cos\theta'} = \frac{\sin\theta}{\frac{z_v+h}{R}+1-\cos\theta}$$

$$T = \frac{2}{c}\frac{(R+h)\sin\theta}{\sin\phi_0}$$

$$D = 2R\theta$$

$$z_v' = z_v.$$

Case (a′) corresponds to the special case treated by Eckersley and Millington, except that, for simplicity of analytical computation, they took $R+h=R$ in the expression for θ', which introduces an error of about $0.016h$ per cent (h expressed in kilometers, not a serious error for $h=100$ kilometers, as we have assumed. This error becomes appreciable, however, if it is attempted to extend this treatment, as they have done, to a height of 400 kilometers or so.

Martyn's equivalence theorem, discussed above, tells us that for a plane ionosphere

$$z_v' = z_v$$

$$T = \frac{2}{c}\frac{z_v}{\cos\phi_0} = \frac{2}{c}\frac{z_v}{\cos\phi_1}$$

$$D' = 2z_v\tan\phi_0 = 2z_v\tan\phi_1$$

$$f = f'\cos\phi_0 = f'\cos\phi_1$$

where

- ϕ_1 = angle of incidence of waves upon the ionosphere
- ϕ_0 = half the vertex angle of the equivalent triangular path
- z_v' = height of equivalent triangular path
- T = time the wave spends in the ionosphere
- D' = horizontal distance the wave travels in the ionosphere.

We see that the relation $z_v=z_v'$ is valid in case (c) but not in any other case. The relation $T=(2/c)(z_v/\cos\phi_1)$ is valid in cases (a′) and (b), but not otherwise in general. The relation $D'=2z_v\tan\phi_1$ holds approximately only in case (a′). And finally

$f = f' \cos\phi_0$ in case (c), $f = f' \cos\phi_1$ in case (a′), and f is a complicated function in every other case. It is therefore concluded that the equivalence theorem, in the form given, cannot be applied to the curved-earth problem.

VII. Transmission Curves for Curved Ionosphere and Earth

Referring again to the summaries of cases (a′), (b), and (c), we may for a given value of z_v take a set of values of z_0, for each of which can be calculated the variation of D with ϕ_1, in each of the three cases. We may also, for these values of z_v and z_0, calculate the variation of f'/f with ϕ_1. By eliminating ϕ_1 graphically, we can determine the variation of f'/f with D for a given z_v and z_0, or, what is the same thing, the variation of f'/f with z_v for a given D and z_0. We can plot a family of transmission curves, with f as abscissas and z_v as ordinates, for each value of D and several values of z_0, corresponding to each of these cases.

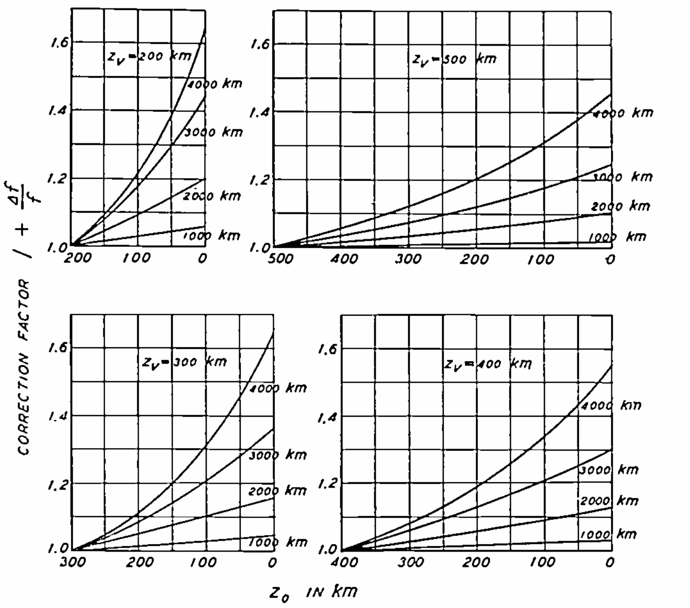


Fig. 9—Approximate factors by which the equivalent vertical-incidence frequency f must be multiplied to give the frequency $f'/\sec\phi_0$ which is to be used in conjunction with the log sec ϕ_0 curves.

Such curves may be superimposed on the (z_v, f) curve obtained at vertical incidence, and the (z_0, f) curve deduced therefrom, to give directly the wave frequency f' corresponding to reflection at a given level z_0 which is characterized by a given z_v. The value of z_0 appropriate to reflection at the given height determines which curve of the family for a given D is to be used, and the point of reflection is determined as the intersection of this transmission curve with the (z_v, f) curve.

The value of z_v' depends only on D and ϕ_1, and so lines of equal z_v' may be plotted on the curve sheet for each D, so that z_v', as well as f' may be read off directly.

Because of the approximations made in case (b) and the assumption of a linear variation of ionic density with height this case is of only special significance. It will be assumed, until further investigation determines more precisely the variation of conditions with z_0, that the curves vary smoothly between those calculated for $z_0 = 0$ and those calculated for $z_0 + z_v$.

A family of curves for each distance is rather cumbersome for rapid use. It is, as was said above, more convenient to use the log sec ϕ_0 transmission curves, and apply a correction to the (z_v, f) curve by multiplying each vertical-incidence frequency by the factor $1 + \Delta f/f$, where

$$1 + \frac{\Delta f}{f} = \frac{\cos\phi_0}{\sqrt{1 - \mu_0'^2}} \cdot$$

This factor is obtained, for a given D, z_v, and z_0, by determining corresponding values of D and $\sqrt{1 - \mu_0'^2}$ for arbitrary values of ϕ_0 or ϕ_1. It is unity for $z_v = z_0$ and is quite easily obtained for $z_0 = 0$. For intermediate values of z_0 it will be assumed that the factor $1 + \Delta f/f$ varies in a manner similar to that determined from the relation (7b) with $z_v' = z_v$; i.e.,

$$\frac{\cos\phi_0}{\sqrt{1 - \mu_0'^2}} = \sqrt{1 - \frac{2(z_v' - z_0)}{R + h} \tan^2\phi_0}$$

but drawn through the values for $z_0 = 0$ and $z_0 = z_v$ determined in the more precise analysis, rather than those indicated on the assumption that $z_v = z_v'$. Cos ϕ_0 is here calculated for an equivalent triangular path of height z_v.

Fig. 9 gives the approximate factors by which f must be multiplied to give $f'/\sec\phi_0$ for values of z_v from 200 to 500 kilometers and for distances up to 4000 kilometers.

VIII. Effect of the Earth's Magnetic Field

The presence of the earth's magnetic field introduces some complications in the use of these transmission curves. These complications are often of minor importance compared with some of the unknown factors (e.g., the geographic uniformity of the ionosphere over the transmission path), especially over long distances, but the effect of the earth's field must at times be taken into account. The anisotropy of the ionosphere due to this field causes the effect of the field on radio transmission to vary with the length, direction, and geographic location of the transmission path.

One effect of the field is to cause the received signal to be split in general into two main components, the one with the lower maximum usable frequency known as the o component and the other as the x component. The refractive index for a frequency f, in the presence of a magnetic field H, whose components along and transverse to the direction of phase propagation are, respectively, H_L and H_T, is given[12] by

$$\mu' = 1 - \frac{f_0^2}{f'^2 - \frac{f'^2 f_T^2}{2(f'^2 - f_0^2)} \pm \sqrt{\left(\frac{f'^2 f_T^2}{2(f'^2 - f_0^2)}\right)^2 + f'^2 f_L^2}}$$

where

$$f_0 = \sqrt{\frac{Nc^2}{\pi m}}$$

$$f_T = \frac{e}{2\pi mc} H_T$$

$$f_L = \frac{e}{2\pi mc} H_L$$

N = ionization density.

The upper sign refers to the o component; the lower to the x component. The frequency of a wave whose x component is returned from a given ionization density, at a given height, is different from the frequency of the wave whose o component is returned from the same level. This frequency separation is in general a function both of frequency and distance, and may often become negligible at great distances.

For practical calculation it may be assumed that only the field and direction of wave propagation in the region near the level of reflection will appreciably affect the propagation of the wave. This assumption is probably better for the o than for the x component, and is, it must be emphasized, only a good approximation.

With this limitation, therefore, the value of μ'^2 may be calculated for a given transmission frequency and transmission path. If we put $\mu' = \sin\phi_0$ and deduce z_v from $\sin\phi_0$ as was done in the previous paper, we may plot transmission curves, of virtual height against effective normal-incidence frequency, for the o and x components. When these curves are applied to the corrected normal-incidence-frequency—virtual-height curves, they may be expected to give reasonably good results. An example of this type of transmission curve is shown in Fig. 10. The curve marked x gives transmission conditions for the x component, and that marked o for the o component. The frequency used is well over the gyrofrequency $f_H = eH/2\pi mc$ so that the x component is returned from a lower level than is the o component and has a higher limiting frequency.

[12] E. V. Appleton, "Wireless studies of the ionosphere," *Jour. I.E.E.* (London), vol. 71, pp. 642–650; October, (1932).

It is not now justifiable to plot the transmission curves logarithmically, since the form of the curves will vary with the transmission frequency. For practical purposes, however, a logarithmic curve may be used within a limited range of frequencies about the frequency for which the curve is plotted; a practical limit might be, say, within ±15 per cent of this frequency.

The logarithmic sec ϕ_0 transmission curves may be used in estimating the maximum usable frequency

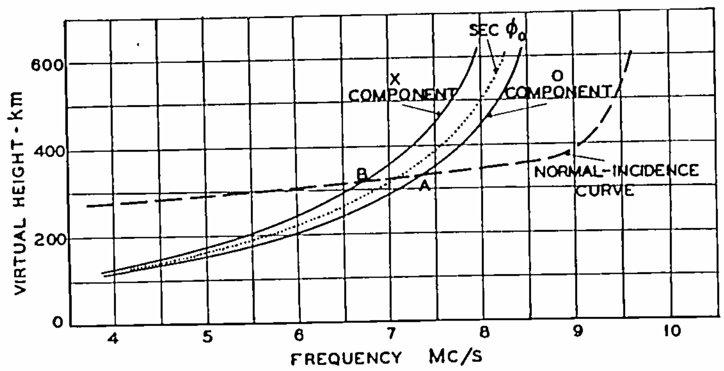


Fig. 10—Transmission curve, including effect of earth's magnetic field at level of reflection. Dotted line = secant-law curve ($f' = f \sec\phi_0$) for the given distance D and wave frequency f'. Dashed line = (z_v, f) curve for the o component, this being the component which, at vertical incidence is a measure of the ionization density N. The o component of the frequency f' over the given distance is reflected at A, and the x component at B.

for each component over a given path by adding or subtracting the separation between the limiting frequencies for the two components, evaluated at that frequency and distance. This separation is in general a function only of the transmission frequency and the quantity sec ϕ_0 and may be estimated within the limits of experimental error in most cases.

Fig. 11 gives the frequency to be added to or subtracted from the maximum usable frequency given by the logarithmic sec ϕ_0 transmission curves for the cases $H_L = 0$, $H_T \doteq 0$, and $H_T = H_L$. The $H_T = 0$ case is uninteresting save for single-hop transmission over the magnetic equator, close to the magnetic meridian. For transmission in the continental United States H_L is much less than H_T and, indeed, is negligible over east-west paths, so that such transmission we may consider as essentially transverse transmission. In this case

$$\mu'^2 = 1 - \frac{f_0^2}{f^2 - \delta\frac{f^2 f_H^2}{f^2 - f_0^2}}$$

where $\delta = 0$ for the o component and 1 for the x component, and the logarithmic sec ϕ_0 transmission curves give the correct maximum usable frequency for the o component. For this reason the o component lies along the sec ϕ_0 axis in the $H_L=0$ case in Fig. 11.

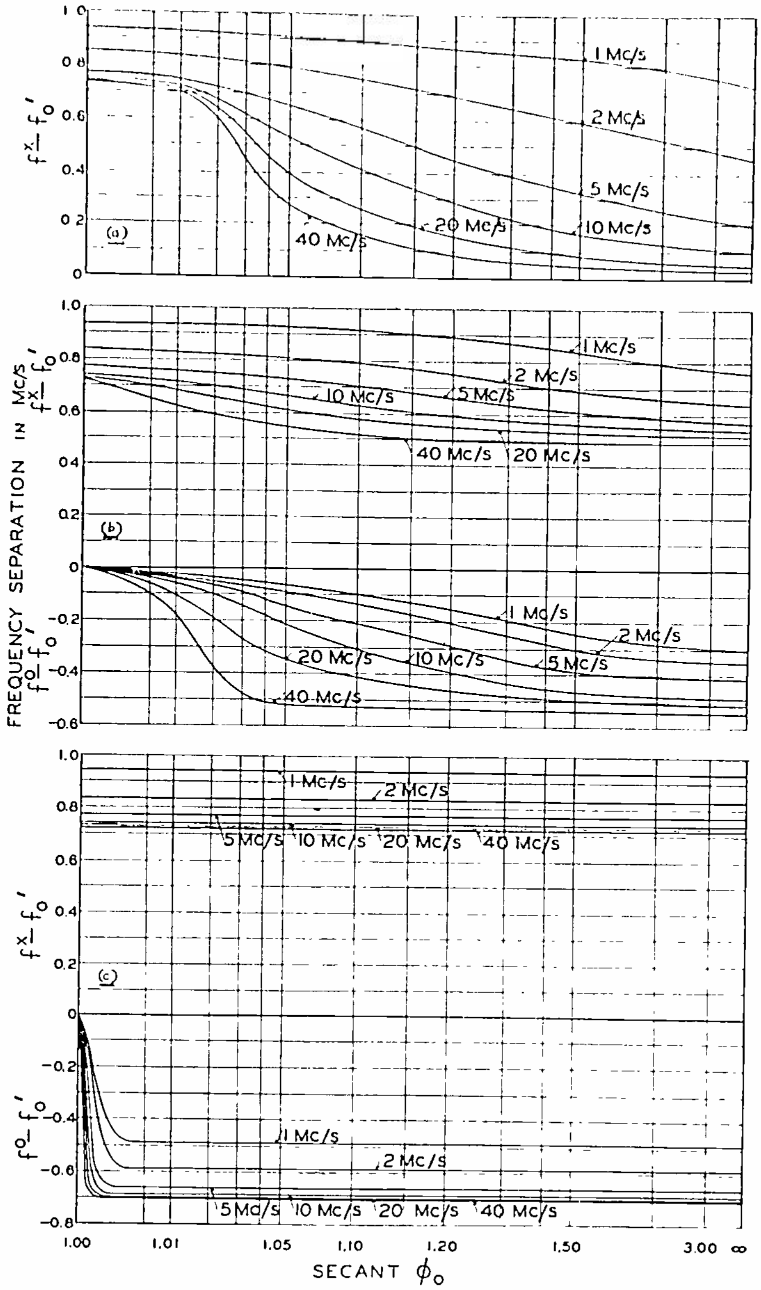


Fig. 11—f^x-f_0' and f^0-f_0' plotted against sec ϕ_0. f^x = frequency whose x component is reflected at the same level as is the o component of f^0. $f_0'=f'\cos\phi_0$. $H=\sqrt{H_T^2+H_L^2}=0.5$ gauss. The value of f' is given, in megacycles, on each curve.
(a) $H_L=0$. Here $f^0-f_0'=0$ and the curves for the o component lie on the sec ϕ_0 axis.
(b) $H_L=H_T\neq 0$.
(c) $H_T\doteq 0$. Here f^x-f_0' and f^0-f_0' are independent of sec ϕ_0, save near sec $\phi_0=1$.

For a maximum usable frequency for the o component below the gyrofrequency $f_H=\sqrt{f_T^2+f_L^2}$ the x component always has a maximum usable frequency above f_H. This is only important for E-layer transmission and only in cases where the x component is not too highly absorbed at frequencies near f_H. This case must not be confused with the case of a transmission frequency f' less than f_H in which case the x component is reflected from a level above the level where the o component is reflected.

Another effect of the anisotropy of the ionosphere due to the earth's field is to cause a difference in the directions of phase and energy propagation in the medium. This results in the wave's being reflected, not at the level where the direction of phase propagation is horizontal, but where the direction of energy flow (group direction) is horizontal. This effect has not been considered in these curves, and is probably not important to the degree of accuracy to which these calculations are carried.

IX. Behavior of the Wave Below the Point of Reflection

The development thus far has been given for any distribution of ionization in the ionosphere below the level of reflection. The effect on the reflection of the passage of the waves through the lower ionosphere, and for that matter through the lower atmosphere itself, has been taken care of by using the normal-incidence virtual height in the calculation, since this effect enters into the measurement of this height. Nothing has been said thus far, however, about phenomena that happen below the level of reflection.

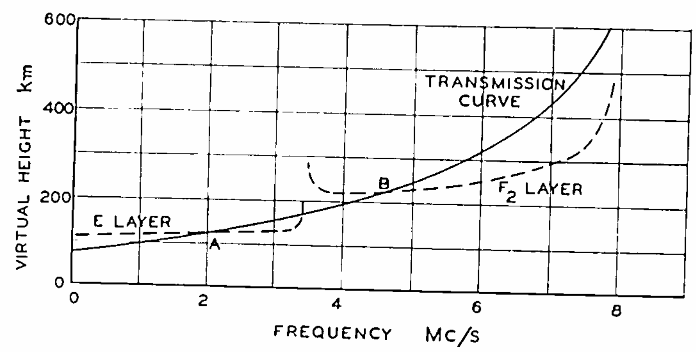


Fig. 12—Transmission by way of both E and F_2 layers.
A = point of reflection on E layer.
B = point of reflection on F_2 layer.
Solid line = transmission curve.
Dashed line = (z_v, f) curve.

An example of such a complication is given in Fig. 12. Here the transmission curve crosses both the E and the F_2 virtual-height curves. Therefore transmission may be expected by way of each layer. The angle of incidence is not the same for each layer, since the virtual heights of the layers are different. The wave reflected from the higher layer has a smaller angle of incidence upon the lower layer than does the wave reflected from the lower layer. Under certain conditions, then, it can penetrate the lower layer, and the waves can reach the receiver by way of the higher as well as the lower layer.

When the critical frequency of the lower layer is higher than the equivalent vertical-incidence frequency for the wave which would be returned from the upper layer, the wave will not penetrate the lower layer, but will reach the receiver by way of the lower layer only. Furthermore, if this equivalent vertical-incidence frequency is not below the critical frequency for the lower layer, but does lie in a region of appreciable absorption in this layer, the wave will be returned from the higher layer, but will be appreciably absorbed in so doing. Experimental observations indicate the absorption of the actual wave to be roughly the same as the absorption of a wave of the equivalent vertical-incidence frequency measured at vertical incidence.

It is possible to plot, on the sec ϕ_0 transmission curves, lines corresponding to the values of sec ϕ_1 (see Fig. 6) for different heights. Such a set of "absorption lines" is shown in Fig. 13. When the transmission curve is superimposed on the (z_v, f) curve the behavior of a wave below the reflection level may be estimated by the region of the (z_v, f) curve sheet through which the absorption line through the reflection point passes. If this line passes through a region of absorption or cuts a lower layer the wave will be absorbed or will not penetrate through to the higher layer.

These absorption lines are the lines sec ϕ_1 = constant for a plane earth and for short distances on a curved earth. They curve toward larger values of sec ϕ_1 for lower heights in the case of greater distances over a curved earth, and approach the transmission curve itself for great distances. An example of the use of the transmission curves and absorption lines is shown in Fig. 14, which gives vertical-incidence curves and the corresponding frequency-height curves derived therefrom for the transmitted wave over a distance. From these latter curves can be deduced, of course, the equivalent paths or group retardation of the waves over the transmission path.

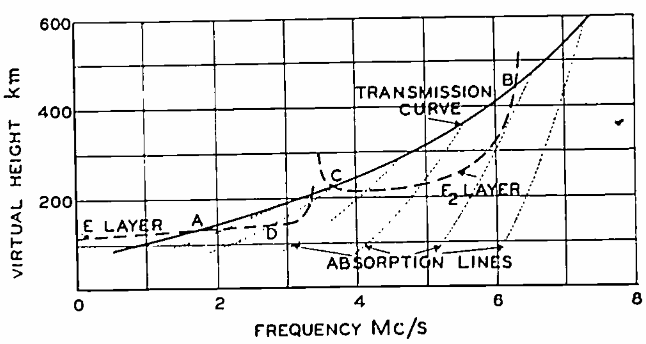


Fig. 13—Absorption lines on transmission curve.
Dashed line = (z_v, f) curve.
Dotted lines = absorption lines.
E reflection takes place at point A and F_2 reflection at point B.
F_2 reflection at C is shielded by the E layer at D.

X. Angle of Departure and Arrival of the Waves

Disregarding any possible asymmetry of the wave trajectory due to the earth's magnetic field the angles of departure of the waves from the transmitter and of arrival of the waves at the receiver are equal, assuming one ionosphere reflection. The amount of energy which is radiated from the transmitter at a given angle from the horizontal depends, especially for low angles, on the design of the transmitting antenna, and to a great extent on the nature of the terrain surrounding the transmitter. Similar factors affect the energy absorbed from the wave at the receiving station.

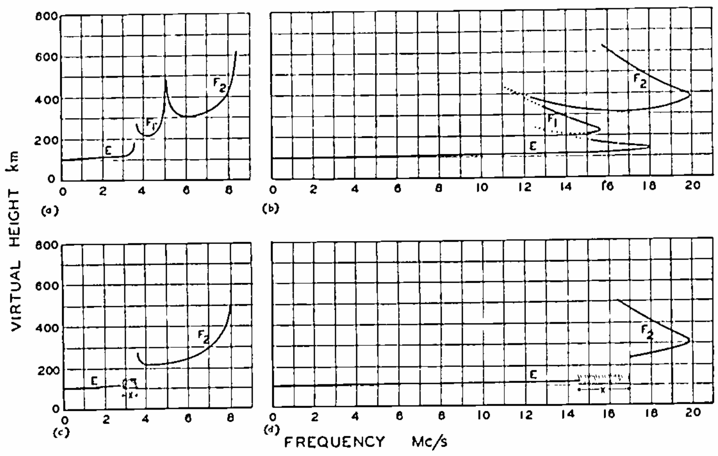


Fig. 14—Virtual heights at vertical and oblique incidence.
(a) (z_v, f) curve showing E, F_1, and F_2 layers.
(b) oblique-incidence heights corresponding to (a). Dotted lines show parts of curves shielded by E layer. Note transmission of some frequencies by several layers. This shows how the F_1-layer reflections are relatively unimportant for transmission.
(c) (z_v, f) curve showing absorption (X) above critical frequency for the E layer.
(d) oblique-incidence heights corresponding to (c). Absorption here (X) completely blocks out some frequencies.
All curves are for the o component.

For a station located on or near the ground no energy will be radiated below the horizontal, and but little until the angle ψ of departure of the waves above the horizontal becomes appreciable. The minimum value of ψ at which sufficient energy is radiated (or received) to produce a readable signal varies with the terrain, the power of the transmitter, and the sensitivity of the receiver. Over sea water ψ can be very nearly zero; over land the minimum value may be several degrees. A fair average approximation may be that ψ must exceed about $3\frac{1}{2}$ degrees.

A simple geometrical calculation gives ψ in terms of sec ϕ_0 for various distances, and these may be noted on the transmission curves. The point of reflection must then correspond to an angle ψ greater than the minimum value assumed for the terrain at the receiver and transmitter.

The effect of this limitation is to limit the maximum distance for single-reflection transmission to about 1750 kilometers (over land) for the E layer and about 3500 to 4000 kilometers (over land) for F_2-layer transmission. Where the transmission path exceeds these values in length calculations must be made on the basis of multireflection transmission.

The condition of the part of the ionosphere traversed by the wave determines the behavior of the wave. In calculating transmission conditions, therefore, the ionosphere around the middle of the path, for single-reflection transmission, and around the middle of each jump, for multireflection transmission, must be considered. The latter requirement means that usually, for practical calculations one must consider the condition of ionosphere over all the transmission path except a few hundred kilometers near the transmitter and receiver. This question has been discussed in detail elsewhere.[13] It must also be kept in mind that at times the o and x components are reflected from different geographical parts of the ionosphere.

XI. Determination of sec ϕ_0

Fig. 15 gives an alignment chart for the rapid determination of the factor sec ϕ_0 to be used in calculating the logarithmic transmission curves. To use the chart, place a straightedge so that it passes through the desired virtual height and the desired distance laid off on the distance scale at the lower left-hand edge of the chart (increasing distances lie to the left). The ordinate of the intersection of the straightedge with the vertical line corresponding to the same desired distance laid off on the main distance scale (increasing distances lie to the right) gives the value of sec ϕ_0. The relation of the point of intersection to the curved dashed lines of equal ψ gives the value of the angle of departure of the waves from the horizontal. A point of intersection falling above the $\psi=0$ degrees line indicates an impossible case, where the ray would have to depart at an angle below the horizon.

For example, a distance of 2400 kilometers and a virtual height of 300 kilometers corresponds to a sec ϕ_0 of 3.07, and an angle of departure of 8.02.

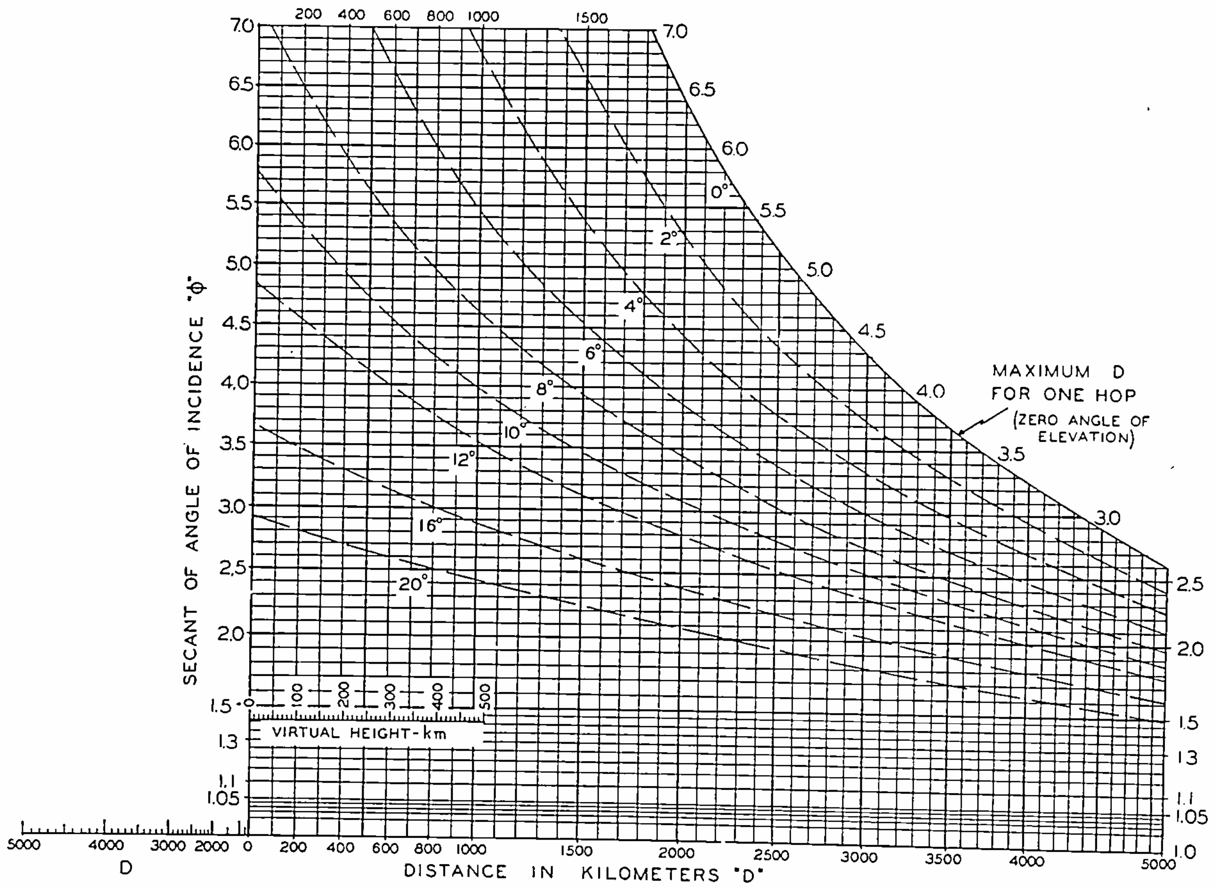


Fig. 15—Alignment chart for determining sec ϕ_0 and ψ, the angle which the ray makes with the horizontal at the transmitter or receiver. The curved dashed lines are lines of equal ψ.

[13] T. R. Gilliland, S. S. Kirby, N. Smith, and S. E. Reymer, "Characteristics of the ionosphere and their applications to radio transmission," Proc. I.R.E., vol. 25, pp. 823–840; July, (1937); *Nat. Bur. Stand. Jour. Res.*, vol. 18, pp. 645–668; June, (1937); the weekly radio broadcasts of the National Bureau of Standards on the ionosphere and radio transmission conditions; Letter Circular LC499.

XII. Transmission Factors

When average transmission conditions over a period of time or when a variety of transmission paths are to be considered, or when an estimate of the maximum usable frequencies is to be made without a precise knowledge of the ionosphere over the transmission path, it is convenient to have available a means by which the maximum usable frequencies may be quickly estimated from an approximate value of the vertical-incidence critical frequency.

The National Bureau of Standards is now beginning a compilation of factors by which the critical frequency for the *o* component, measured at vertical incidence, may be multiplied in order to obtain the maximum usable frequencies. These factors are based on average observations over a period of time, and may be applied either to average critical frequencies to give average transmission conditions, or to a given observation of a critical frequency to obtain approximate transmission conditions at a given time.

XIII. Conclusions

The type of transmission curves described above has been in use at the National Bureau of Standards for the past two years in studying the correlation of high-frequency radio transmission conditions with regular ionosphere observations. The results of continuous measurements of high-frequency broadcast stations and observations on other high-frequency signals, as well as the results of some specific experiments have been compared with vertical-incidence data. Practically all the available data agree with what would be expected on the basis of the theory outlined above, and the exceptions may in most cases be accounted for.

Characteristics of the Ionosphere at Washington, D. C., March, 1939*

T. R. GILLILAND†, ASSOCIATE MEMBER, I.R.E., S. S. KIRBY†, ASSOCIATE MEMBER, I.R.E.
AND N. SMITH†, NONMEMBER I.R.E.

DATA on the critical frequencies and virtual heights of the ionosphere layers during March are given in Fig. 1. Fig. 2 gives the monthly average values of the maximum frequencies which could be used for radio sky-wave communication by way of the regular layers. Fig. 3 gives the distribution of the hourly values of F- and F_2-layer critical frequencies and maximum usable frequencies about the average for the month. Fig. 4 gives the expected values of the maximum usable frequencies for transmission by way of the regular layers, average for June, 1939. The ionosphere storms and sudden iono-

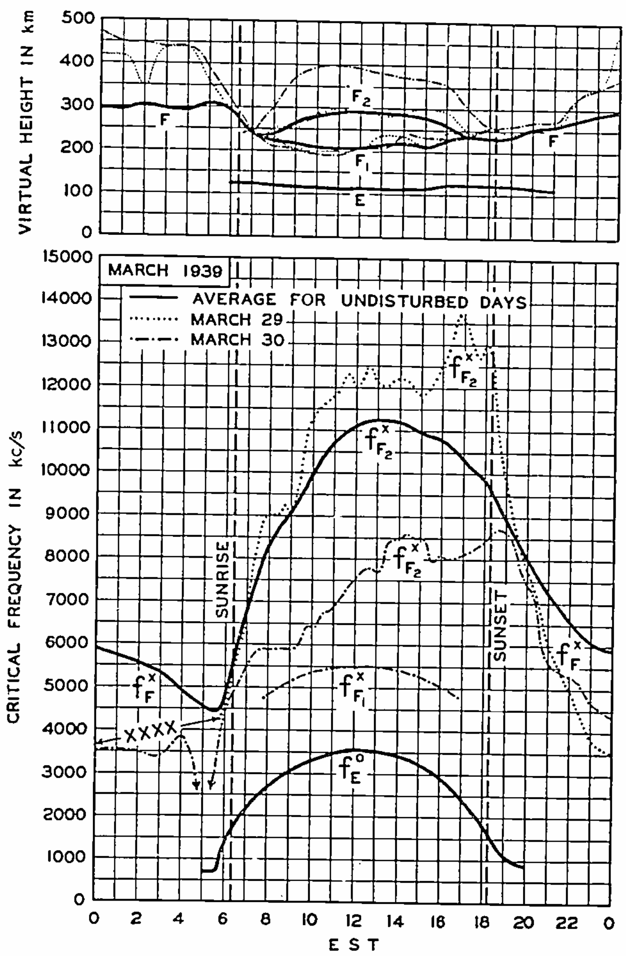


Fig. 1—Virtual heights and critical frequencies of the ionosphere layers, March, 1939. The solid-line graph is the undisturbed average for March, 1939. The dotted-line and dashed-line graphs are for the ionosphere storm days of March 29 and 30, respectively. Parts of two separate storms on March 29 are shown. The first ended at 0600 and the second began at 1500 E.S.T. The crosses represent the times on March 29 when the reflections were so diffuse that the critical frequencies could not be determined.

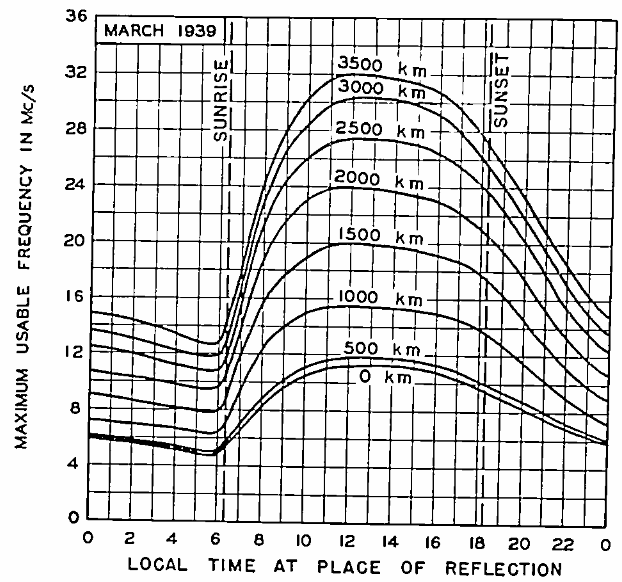


Fig. 2—Maximum usable frequencies for sky-wave radio transmission. Averages for March, 1939, for undisturbed days, for dependable transmission by the regular F and F_2 layers.

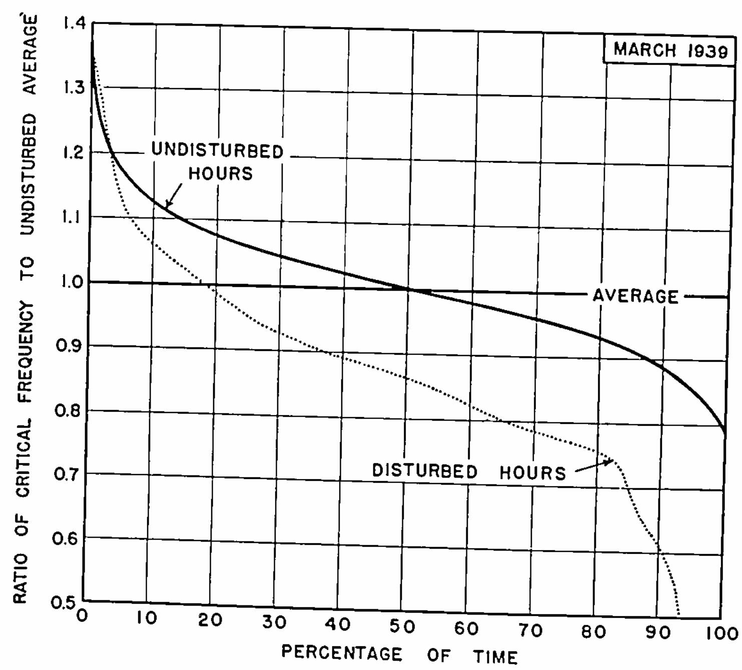


Fig. 3—Distribution of F and F_2 critical frequencies (and approximately of maximum usable frequencies) about monthly average. Abscissas show percentage of time for which the ratio of the critical frequency to the undisturbed average exceeded the values given by the ordinates. The graphs give data as follows: solid line, 399 undisturbed hours; dotted line, 140 disturbed hours listed in Table I.

* Decimal classification: R113.61. Original manuscript received by the Institute, April 10, 1939. These reports have appeared monthly in the PROCEEDINGS starting in vol. 25, September, (1937). See also vol. 25, pp. 823–840; July, (1937). Publication approved by the Director of the National Bureau of Standards of the U. S. Department of Commerce.

† National Bureau of Standards, Washington, D. C.

sphere disturbances are listed in Tables I and II, respectively.

It was observed this month, as previously, that during ionosphere storms at night, additional reflections frequently appear with retardations equivalent to several hundred kilometers in excess of the F-layer

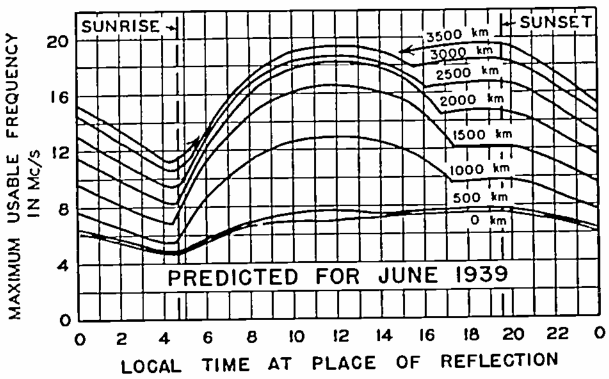


Fig. 4—Predicted maximum usable frequencies for dependable sky-wave radio transmission by way of the regular layers, for June, 1939. The F layer will ordinarily determine the maximum usable frequencies at night, and, except for nearly vertical incidence, the E and F_1 layers will determine them during the day. The effect of the E and F_1 layers is shown by the hump on the graphs during the middle of the day. The values shown will be considerably exceeded during occasional irregular periods by reflections from clouds of sporadic-E layer.

retardations. The ionized region responsible is often sharply defined. The retardations from it usually decrease over a period of several hours so that the reflections merge with the F-layer reflections. The retardations sometimes decrease with increasing frequency. This is similar to the phenomenon reported by Leithäuser and Beckmann.[1]

Sporadic-E reflections were observed at 6 megacycles or above at no time during the month, and at 4.5 megacycles for only three scattered hours.

[1] *Zeit. für tech. Phys.* vol. 18, p. 290, (1937).

TABLE I

IONOSPHERE STORMS (APPROXIMATELY IN ORDER OF SEVERITY)

Date and hour E.S.T. 1939 March	h_F before sunrise (km)	Minimum f_F^z before sunrise (kc)	Noon $f_{F_2}^z$ (kc)	Magnetic character[1] 00–12 G.M.T.	12–24 G.M.T.	Ionosphere character[2]
29 (after 1500)	—	—	—	1.6	1.2	1.7
30	445	below 2500	7500	0.9	0.8	1.5
31 (until 0600)	338	3500	—	0.6	0.4	0.5
27 (after 2000)	—	—	—	0.4	0.9	0.3
28	350	3800	about 10,500	0.9	1.7	1.1
29 (until 0600)	402	diffuse	—	1.6	1.2	1.5
21 (after 0200)	328	diffuse	8900	0.9	0.9	1.0
22 (until 0600)	330	3600	—	0.9	0.8	0.5
6 (until 0600)	328	3400	—	0.3	0.4	0.4
3 (after 2200)	—	—	—	0.5	0.7	0.1
4 (until 1400)	332	4600	about 10,500	0.9	0.6	0.4
22 (after 2200)	—	—	—	0.9	0.8	0.1
23 (until 1200)	318	4000	10,200	0.7	0.7	0.2
8 (0100 to 0500)	322	4700	—	0.6	0.4	0.1
For Comparison: Average for undisturbed days	299	4540	11,100	0.4	0.4	0.0

[1] American magnetic character figure, based on observations of seven observatories.

[2] An estimate of the severity of the ionosphere storm at Washington on an arbitrary scale of 0 to 2, the character 2 representing the most severe disturbance.

TABLE II

SUDDEN IONOSPHERE DISTURBANCES

Date 1939	G.M.T. Beginning	End	Location of transmitter	Relative intensity at minimum[1]
March 20	1510	1530	Ohio, Mass., Ont., D. C.	0.0
21	1516	1545	Ohio, Mass. Ont. D. C.	0.0
21	1823	1835	Ohio, Ont., D. C.	0.01

[1] Ratio of received field intensity during fade-out to average field intensity before and after; for station W8XAL ,6060 kilocycles, 650 kilometers distant

Institute News and Radio Notes

Board of Directors

A regular meeting of the Board of Directors was held on April 5 and attended by R. A. Heising, president; Melville Eastham, treasurer; H. H. Beverage, Ralph Bown, F. W. Cunningham, Alfred N. Goldsmith, Virgil M. Graham, O. B. Hanson, L. C. F. Horle, I. J. Kaar, F. B. Llewellyn, Haraden Pratt, B. J. Thompson, H. M. Turner, A. F. Van Dyck, and H. P. Westman, secretary.

C. R. Barhydt, J. G. Brainerd, and Alfred Decino were transferred to Member grade.

Fifty-three applicants for Associate membership, three for Junior, and thirty for Student were admitted.

A number of letters approving the changes made in the PROCEEDINGS starting with Volume 27 were read. To date, no criticisms of these changes have been received.

The progress being made in preparations for the San Francisco Convention was reported.

Committees

Admissions

The Admissions Committee met on April 5. Those present were F. W. Cunningham, chairman; H. H. Beverage, Melville Eastham, J. F. Farrington, L. C. F. Horle, H. M. Turner, A. F. Van Dyck, and H. P. Westman, secretary.

One of three applications for transfer to Fellow grade was approved. These applications were submitted prior to the change in the Institute Constitution and the Board of Directors agreed that if an application were approved, the Board would issue an invitation to the applicant to transfer to Fellow grade.

There were twenty applications for transfer to Member grade acted on. Seventeen of these were approved, two were denied, and one was tabled pending the obtaining of additional data. Three applications for admission to the grade of Member were acted on and approved.

Awards

A luncheon meeting of the Awards Committee was held prior to the Board meeting on April 5 and recommendations prepared for the Institute Medal of Honor and Morris Leibmann Memorial Prize for 1939. The names of the recipients were agreed on but citations could not be prepared because of lack of time. These will be prepared prior to the May meeting of the Board of Directors at which time final action will be taken. Those present at the meeting were Haraden Pratt, chairman; H. H. Beverage, Ralph Bown, Melville Eastham, H. M. Turner, and H. P. Westman, secretary.

Membership

The Membership Committee met on March 14. Those present were C. E. Scholz, chairman; I. S. Coggeshall, E. D. Cook, H. F. Dart, H. C. Gawler, L. G. Pacent, J. R. Poppele, C. R. Rowe, Bernard Salzberg, and H. P. Westman, secretary.

It was agreed that in the future, minutes of Membership Committee meetings would be forwarded to the chairman of each section.

The desirability of preparing a standard form which can be posted on bulletin boards in various organizations whose personnel might be interested in the Institute was discussed. Arrangements are being made for the preparation of such a form on which the card announcing the latest meeting may be mounted.

Because the changes in the Institute Constitution which were voted by the membership recently involved the qualifications for membership, new application blanks are required. The arrangement and wording of these blanks were agreed on.

A lengthy discussion on Student memberships was held and a number of proposals adopted which should increase the number of active Students.

New York Program

The New York Program Committee met on February 24 and those present were I. S. Coggeshall, chairman; Austin Bailey, A. B. Chamberlain, J. D. Crawford, assistant secretary, A. B. Goodall, A. F. Van Dyck, and H. P. Westman, secretary. Arrangements were made for the next several New York meetings of the Institute.

Registration of Engineers

A special committee appointed to investigate the situation in regard to the New York State law for the registration of professional engineers met on February 10. Those present were L. M. Hull, chairman; L. C. F. Horle, E. L. Nelson, A. F. Van Dyck, and H. P. Westman, secretary. The committee completed a report for submission to the Board of Directors.

Tellers Committee

The Tellers Committee met on February 25 to count the ballots cast in the vote on the amendement of the Institute Constitution. The results of this vote have already been reported to the membership. Those present at the meeting were Austin Bailey, chairman; H. F. Dart, D. G. Fink, and H. P. Westman, secretary.

André Blondel

André-Eugène Blondel was born on August 28, 1863. In 1888 he was graduated from the National School of Bridges and Highways with high honors, finishing his studies in mathematics and physics in 1885 and 1889, respectively. He entered the government engineering service and was appointed to the Central Service of Lighthouses and Beacons.

After the historical experiments of Marconi, he became interested in wireless telegraphy. In 1898 he disclosed the first exact theory of electromagnetic-wave propagation. He compared the antenna-and-ground system to a vertical Hertzian oscillator insulated in space and of length double that of the antenna (this theory being completed in 1903). During the same period, he tested the use of the telephone as a receiver when combined with a vacuum tube. In 1900 he employed alternating current to charge the condenser of the oscillatory circuit in the generation of radio waves as had been done by Tesla and d'Arsonval.

ANDRÉ BLONDEL

From 1898 to 1900 he recommended the use of the musical spark which permits acoustic selectivity, thus improving the ability to operate through interference.

In 1902 he had already indicated the principles of continuous-wave telephony and proposed in 1907 the use of a direct-current arc as a suitable generator. Experiments of this nature were made on the Eiffel Tower by Lieutenant Paul Brenot.

During the same period, Blondel published a theoretical and experimental study of resonant transformers.

By 1902 he had disclosed radiating systems capable of concentrating waves into beams by applying a principle which was similar to that of optical gratings. He

Redwood Empire Association

Every ship entering San Francisco Bay must pass through its famed Golden Gate which is now spanned by one of the most recent large bridges built in this country. It offers a new connecting link between San Francisco on the southern peninsula and much interesting country to the north.

had also recognized the basic characteristics of loop antennas, and thus had established the fundamentals of radio direction finding. Verifying experiments were made in 1908 by Captain Ferriè with whom he had collaborated since about 1900 and Lieutenant Brenot. In 1911 and 1919, Blondel described further advances in the theory of these devices and described newer models.

Blondel visualized the possibility of using electromagnetic waves for the remote control of equipment and with the advent of the vacuum tube he devoted considerable time to the experimental and theoretical study of this subject.

In 1919 and 1920 he published papers on the operation of vacuum tubes as amplifiers and oscillators and classified the various possible methods of operation. This work is evident in the current method of designating tube operation as classes A, B, and C.

He received many honors. In 1927 he was nominated Commander of the Legion of Honor. As a member of the Academy of Sciences, he was Honorary President of the French Society of Electrical Engineers. He received the Mascart Medal, Rouville Prize of the Department of Public Works, Medal of the Franklin Institute, Montefiore Prize, and the Lord Kelvin Medal. The Institution of Electrical Engineers awarded him the Faraday Medal and the Belgian Electrical Engineers bestowed on him the Gustave Trasenster Medal.

He died on November 15, 1938, after a short illness.

Sections

Atlanta

Five speakers presented reports on the Broadcast Engineering Conference held at Ohio State University.

A. W. Shropshire, transmitter engineer of WSB, outlined the division of the material to be presented and introduced the speakers. The first speaker was F. S. Holliday, also a transmitter engineer at WSB, who discussed the paper by E. E. Spitzer on the "Design of Transmitting Tubes." Tubes were treated element by element considering the characteristics of the materials used, design considerations, and production methods. This was followed by a discussion of modes of operation and ultra-high-frequency applications.

J. M. Burke, Jr., transmitter engineer at WAGA, reviewed the paper on "Electron Optics," by V. K. Zworykin. The effects on the electron stream of the arrangement of the various elements in a cathode-ray tube were considered. A comparison of electron and light optics followed. Data were then given on the iconoscope, kinescope, and projection kinescope. The review was concluded with a discussion of television standards and ultra-high-frequency propagation.

The paper on "Measurements on Broadcast Antennas," by D. B. Sinclair, was reviewed by J. M. Comer, Jr., chief engineer of WTAL. Substitution-method and resistance-variation-method errors were particularly stressed. The paper was concluded with a consideration of bridge methods in general and the more prominent errors encountered.

The paper by J. F. Morrison on "Antennas" was reported by A. W. Shropshire. The performance characteristics of broadcast transmitting antennas were reviewed. Structures of different cross sections and electrical heights were compared as to performance. The paper was concluded with a discussion of coaxial transmission lines with particular emphasis on harmonic suppression and tower-lighting circuits.

Ben Akerman, chief engineer of WGST, reviewed the panel discussion on "Standards of Good Engineering Practice" which had been led by A. D. Ring, J. H. DeWitt, Jr., and S. L. Bailey. This was followed by a discussion of the new volume indicator recently adapted for broadcasting. The Paper by H. A. Chinn on "Studio Equipment" was covered briefly. This report was concluded with a short discussion on studio reverberation and a device for recording it.

February 24, 1939—Ben Akerman, chairman, presiding.

Under the guidance of J. W. Spratlin, communication supervisor of the Delta Air Corporation, an inspection tour of the radio communication facilities at Candler Field was made.

The first trip was to the airport traffic-control tower which controls all aircraft using the field. Next, the operating room of the Civil Aeronautics Authority plant where teletypewriter equipment, radio receivers, and the controls for the radio range station are installed was visited.

At the station of the Delta Airlines, the functions of air-line communications were explained and normal operation and equipment were observed.

The Eastern Airlines plant was next visited and information was supplied on its operation and the switching system whereby any operator on duty has immediate access to any of the transmitters installed at a remote location. These transmitters were described and later inspected.

At the radio maintenance shops of Delta Airlines, the features of a new transmitter under construction were described. A complete transport-aircraft station was set up and its operation explained. At the Delta plane shops, one of the planes was examined and its operation and the functions of the various instruments and controls used in flight were described.

The tour was completed by a visit to the transmitting station of the Civil Aeronautics Authority. A radio range transmitter of recent design was inspected. In addition, a 75-megacycle transmitter which had been used in blind landing experiments was described and displayed.

March 16, 1939—Ben Akerman, chairman, presiding.

Buffalo-Niagara

J. R. Nelson of the Raytheon Production Corporation, presented a paper on "Input Characteristics of Converter Tubes." In it numerous constructions of these tubes were considered and included those using a single electron stream and two independent streams. Characteristics of tubes of these types were given. The paper was concluded with a discussion of the reactions of the elements, neutralization of space-charge coupling, and electron hysteresis currents.

March 8, 1939—H. C. Tittle, chairman, presiding.

Chicago

C. S. Roys, professor in the department of electrical engineering of Purdue University, presented a paper on "Predetermination of the Performance of Radio-Frequency Amplifiers." Dr. Roys described a method for predicting the upper limits at which a vacuum tube will oscillate. This method neglects the transit time of the electrons and gives predictions of higher limits than those obtained experimentally. It was pointed out that the method rather than the actual values obtained was the important feature.

The interelectrode capacitances are shunted by external capacitances which are larger by a definite factor and the measurements are made at audio frequencies. From these measurements the radio-frequency behavior when the shunted capacitances are removed is predicted. This method has been used in the past in the testing of high-power transmitting tubes.

February 17, 1939—V. J. Andrew, chairman, presiding.

Cincinnati

"Facsimile Transmission and Reception" was the subject of a paper by J. F. Silver, special development engineer of the Crosley Corporation.

The subject was introduced with a résumé of early developments in the field. Several methods now being used were described and shown to be basically similar with the principal differences being in the mechanical arrangements used in moving the scanning spot over the copy. Systems using both linear and constant acceleration of the scanner spot and recorder stylus were described. Various methods for recording impressions on paper were described and included photographic, pressure on carbon paper, and dry and wet electrolytic systems.

A complete facsimile system was demonstrated. It consisted of a Finch scanner and Crosley "Reado" or facsimile recorder. A cathode-ray oscilloscope reproduced visually the print and synchronizing signals generated by the scanner.

March 14, 1939—H. J. Tyzzer, chairman, presiding.

Connecticut Valley

V. K. Zworykin of the RCA Manufacturing Company, presented a paper on "Electron Optics as Applied to Television."

He presented analogs relating electron and light optics as applied to television, an historical discussion of scanning tubes, and analyses of recently developed iconoscopes and kinescopes. Television-tube focusing and deflection were explained by comparison to optical prisms or lenses of varying density and refractive index.

In the historical discussion of television systems that followed, he described the Nipkow system, secondary-emission amplifiers, the Farnsworth dissector, the iconoscope, and the present RCA kinescope. The difference between non-storage and storage-type television cameras was pointed out.

March 15, 1939—E. R. Sanders, chairman, presiding.

Detroit

"Radio-Frequency Measuring Equipment, was the subject of a paper by D. B. Sinclair, research engineer of the General Radio Company.

Dr. Sinclair pointed out that either resonance or null measurements are commonly used at radio frequencies. They are similar in that they compare the unknown with a known impedance but differ fundamentally in the means of indication.

Resonance methods depend on tuning resonant circuits to a voltage or current maximum while the null method requires balancing to a voltage or current null. The major point of difference lies in the manner of determining the resistive component.

Series-resonant methods, which depend on measuring the current in series-resonant tuned circuits, are best adapted for the measurement of low impedances because primarily they are methods for determining the series resistance of such circuits.

The magnitude of unwanted residual parameters in the standard and in the wiring usually determines the accuracy and range of any given circuit. The magnitudes of both the main and residual parameters change with frequency because of current redistribution or skin effect.

The variable condenser has been found, in general, to be the best reactance standard at radio frequencies. The residual inductance and resistance are very small compared with the capacitance, and skin effect has only a second-order effect on the capacitance.

The designer of decade resistors may reduce these residual effects by substituting compensating copper coils for resistance cards. By careful matching of inductances of the copper coils and resistance units, the inductance of the assembly may be maintained constant to within about 0.05 microhenry.

February 17, 1939—Lynne C. Smeby, chairman, presiding.

Emporium

President Heising, who is the Director of Radio Research at Bell Telephone Laboratories, is planning to visit as many of the Institute sections as possible. His first visit was to Emporium.

After discussing the recent changes in the Institute Constitution and various other aspects of the operation of the organization, he presented a paper on "Radio Extension Links to the Telephone System."

He reviewed first the early history of radio communication with particular emphasis on the progress which has been made in transoceanic and transcontinental lines as well as service to marine stations.

Various devices associated with the transmitting and receiving equipment and which have been vital factors in the development and usefulness of these types of communication were described. These special instruments include the vodas, vogad, codan, compander, and calling devices.

Improvements gained by the introduction of single-side-band transmission and its extension to twin-channel single-side-band operation were described.

The paper was concluded with a discussion of various directive antennas with special emphasis on the rhombic type which in multiple form is called the musa.

March 7, 1939—R. M. McClintock, chairman, presiding.

Los Angeles

Chester Davis, an engineer for the RCA Manufacturing Company, presented "A Discussion of the RCA Television Exhibit at the San Francisco World's Fair."

The iconoscope transforms light impulses into electrical signals. The electron gun, which projects a narrow beam of electrons onto the mosaic of photoelectric elements on a screen within the tube, was described. A multiplier in the newer iconoscopes increases the sensitivity over previous models by a factor of ten. The beam is deflected electromagnetically to scan the screen 30 times per second, which, because of interlacing, gives the equivalent of 60 frames per second. 441-line scanning is employed. The output of the tube, having a frequency range of several megacycles, is passed through wide-band amplifiers and modulates the transmitter.

At the receiver, the amplified signal modulates the electron beam in the kinoscope to vary the light intensity of each elemental area. Electromagnetic deflection of the beam is synchronized with that of the iconoscope at the transmitter.

Transmission circuits must be able to pass a band approximately 5 megacycles wide without serious attenuation or phase shift. Transmitters operate in the range from 40 to 120 megacycles and the maximum distance at which they can be received is limited by the characteristics of these waves.

The National Broadcasting Company's system in New York was described. The transmitter in the Empire State Building is connected to the studios in the RCA Building by a transmitter operating at 177 megacycles. The distance is approximately one mile.

The paper was concluded with a short description of recent developments in facsimile broadcasting and equipment.

A motion picture on "Broad-Band Transmission over Coaxial Cable" was shown through the courtesy of Electrical Research Products, Inc. It indicated that high-quality television signals can be transmitted over such circuits.

H. R. Lubcke, engineer for the Don Lee Broadcasting System, presented a "Discussion of the Technique of Television Broadcasting."

The pickup camera has been made portable so that it may be easily and quietly moved about the studio. Its output is passed through wide-band amplifiers and then through a coaxial cable to the top floor of the Don Lee Building.

A 5-kilowatt lighting load is usually employed. Approximately 20 hours of preparation is required to produce one episode of a show. The script indicates lighting and camera position as well as sound effects.

The staff comprises the television producer who is in charge of the show, a camera man and assistant, an operator to handle the microphone boom, to keep the microphone as close to the action as possible but out of the camera range, one or two prop men, a mixer operator, a sound-effects man, and one or two lighting men. A television supervisor is located at a near-by receiving point and keeps in touch with the studio by telephone. He makes suggestions regarding lighting, focus, composition, and sound.

Photographs were exhibited of television images received approximately 30 miles from the transmitter. Reliable operation is secured up to 20 miles.

In the discussion, it was brought out that satisfactory synchronization may be secured when both transmitter and receiver are operated from power circuits of the same frequency although not necessarily from the same generating station. However, if the transmitter is operated from a 60-cycle circuit and the receiver from a 50-cycle line, considerable difficulty is encountered.

February 21, 1939—F. G. Albin, chairman, presiding.

Convention Headquarters in San Francisco will be at the Hotel Mark Hopkins shown above. It is not too early to make reservations as all hotels will be well occupied by visitors to the San Francisco World's Fair.

"Carrier-Current Equipment and Application on the Boulder Dam Transmission System" was the subject of a paper by J. D. Laughlin of the Design and Construction Department, Bureau of Light, City of Los Angeles.

This transmission line connects Boulder City, Nevada, and Los Angeles, California, which are 266 miles apart. It normally handles 240,000 kilowatts, about 80 per cent of its total capacity. The copper conductors are 1.4 inches outside diameter and hollow to reduce weight. The line is operated at about 275,000 volts.

A carrier-current supervisory control provides pilot channels for switching, control channels for operating remote switches, and a telephone circuit between the terminals of the line. Ten bands, each ten kilocycles wide are used between 50 and 150 kilocycles. Terminal equipment is of conventional design and is coupled to the line through capacitors. Since there are two three-phase power lines, six conductors are available. Telephone and supervisory control is operated between two conductors and ground which permits operation even if one line is out of the circuit. Chokes and by-pass capacitors are employed at switching points to reduce the losses in the carrier-current system.

The telephone circuit used is a single-side-band suppressed-carrier-type of transmission. Voice-operated relays control the direction of transmission. Ringing is accomplished by a 1600-cycle tone.

The paper was concluded with the showing of a sound motion picture "Modern Pioneers" which described the construction of the line across the mountains and the desert of Southern California.

March 21, 1939—F. C. Albin, chairman, presiding.

New York

Two papers on facsimile were presented. The first on "The Handling of

Telegrams by Facsimile Methods" was by R. J. Wise and I. S. Coggeshall of the Western Union Telegraph Company.

The particular demands for point-to-point commercial services of the type which have been established for public use between New York City, Buffalo, and Chicago resulted in the development of unusual equipment. Automatic loading and discharge of messages to be transmitted, dry-process recording, minimizing of wasteful scanning, and an unusually high speed of trace and scanning rate have been obtained. Equipment capable of handling fifty telegrams per hour is now in service.

A message-filing equipment for unattended service was described and demonstrated. The customer need only deposit the blank into a slot in order to file it. The blank is automatically clipped to a rotating cylinder, scanned, stripped off, and deposited in the bottom of the container to be picked up at a later time by the operating company for record purposes.

The second paper, "Finch Facsimile," was described by R. H. Marriott, consulting engineer. In this system, transmission is from a continuous strip of copy which moves past the scanning head. The head is mounted on an axle and is turned from side to side so that a spot of light about 0.01 inch in diameter projected from it traces a series of adjacent parallel paths across the copy. The copy paper is bent concave with respect to the scanning head so that the light path is of constant length.

The amount of light reflected from the copy and picked up by the photocell, which is also mounted in the scanning head, depends on the variation from white to black of the copy. The current from the photocell varies the intensity of a modulating signal of about 2000 cycles which actuates the broadcast transmitter.

The scanning signal is interrupted at the beginning of each scanning swing by a brief synchronizing pulse of about 500 cycles. This pulse, in combination with speed governors on the motors, keeps the recorders in step with the scanner.

At the recorder, a stylus is used which has contact with the paper over an area equal to that covered by the scanning beam at the transmitter. An electric current passes through the paper from the stylus and changes the white heat-sensitive paper to gray or black depending on the intensity of the current.

March 1, 1939—President Heising, presiding.

"Synthetic Reverberation," by P. C. Goldmark and P. S. Hendricks of the Columbia Broadcasting System, was presented and demonstrated.

A disk, twenty inches in diameter, is rotated at a speed of about 400 revolutions per minute. Around the periphery of the disk fluorescent material has been deposited. A light, modulated by the signal, is projected onto the disk and results in fluorescence. Phototubes are arranged at one or more points around the disk and are actuated by the fluorescence. The phototube currents pass through amplifiers back to the main signal circuit. By adjusting the gain of these amplifiers, the "echo" can be made of any desired intensity. The "reverberation time" of the system can be adjusted by varying the speed of the disk, the spacing of the phototubes, or the size of the disk.

A demonstration of recorded music, a soloist, and various sound effects was given.

April 5, 1939—President Heising, presiding.

Philadelphia

"Wave-Guide Radiators and Electromagnetic Horns," the demonstration-lecture by G. C. Southworth and A. P. King of Bell Telephone Laboratories, which was reported for the New York meeting in the April issue, was presented.

March 2, 1939—H. J. Schrader, chairman, presiding.

Pittsburgh

A joint meeting was held with the Westinghouse Engineering Society to hear the paper on "Electron Optics," by V. K. Zworykin of the RCA Manufacturing Company. This was in the main the same paper as is reported for the Connecticut Valley Section is this issue.

W. R. Jones of the Hygrade Sylvania Corporation, presented a paper on "Electron Amplifiers."

The problems which the manufacturer of an expanding number of vacuum tubes presents were discussed. Specific tubes were then considered and the 1231 treated first. This tube makes possible amplification over a band five to six megacycles wide at ultra-high frequencies and was designed for use in television reception. However, it may be used as a conventional triode, tetrode, or pentode. Its characteristics were described and it was pointed out that high-μ tubes must be operated self-biased if distortion is to be avoided. The construction of the tube was described and illustrated by showing the various parts that comprise it and its complete form. A special wire acts as a contact pin, seal, and lead-in wire to which the elements are attached.

The paper was closed with a description of the 1.4-volt series of tubes for battery operation.

March 21, 1939—W. P. Place, chairman, presiding.

Portland

"Feedback Amplifiers" was the subject of a paper presented by F. E. Terman, Professor of Electrical Engineering at Stanford University. A report on this paper is given for its presentation at Seattle.

As a method of teaching safety, a rescue crew of the American Red Cross, under the direction of W. L. Dawson, gave a demonstration of artificial resuscitation.

March 29, 1939—H. C. Singleton, chairman, presiding.

San Francisco

"Television" was the subject of a paper by Chester Davis of the RCA Manufacturing Company. This paper is summarized in a report on the Los Angeles Section.

March 15, 1939—F. E. Terman, chairman, presiding.

Seattle

H. H. Willis, Chief Research Engineer of the Sperry Gyroscope Company, presented a paper on "Recent Developments in Aerial Navigation."

Dr. Willis discussed first the general problems of aerial navigation. The "automatic pilot" employs an indicating instrument which shows the direction of flight and the amounts of banking, turning, and climbing but does not locate the plane in space nor provide for blind landing.

Radio beacons locate the plane only with reference to a line except when immediately above the radiator in the "cone of silence."

A newly developed radio direction finder is completely automatic, its indicator showing constantly the direction of the station to which the receiver is tuned. The location of the ship can then be determined by triangulation measurements taken on two transmitters. In this system signals are picked up by a straight-wire antenna and by a loop. The loop signal is modulated by a 90-cycle voltage and delivered to the receiver. After passing through the receiver and a band-pass filter, the 90-cycle component is applied to a control circuit which turns the loop to a null-signal position.

A development was described which includes the features of the automatic pilot and a new blind-landing indicator. On the screen of a cathode-ray tube there is first traced a line horizontal with respect to the earth and controlled by the artificial-horizon gryo. A pair of spots are then projected on an imaginary line horizontal with respect to the airplane but shifted to the right or left under control of the rate-of-turn gyro. A centrally located spot moves up, down, or to either side under control of a new ultra-short-wave radio beam. In operation, if the pilot, after having entered the beam, keeps the beam spot in the center of the screen and the two rate-of-turn spots equidistant from the center with their center line coincident with the artificial horizon, the ship will land at a proper angle requiring neither turning nor banking

Four special "pancake" beams, two vertical and two nearly horizontal, are projected by electric horns of the type developed at Massachusetts Institute of

Technology and the Bell Telephone Laboratories. They mark the boundaries of the blind-landing path.

An ultra-high-frequency oscillator developed at Stanford University and suitable for generating substantial power at frequencies as high as 30,000 megacycles is used. It employs two rhumbatrons which are rigid toroidal-shaped copper shells forming oscillatory circuits having very high values of Q. By shooting a stream of electrons through the centers of two rhumbatrons in tandem and having them coupled by small dipoles, a momentary oscillation is set up in the second rhumbatron from the initial electron flow and is fed back to the first one. This feedback voltage increases or retards the speed of the electrons in the first rhumbatron which, when properly adjusted, results in sustained oscillations.

March 7, 1939—R. O. Bach, chairman, presiding.

"Feedback Amplifiers" was the subject of a paper by F. E. Terman, Executive Head of the Electrical Engineering Department at Stanford University.

It was shown both mathematically and experimentally that an amplifier employing negative feedback and having AB large compared with unity, where A is the gain before the feedback is applied and B is the fraction of the output fed back to the input, will have the following characteristics throughout the frequency range for which B is constant and the phase shift of AB is kept within certain limits: frequency distortion, amplitude distortion, and cross modulation will be greatly reduced, amplifier noises will be from slightly to greatly reduced according to where introduced, and the effective plate impedance will be decreased. However, the net gain also will be reduced and will be substantially equal to $1/B$, and practically independent of the original gain A, even, for example, with large changes in plate voltage.

The problems of applying feedback to single-stage, two-stage, and three-stage amplifiers and the design requirements in each case to prevent oscillation resulting from phase shifts of AB at the extreme frequencies were discussed. In general, such oscillation can be prevented by not allowing AB to be both positive and greater than unity.

The uses to which negative feedback can be applied appear to be unlimited. Some of them are: to improve frequency response and to reduce noise and amplitude distortion in audio-frequency amplifiers, to dampen speaker transients since the effective amplifier output impedance is lowered, to reduce cross modulation and harmonic distortion in radio-frequency amplifiers, to allow the use of cheaper power-supply equipment in transmitters and receivers since the hum can be greater and the voltage regulation poorer, to improve the balancing of a push-pull amplifier by equalizing the gains in each side, to improve the balancing of the resistance-coupled push-pull stage by stabilizing the gain of the associated phase inverter, to improve selectivity by having the feedback increase on each side of the desired frequency under control of a bridge circuit, to reduce the harmonics in a simple audio-frequency oscillator so that its output purity would compare favorably with that of a more expensive beat-frequency oscillator, and as a simple frequency equalizer incorporated in an amplifier for if the feedback network has the same characteristics as the distorting circuit, the equalization is automatic.

Dr. Terman also discussed current feedback in which the feedback voltage is produced by the output current rather than the output voltage. A three-stage vacuum-tube voltmeter employing current negative feedback was described which had a flat gain within two per cent over a frequency range of 30 to 120,000 cycles.

March 31, 1939—R. O. Bach, chairman, presiding.

Toronto

Beverly Dudley, associate editor of *Electronics*, presented a paper on "Analogies in Radio and Photography and Problems of Reproduction of Television Images."

The almost perfect analogy between the characteristics of a photographic film and a vacuum tube was demonstrated. It was shown that the slope of the plate-current—grid-voltage curve for fixed plate voltage decreased and became more linear as the load resistance was increased. In the photographic case, by plotting the logarithm of the exposure time against the density of the film for different values of developing time, it was shown that for infinite developing time the curves were almost identical to the vacuum-tube case when no load resistance was included in the tube circuit. Decreasing the developing time produced a reduction in the slope of the curve which increased its linear portion.

An analogy between the control-grid cutoff and the inertia of a film was pointed out as well as the fact that both show saturation if carried too far. This is particularly true of vacuum tubes employing tungsten filaments.

Analogies were shown between class B operation of vacuum tubes with its resulting increase in efficiency and distortion, and class C operation which carries these characteristics beyond the class B limits. In the photographic field the analogy was described by talking of underexposure where the latitude of contrast was considerably reduced by operation in an equivalent class B or class C region, the efficiency being improved at the expense of fidelity.

In television, it was pointed out that it is not desirable for the kinescope brightness characteristic to follow a linear relation with control-grid variation since the eye responds to brightness variations more in accordance with a logarithmic function. It is desirable, therefore, for the relation between the brightness on the screen and the control-grid voltage to follow a logarithmic exponential law. Amplifiers having such nonlinear characteristics may be used to obtain additional contrast or to reduce the contrast as necessary. Photographs of recent television pictures were shown and some of the problems encountered in taking them described.

March 27, 1939—R. C. Poulter, chairman, presiding.

Washington

"Precipitation-Static Interference in Aircraft and Ground Stations" was presented by H. M. Hucke, Communications Engineer for the Air Safety Board of the Civil Aeronautics Authority. This paper appears in full elsewhere in this issue.

March 13, 1939—Gerald C. Gross, chairman, presiding.

Two papers originally reported in the March Proceedings and given before the Connecticut Valley Section were presented. They were "Frequency and Amplitude Modulation Field Tests," by I. R. Weir of the General Electric Company (Schenectady), and "A New Frequency-Modulation Receiver," by G. W. Fyler of the General Electric Company (Bridgeport).

April 10, 1939—Gerald C. Gross, chairman, presiding.

Personal Mention

The following members have recently informed us of changes in their company affiliations or titles to those given below.

Adams, J. J.; Zenith Radio Corporation, Chicago, Ill.

Bernstein, H. E.; Lieutenant, U.S.N.; Department of Electrical Engineering and Physics, U. S. Naval Academy, Annapolis, Md.

Chun, H. H.; Arcturus Radio Tube Company, Newark, N. J.

Cogswell, W. P.; Lieutenant, U.S.N.; U.S.S. *Lexington*, New York, N. Y.

Cornell, J. I.; Solar Manufacturing Corporation, Bayonne, N. J.

Erstad, Johannes; Philips A/S, Copenhagen, Denmark.

Esperson, G. A.; National Union Radio Corporation, Newark, N. J.

Gregory, S. D.; National Broadcasting Company, Pittsburgh, Pa.

Hamvas, L. U.; Kenrad Lamp and Tube Company, Owensboro, Ky.

Hart, H. C.; Patent Lawyer, 30 Rockefeller Plaza, New York, N. Y.

Hendricks, P. S.; Columbia Broadcasting System, New York, N. Y.
Hodgson, A. D.; Signals Officer, Air Ministry, London, England.
Hollis, J. L.; Television Department, Crosley Corporation, Cincinnati, Ohio.
Kiebert, M. V., Jr.; Federal Communications Commission, Washington, D. C.
King, Y. C.; Central Radio Manufacturing Works, Changsha, Hunan, China.
Kohno, S.; Riken Sinkukogyo Company, Itakawa, Tiba-ken, Japan.
Lack, F. R.; Western Electric Company, New York, N. Y.
Landsberg, K. U.; National Broadcasting Company, New York, N. Y.
Lawlor, R. D.; Signal Corps Laboratories, Fort Monmouth, N. J.
Loucks, I. H.; Federal Communications Commission, Washington, D. C.
McGirr, W. P.; Lieutenant, U.S.N., District Communications Office, San Diego, Calif.
Mittlemann, Eugen; Consulting Engineer, 1118 North State Street, Chicago, Ill.
Miyakosi, Kazuo; Osaka Imperial University, Osaka, Japan.
Murrough, J. P.; Inspecting Engineer, Egyptian Government, Brussels, Belgium.
Passow, C. B.; Zenith Radio Corporation, Chicago, Ill.
Polydoroff, W. J.; Consulting Engineer, 2020 Engineering Building, Chicago, Ill.
Roder, Hans; Telefunken G.m.b.H., Berlin, Germany.
Sarault, G. E.; Canadian Broadcasting Corporation, Montreal, Que., Canada.
Sharpe, M. O.; Inspector, Federal Communications Commission, Galveston, Tex.
Sowley, H. A.; Telecommunication Department, General Post Office, Kingston, Jamaica, B.W.I.
Speir, F. H.; Southwestern Engineering Company of P. I., Inc., Manila, P. I.
Stannow, J. C.; Engineer, Nybrogatan 20, V, Stockholm, Sweden.
Tyrer, A. R.; Cable and Wireless Ltd., London, England.
Vasilevskis, Henriks; Valsts Elecktrotechniskā Fabrīka, Riga, Latvija.
Wolcott, C. F.; Thos. L. Siebenthaler Manufacturing Company, Kansas City, Mo.
Zottu, P. D.; Thermal Engineering Corporation, Richmond, Va.

Membership

The following indicated admissions and transfers of memberships have been approved by the Admissions Committee. Objections to any of these should reach the Institute office by not later than May 31, 1939.

Transfer to Fellow

Guy, R. F., National Broadcasting Co., 30 Rockefeller Plaza, New York, N. Y.

Transfer to Member

Bond, C. H., 103 W. Luray Ave., Alexandria, Va.
Eells, M. M., c/o Breeze Corp., Inc., 41 S. Sixth St., Newark, N. J.
Friend, A. W., 316 Forest Ave., Morgantown, W. Va.
George, E. E., 417 Power Bldg., Chattanooga, Tenn.
Giannini, G. M., Transducer Corp., 30 Rockefeller Plaza, New York, N. Y.
Hamburger, F., Jr., Johns Hopkins University, Baltimore, Md.
Irwin, G. J., Philco Products, Ltd., 1244 Dufferin St., Toronto, Ont., Canada.
Jacker, E. W., WCBD Transmitter, Church Rd., Elmhurst, Ill.
Janes, R. B., RCA Manufacturing Co., Inc., Harrison, N. J.
Jones, G. I., 2331 Cathedral Ave., N.W., Washington, D. C.
Kapus, E. E., J. B. Alberdi St. 2775, Buenos Aires, Argentina.
Khastgir, S. R., Physics Dept., Dacca University, Dacca, East Bengal, India.
Narayanan, P. L., c/o The Gramophone Co., Ltd., Box 118, Bombay, India.
Nimmcke, F. E., Bell Telephone Labs., Inc., 180 Varick St., New York, N. Y.
Osterbrock, W. C., University of Cincinnati, Cincinnati, Ohio.
Tyzzer, H. J., 2387 N. Bend Rd., Cincinnati, Ohio.
Wells, J. M., c/o American Optical Co., Southbridge, Mass.

Admission to Member

Bishop, J. B., 47 Oakland Ave., Bloomfield, N. J.
Dow, W. G., Elec. Eng. Dept., University of Michigan, Ann Arbor, Mich.
Hershey, L. M., 895 E. 95th St., Brooklyn, N. Y.

Admission to Associate (A), Junior (J), and Student (S)

Allen, W. G. D., (A) 403 Blowey-Henry Bldg., Edmonton, Alta., Canada.
Bourgonnier, C., (A) 15 rue Pergolese, Paris, France.
Carroll, D. V., (A) 922 Inverness Rd., Victoria, B. C., Canada.
Chelgren, A. E., (J) 2355 N. Eighth St., Milwaukee, Wis.
Conn, J. T., (S) 4134—12th Ave., N.E., Seattle, Wash.
Corderman, R. C., (A) 4401 Leland St., Chevy Chase, Md.
Dearing, J. B., (A) 161-08—84th Dr., Jamaica, L. I., N. Y.
Draper, C. A., (A) 4281 W. 30th St., Cleveland, Ohio.
Dunann, R. P., (S) 35 Sylvan Way, Piedmont, Calif.
Emigh, W. F., (S) Idaho Club, Moscow, Idaho.
Erdman, D. C., (S) 536 Berkeley Ave., Claremont, Calif.
Erlandson, P. M., (S) M.I.T. Dormitories, 413 Goodale, Cambridge, Mass.
Goette, C. A., (S) R.F.D. 2, Box 268, Woodland, Calif.
Gymer, A. F., (A) 6 Elizabeth Ave., Christchurch, Hants., England.
Heller, J. I., (A) 1732 E. 13th St., Brooklyn, N. Y.
Helwig, N. C., (A) R.F.D. 1, Box 243 C, Menlo Park, Calif.
Hexem, J., (S) Box 605, Ely, Nev.
Huffaker, I. R., (A) 310 Mathewson Pl., S.W., Atlanta, Ga.
Jesse, L. R., (A) 945 Hurt Bldg., Atlanta, Ga.
Jones, F. A., Jr., (S) 2429 Channing Way, Berkeley, Calif.
Joyce, M. W., (A) 21 Berston Ave., West Roxbury, Mass.
Kalmus, H. P., (A) Hotel Warrington, 161 Madison Ave., New York, N. Y.
Krishnan, L. M., (S) 410 Memorial Dr., Cambridge, Mass.
Leigh, J. S., (A) 211 Victoria St., Merchantville, N. J.
Lo Piparo, M., (A) Via Gran Sasso 28, Milan, Italy.
Luedtke, A., (S) Radio Station WRUF, Gainesville, Fla.
Maeder, A. S., (S) 16 Hurlcroft Ave., Medford, Mass.
Mahoney, W. P., (A) U.S.S. *Clark*, c/o Postmaster, New York, N. Y.
Mair, A. D., (A) California Arabian Standard Oil Co., Jidda, Arabia.
Martitsch, F., (A) 5 Schonbrunnerstrasse 75, Vienna, Germany.
Mason, M. A., (S) 2447 Derby St., Berkeley, Calif.
Mayer, R., (S) 107 E. Stewart Ave., Lansdowne, Pa.
Merchant, R., (S) Nassau, N. Y.
Mesak, C., (S) 2537 Ridge Rd., Berkeley, Calif.
Miller, W. E., (S) Bowles Hall, University of California, Berkeley, Calif.
Moon, B. F., (A) 1217 E. Windsor Rd., Glendale, Calif.
Nichandros, G., (S) 2624 Channing Way, Berkeley, Calif.
Noll, E. M., (J) Denver, Pa.
Olsson-Seffer, R. I., (A) Hede Soteri, Stenkullen, Goteborg, Sweden.
Palmer, V. L., (S) 4541 Brooklyn Ave., Seattle, Wash.
Pedersen, H. W., (A) Patrol Squadron 2, Coco Solo, Canal Zone.
Prager, J. H., (A) 3812 S. Hope St., Los Angeles, Calif.
Price, W. E., (A) 34-60—112th St., Corona, L. I., N. Y.
Radcliffe, F. E., (A) 157 Washington Ave., Chatham, N. J.

Ramachandra Rao, R., (A) Chennuru (P.G.), Via Gudlavalleru, Kistna Dist., Madras Presidency, India.
Rao, M. V. S., (A) Indian Radio & Television Corp., Box 1363, Madras, India.
Rosenblith, W. A., (A) 86 rue de l'Assomption, Paris 16e, France.
Shew, L. F., (S) 2968 Telegraph Ave., Oakland, Calif.
Shimizu, S., (A) Kawaguchi Hosojo, Kawaguchi-Shi, Saitama-Ken, Japan.
Sloan, C. B., (A) 3473 Mayfair Ave., Cheviot, Ohio.
Smith, A. E., (A) 544 Nettie Ave., Fort Wayne, Ind.
Snyder, R. F., (S) 215 E. Durham St., Philadelphia, Pa.
Spencer, H. S., (A) 131 Inwood Pl., Cincinnati, Ohio.
Stuedeman, W., (A) 2 White Ter., Newark, N. J.
Swingle, T. M., (A) 4009 Norwood Ave., Red Bank, Tenn.
Toshniwal, B. D., (S) Massachusetts Institute of Technology, Cambridge, Mass.
Varma Raja, M. V. R., (A) Power House, Trivandrum, Travancore, India.
Vasan, S. S., (A) Nariman Ter., Kohinoor Rd., Dadar, Bombay, India.
Whaley, C. D., (A) 52 Garden Pl., Brooklyn, N. Y.
Winans, R. C., (A) Bell Telephone Labs., Inc., 463 West St., New York, N. Y.

Books

The Engineers' Manual, by Ralph G. Hudson.

John Wiley & Sons, 440 Fourth Ave., New York, N. Y. 340 pages. 5 inches × 8 inches. Price $2.75.

For years Hudson's Manual has been the standby of a great number of engineers. Many of them first bought it at the suggestion of their instructors, others on the recommendation of their professional associates, for those who use it seem consistently to recommend it. One likes the trigonometric tables which are in decimal parts of a degree instead of minutes. Another will emphasize the convenience of the table of conversion factors which are arranged alphabetically in one inclusive list:

Multiply	by	to obtain
Abamperes	10	amperes
Years (leap)	8784	hours

The first edition appeared in 1917. Now a second edition has appeared. In its general form, and in much of its material, it does not differ greatly from the first edition. The first three sections, 67 pages on Mathematics, 74 pages on Mechanics, and 14 pages on Hydraulics, are almost unchanged.

The section on Heat, 33 pages, has been revised and rearranged. There is more emphasis on heating and ventilating, probably in recognition of the growing importance of this branch of engineering.

The section on Electricity has been expanded from 48 pages to 60 pages. Outstanding is the change from abhenrys, abamperes and ergs per second to henrys, amperes, and watts. It is unfortunate that gilberts, statvolts, and statcoulombs could not also have been superseded by the units the engineer uses. An addition is the set of formulas for equivalences of delta- and Y-connected impedances. This appears under the subject of three-phase circuits, and many readers will overlook the fact, which might well have been stated explicitly, that these equivalences also apply when the impedances are called T- or V-networks and are used in single-phase or even multifrequency systems. Under a new heading, "AC Power Transmission," formulas on this subject have been grouped and new material on transmission lines has been added.

The Mathematical Tables have been enlarged from 65 to 73 pages. This section includes, in addition to the usual mathematical tables, several tables of coefficients for hydraulics formulas and eleven short tables of properties of materials such as specific heat, heat of combustion, and electrical resistivity. This section has been enlarged mainly by a general expansion of these last.

The final section, conversion factors, covers 15 pages. It has been increased a little. The most significant change is the omission of monetary conversion factors!

The book is easy to use. It is principally a list of formulas, with a minimum of necessary explanatory matter. There is little fine print, while the headings and formulas are in rather large bold type. The first edition was noted for its freedom from errors. As to the new edition only use can tell. It was observed that circular mills are converted to square mills, and that the mathematical symbol for natural logarithm is *In*.

There are various omissions that might be pointed out. There are no formulas for permutations and combinations, nor for the vector calculus. Weights of materials are given only in pounds per cubic feet, even for xenon and platinum, while some materials such as helium and the modern plastics are omitted. The common logarithms are only to four places. But the answer of the typical user of Hudson's Manual is, "It has in it what I need. If you put all that other junk in it, it wouldn't be so simple and useful as it is now."

E. B. Ferrell
Bell Telephone Laboratories, Inc.
New York, N. Y.

The "Radio" Handbook, Fifth Edition, 1939, by the Editors of *Radio*.

Published by Radio, Ltd., 7460 Beverly Blvd., Los Angeles, California. 592 pages. 6 inches by 9 inches. Paper covers. $1.50 in Continental United States; elsewhere, $1.65 or 7s.

The "Radio" Handbook is particularly suited to the needs of the radio amateur. Sections on fundamental theory, vacuum tubes, and basic mathematics are followed by a particularly good section on the subject of antennas. Many new designs of transmitting and receiving apparatus are described, with sufficient detail for duplication by the amateur in his home workshop. In other sections are concisely treated such matters as test equipment, radio therapy, radio laws, and learning the code.

H. O. Peterson
R.C.A. Communications, Inc.
Riverhead, L. I., N. Y.

Correspondence

Time

In the March, 1939, issue of the Proceedings, in the Correspondence Section, the old question of G.C.T. versus G.M.T. was raised by J. B. Moore. Since this subject engaged my serious interest several years ago, I wonder if I might be permitted to present some of my views thereon?

First, I shall quote portions of a letter received from the International Bureau of the Telegraph Union, Berne, dated July 19, 1933.

"It is since the International Radiotelegraphic Conference of Washington, in 1927, that the term "Greenwich Mean Time" figures in the regulations annexed to the Radiotelegraphic Convention. The last Radiotelegraphic Conference of Madrid, in 1932, has maintained this term, which is accepted with few exceptions, throughout the entire world.

"We utilize G.M.T. in our nomenclatures, because this is the abbreviation which we must employ in the expressions of several provisions of the Regulations for Radio-communications.

"General Regulations for Radio-communications: art. 19; art. 23, sec. 4, D;

Appendix 3, note e) *Must be expressed by a group of 4 figures (0001 to 2400), Greenwich Mean Time; appendices 4, 5, and 6. Additional Regulation: art. 4, sec. 2. To indicate the hour of filing of radio-telegrams accepted in mobile stations, the rule is based on Greenwich Mean Time and utilizes the notation according to the dial of 24 hours. This hour is always expressed and transmitted by means of four figures (0001 to 2400).*

"As you state, the abbreviation G.M.T. is universally accepted in radio-communications, . . . For the moment, please understand that G.M.T. should signify, in radio-communications, without any possible doubt, the time commencing at midnight."

From the foregoing, it is evident that the term "Greenwich Mean Time" has been standard radio terminology for many years. There has never been any confusion in its use among radio men. Now it is suggested that we change from what we have been using these many years (G.M.T.) to something else (G.C.T.), merely because what we are using does not conform with what the astronomers use. With all due respect to the astronomers, wouldn't it be just as logical to expect *them* to adopt *our* terminology?

Mr. Moore uses as his authority for suggesting the change the *American Ephemeris and Nautical Almanac.* However, this publication is not the authority for the entire world. The *British Nautical Almanac* does not recognize the term "Greenwich Civil Time"; it does use "Greenwich Mean Time."

In 1922 the following notice appeared on the cover of the then current *British Nautical Almanac:*

"In both the abridged and complete Nautical Almanac the times styled G.M.T. are at present reckoned from noon, corresponding to 12 hours (Civil Time); but from the year 1925 inclusive and thenceforward the times styled G.M.T. in these publications will be given commencing at midnight, to conform with Civil Time; the term "Greenwich Mean Time" will then be considered to be the Standard Time of the Meridian of Greenwich, commencing at midnight and reckoned throughout the 24 hours."

It is seen that in the above quotation a very clear and concise definition of G.M.T. is given. There can be no uncertainty as to its reference point.

Prior to 1925, astronomers reckoned time from noon, so that the local time was identical with the mean sun's hour angle. By civil reckoning, the day began twelve hours earlier, at midnight. In 1925, by international agreement, it was decided that the astronomical reckoning should also be made from midnight. The two times thus became one and the same. Any confusion which now exists can be laid directly at the door of the astronomers. Instead of merely redefining the *hour* at which the day began, and leaving the *name* the same, as was done by the British authorities, they brought into use the new term "G.C.T."

We are not measuring a new kind of time when we say "G.C.T." Since Mean Time is time measured by the apparent motion of the mean sun, it is evident that we are still using the same mean time, as of old. There is no necessity for having two different names for the same kind of time. The term "Greenwich Mean Time," being accurately descriptive, is, without doubt, the more logical term to retain.

"G.M.T." is universally accepted and used in radio communications. Notwithstanding the fact that the *American Ephemeris and Nautical Almanac* employs "G.C.T." in its tables, it appears that "G.M.T." is the term actually used by seafaring men. A search of technical publications will reveal that astronomers are practically the only people that now use "G.C.T." Since they comprise only a small portion of the many that use time referred to the meridian of Greenwich, would the suggested change in terms eliminate confusion, or would it only tend to create more?

Arthur M. Braaten
Box 626
Riverhead, L. I., N. Y.

Contributors

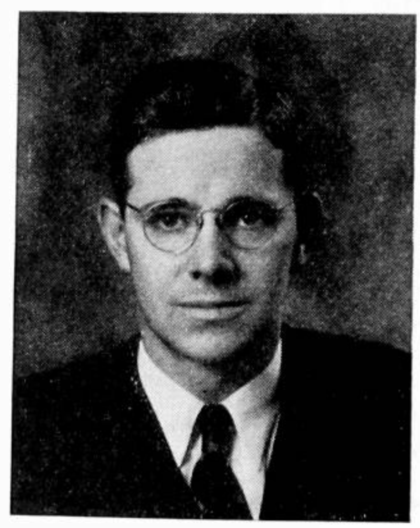
Joseph G. Chaffee

Joseph G. Chaffee (A'26) was born on March 6, 1901, at Hackensack, New Jersey. He received the S.B. degree from Massachusetts Institute of Technology in 1923. From 1923 to 1925 he was in the engineering department of the Western Electric Company, and since 1925 he has been with the Bell Telephone Laboratories.

❖

Herbert M. Hucke (A'31) was born at Woodland, California, on September 26, 1904. In 1927 he received the B.S. degree in electrical engineering from the Polytechnic College of Engineering at Oakland, California. From 1925 to 1931, Mr. Hucke was in the Sales and Engineering Products Division of the Radio Corporation of America. He was with the United Air Lines working on aircraft receiver design from 1931 to 1934, and from 1934 to 1938 he was Chief Communications Engineer. In October of 1938 he became Communications Engineer for the Air Safety Board of the Civil Aeronautics Authority at Washington, D. C.

❖



Herbert M. Hucke

For biographical sketches of T. R. Gilliland, S. S. Kirby, and Newbern Smith see the Proceedings for January, 1939.

Commercial Engineering Developments

These reports on engineering developments in the commercial field have been prepared solely on the basis of information received from the firms referred to in each item.

Sponsors of new developments are invited to submit descriptions on which future reports may be based. To be of greatest usefulness, these should summarize, with as much detail as is practical, the novel engineering features of the design. Address: Editor, Proceedings of the I.R.E., 330 West 42nd Street, New York, New York.

●

New Equipment for the National Park Service

A headquarters type set furnishing 50 watts of unmodulated carrier output has recently been placed in operation by the National Park Service* at Mount McKinley National Park, Alaska.

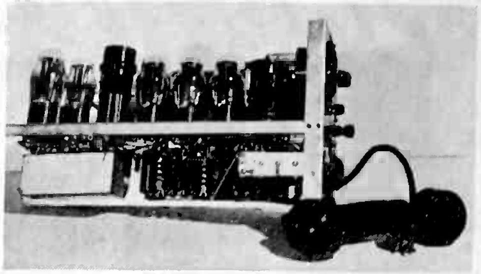
Requirements for a complete and compact station that could be placed on the desk of a chief ranger or fire dispatcher were met by careful selection of component parts and circuit arrangement. Unforeseen overloading and long continuous periods of operation had to be considered in design in view of the fact the set was going to the interior of Alaska to be operated by untrained personnel, and the nearest radio technician after the installation was completed would be 300 miles away. Acceptance tests required an 8-hour run under full modulation without exceeding specified heat rises in transformers, chokes, etc., thereby giving assurance against underrating of parts.

Special features included an antenna-patching panel to allow matching impedances of 72- and 600-ohm lines and a loaded end-fed antenna. The speech input as well as the receiver output included windings for matching a 600-ohm open-wire line. Band switching to three frequencies—2670, 3235, and 3410 kilocycles—for emergency use was effected by selection of different variable condensers with shaft lock nuts, thereby eliminating tapped coils and giving a much stronger and neater type of construction. The difference in the frequencies was not great, so an effectual reduction in *LC* ratio was not noted.

The receiver is conventional with the inclusion of recent development features applicable to the communication-type receiver.

A field set used on forest fires and temporary stations in remote sections of a national park is shown in the accompanying photographs. This set is complete in itself including power supply and antenna. Three equal-voltage dry-battery supplies, differing only in milliampere-hour capacity, are available for use with the set. Choice depends on the type of country to be covered and the amount of use the set will be given. The set and power supply combinations weigh, respectively, 21 pounds for 7-hour, 36 pounds for 150-hour, and 96 pounds for 70-day continuous use.

The transmitter is crystal controlled for operating between 2496 and 3415 kilocycles and uses a type 30 tube in the oscillator stage which in turn excites 2 30-type tubes in parallel for the power amplifier. The output circuit can be connected by a toggle switch for either a 72- or 600-ohm load, and the unmodulated power output is approximately 3 watts. Audio power is derived from a single-button carbon microphone into a 1F4 which drives a 19 class-B modulator.



National Park Service field set

The receiver employs a 1A6 mixer tube, a 1A4 iron-core intermediate, and a 1B5, which has one diode connected as the second detector, the other diode as an automatic-volume-control-voltage source, and the triode section as the audio amplifier. The power output levels off at 50 milliwatts which is more than ample for earphones or a small permanent-magnet loudspeaker.

The sensitivity standard required for these receivers is 0.9 volt root-mean-square on a vacuum-tube voltmeter across the audio output shunted by a resistance equal to the earphone. By calculation this is an equivalent output of 50 milliwatts were there a 1F4 audio amplifier. Eleven microvolts input gives this standard output and readable signals can be maintained down to less than one microvolt input.

This type of field set, including a few improvements from time to time, has been in use in the National Parks for the past two years and has proved entirely satisfactory both electrically and mechanically for the service it is required to furnish.

* U. S. Department of the Interior, National Park Service, Washington, D. C.—Oliver G. Taylor, Chief of Engineering; W. C. Hilgedick (A'29) Radio Engineer.

Headquarters-type transmitter and receiver of the National Park Service at Mount McKinley National Park, Alaska

Single-Adjustment Push-Button Tuner

A push-button tuner in which the tuning of two circuits can be varied simultaneously by a single adjustment has been developed by Sprague Specialties Company.* This permits the antenna and oscillator condensers in a superheterodyne receiver to be adjusted without, in many cases, making necessary the use of a separ-

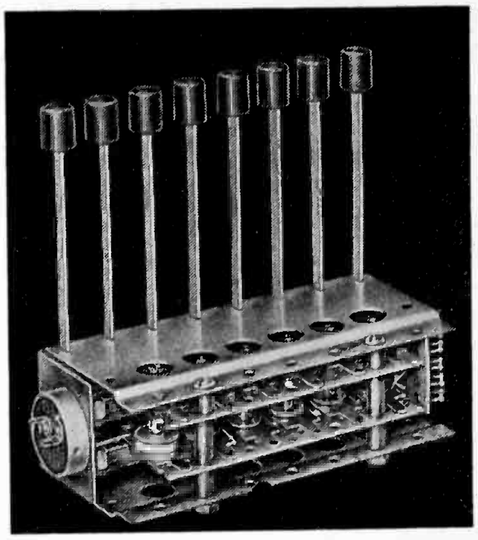
* Sprague Specialties Company, North Adams, Massachusetts.

You are cordially invited to visit the Bell System exhibit at Golden Gate International Exposition, San Francisco

(*Continued from page ii*)

ate aligning oscillator when the receiver is set up in the home.

The assembly uses pairs of compression-type mica condensers, each mounted on a separate deck. Reasonably close tracking



One adjustment controls the frequency of both the antenna and oscillator circuits in this 2-circuit push-button tuner assembly

is obtained. In a typical unit the maximum deviation from equal capacitance was 17 micromicrofarads for a capacitance setting of 150 micromicrofarads. Where exact tracking is required a vernier compensation adjustment for the antenna-tuning unit is provided.

Special precautions to insure stability in the face of temperature and humidity changes have been taken. Design of the plate shape and selection of materials are reported to result in a combination of elements which almost entirely eliminates capacitance changes resulting from temperature, mechanical instability, and setting drift (i.e., the permanent set taken by a trimmer after initial setting). Complete assemblies are treated with a material which breaks up moisture paths and, without interfering in any way with the ability to adjust the trimmer within a wide capacitance range, eliminates humidity drift. When applied to a reasonably stable receiver, total drift is held to within 2 to $2\frac{1}{2}$ kilocycles over the entire broadcast-frequency range.

Ultra-High-Frequency Power Tube

A new ultra-high-frequency vacuum tube designed for operation in the range from 30 to 300 megacycles, has just been introduced by Western Electric.* This tube, the 356A, utilizes the stemless type of construction† used in earlier tubes developed at the Bell Telephone Laboratories.

The tube is a filamentary, air-cooled, high-mu triode. As a radio-frequency oscillator it may be used at the full 50-watt rating up to 100 megacycles and at reduced

* Western Electric Company, 195 Broadway, New York, New York.

† C. E. Fay and A. L. Samuel, "Vacuum tubes for generating frequencies above one hundred megacycles," Proc. I.R.E., vol. 23, pp. 199–212, March, (1935).

(Continued from page v)

rating up to 250 and even 300 megacycles. It is also suitable for use at audio frequencies, particularly in Class-B audio amplifiers or modulators, where it may be used without grid bias at plate potentials as high as 1000 volts.

All of the tube elements are supported by the leads, which are themselves supported solely by the glass envelope. The plate terminal at the top of the tube is connected to its lead by a welded connection. The tube envelope is made of Nonex glass.

The leads at the base of the tube are connected externally to a special 4-prong, wafer type base made of lavite and designed for use in the Bell System's standard 4-terminal socket.

New Western Electric power tube for ultra-high-frequency applications

Basic Characteristics

Filament: $E_f = 5.0$ v, $I_f = 5.0$ a, a.c. or d.c.
Average thermionic emission: $I_s = 1.0$ a.
Average characteristics with $I_b = 100$ ma: $\mu = 50$; $g_{gp} = g_m = 3800$ μmhos; $r_p = 13{,}000$ ohms.
Average direct interelectrode capacitances $C_{gp} = 2.75$ μμf, $C_{pk} = 1.0$ μμf, $C_{gk} = 2.25$ μμf.
Maximum ratings for maximum frequency of 100 Mc: $E_b = 1500$ v, $I_b = 120$ ma, $P_p = 50$ watts; $(I_g)_{rf} = 6$ a, $I_c = 35$ ma.
Maximum plate voltage for upper frequency limit of 250 Mc: $E_b = 1000$ v.
Connections: Plate, to cap on glass; grid, filament, and filament center-tap, to Western Electric 4-pin base.

Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements.

A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I. R. E.

* ANTENNEN: Ihre Theorie und Technik (Antenna Theory and Engineering). By DR. ING. HELLMUT BRÜCKMANN, Staff Member in the German Postal Administration. Leipzig: S. Hirzel, February, 1939. xiv+339 pages+2 charts, illustrated, 6×9 inches, paper. 20.50 rm.

* BBC HANDBOOK, 1939. London: The British Broadcasting Corporation, 1939. 176 pages, illustrated, 5×7½ inches, cloth. 2 shillings.

* EINFÜHRUNG IN DIE SIEBSCHALTUNGSTHEORIE DER ELEKTRISCHEN NACHRICHTENTECHNIK (Introduction to the Theory of Filters for Electrical Communication). By DR. R. FELDTKELLER, Ord. Professor and Director of the Institutes for Communications, Technische Hochschule, Stuttgart. Leipzig: S. Hirzel, January, 1939. x+174 pages, illustrated, 6×9 inches, paper. 10.80 rm.

* ELECTROLYTIC CONDENSERS: Their Properties, Design, and Practical Uses. By PHILIP COURSEY, Technical Director, Dubilier Condenser Company, Ltd., London. New York: John F. Rider, 1938. 164+8 index pages, illustrated, cloth. $3.00.

* ELECTRON OPTICS. By the Research Staff of Electric and Musical Industries, Limited; compiled and written by OTTO KLEMPERER. Cambridge: at The University Press; New York: The Macmillan Company, February 28, 1939. x+107 pages, illustrated, 5½×8½ inches, paper. $1.75.

* HIGH-FREQUENCY ALTERNATING CURRENTS, Second Edition. By KNOX MC ILWAIN and J. G. BRAINERD, Moore School of Electrical Engineering, University of Pennsylvania. New York: John Wiley & Sons, 1939. 484+25 appendix +20 index pages, illustrated, 6×9 inches, cloth. $6.00.

INTRODUCTION TO CONTEMPORARY PHYSICS, Second Edition. By KARL K. DARROW, Ph.D., Research Physicist, Bell Telephone Laboratories, Inc. New York: D. Van Nostrand Company, Inc., 1939. 648 pages, illustrated, 6¼×9¼ inches, cloth. $7.00.

* SPONTANEOUS FLUCTUATIONS OF VOLTAGE, Due to Brownian Motions of Electricity, Shot Effect, and Kindred Phenomena. By E. B. MOULLIN, Fellow of Magdalen College, Donald Pollock Reader in Engineering Sciences in the University of Oxford. London and New York: Oxford University Press, January 26, 1939. 248+3 index pages, 6×10 inches, cloth. $6.00.

* THEORY AND APPLICATIONS OF ELECTRON TUBES. By HERBERT J. REICH, Associate Professor of Electrical Engineering, University of Illinois. New York: McGraw-Hill Book Company, Inc., January, 1939. xviii+631+11 appendix+26 index pages, illustrated, 6×9 inches, cloth. $5.00.

Associate membership affiliates you with the Institute and brings you the PROCEEDINGS each month as well as notices of meetings held near you.

(Typewriting preferred in filling in this form) No..........

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

Name ....
(Give full name, last name first)

Present Occupation ....
(Title and name of concern)

Business Address ....

Home Address ....

Place of Birth .... Date of Birth .... Age ....

Education ....

Degree ....
(College) (Date received)

TRAINING AND PROFESSIONAL EXPERIENCE

(Give dates and type of work, including details of present activities)

Record may be continued on other sheets of this size is space is insufficient.

Receipt Acknowledged .... Elected .... Notified ....

Requirements—For Associate membership, an applicant must be at least twenty-one years of age, of good character, and be interested in or connected with the study or application of radio science or the radio arts.

Sponsors—Three sponsors who are familiar with the work of the applicant must be named. Preferably these should be Associates, Members, or Fellows of the Institute. In cases where the applicant is so located as not to be known to the required number of member sponsors, the names of responsible nonmember sponsors may be given.

Dues—Dues for Associate membership are six dollars per year. The entrance fee for this grade is three dollars and should accompany the application.

Other Grades—Those who are between the ages of sixteen and twenty-one may apply for Junior grade. Student membership is available to full-time students in engineering or science courses in colleges granting degrees as a result of four-year courses. Member grade is open to older engineers with several years of experience. Information on these grades may be obtained from the Institute.

Western Electric

EQUIPPED FOR BETTER BROADCASTING

These six words convincingly testify to a station's ability to provide the finest technical facilities for the job of broadcasting.

Distributed by Graybar Electric Company

DISTRIBUTORS: Graybar Electric Co., Graybar Building, New York. In Canada and Newfoundland: Northern Electric Co., Ltd. In other countries: International Standard Electric Corp.

ENGINEERING DIRECTORY

Consultants, Patent Attorneys, Laboratory Services

BRUNSON S. McCUTCHEN and CHARLES B. AIKEN
Consulting Engineers

Technical cooperation with Attorneys in connection with patent litigation—Design and Development work—Audio and radio frequency measurements—Equipment studies—Receiver and transmitter problems—A well equipped laboratory.

75 West Street New York City — Telephone WHitehall 4-7275

Specialist in
AMPLIFIER DESIGN

engineering and construction. Complete facilities for developing any type of industrial, commercial, or professional amplifier equipment, audio filters, transformers, reactors, power supplies, etc.

N. M. HAYNES
17-19 West 20th Street New York, N.Y. — Telephone Watkins 9-0916

Use this directory . . .

when you need consulting services

when you are asked to suggest the name of a specialist on an engineering or patent problem

Consulting Engineers, Patent Attorneys, Laboratory Services . . . Applications for card space in the Engineering Directory are invited. Complete data will be sent on request to

PROCEEDINGS of the I.R.E.
330 West 42nd Street New York, N.Y.

INDEX

DISPLAY ADVERTISERS

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open on April 20. Make your application in writing and address it to

Box No. . . . . . .
PROCEEDINGS of the I.R.E.
330 West 42nd Street, New York, N.Y.

Please be sure that the envelope carries your name and address

DESIGN ENGINEERS

The Proceedings is receiving an increasing number of requests from employers for engineers with *commercial—design experience* on various specialized types of radio-communication equipment: receivers and transmitters, audio and video, for the broadcast, police, and aircraft services.

If you are qualified and are interested in openings of this type, we suggest that you file a statement of your education and experience record with I.R.E. headquarters. Write to Box 200 and request a copy of the "I.R.E. Employment Record Form."

TRANSMITTER DEVELOPMENT

Graduate engineer wanted for radio transmitter development and design work. Practical experience essential. State salary expected. Must be British subject. Location Canada. Box No. 195.

CRYSTAL-MANUFACTURING ENGINEER

thoroughly familiar with all phases of modern quartz crystal manufacture—cutting, grinding, processing—wanted immediately. Apply in writing to I.R.E. Box 196, giving age, experience and salary desired.

TRANSFORMER DESIGNER

A young engineer, college graduate, with several years of transformer-design experience is wanted by a well-known manufacturer in the East. The present staff knows of this opening. Box 197.

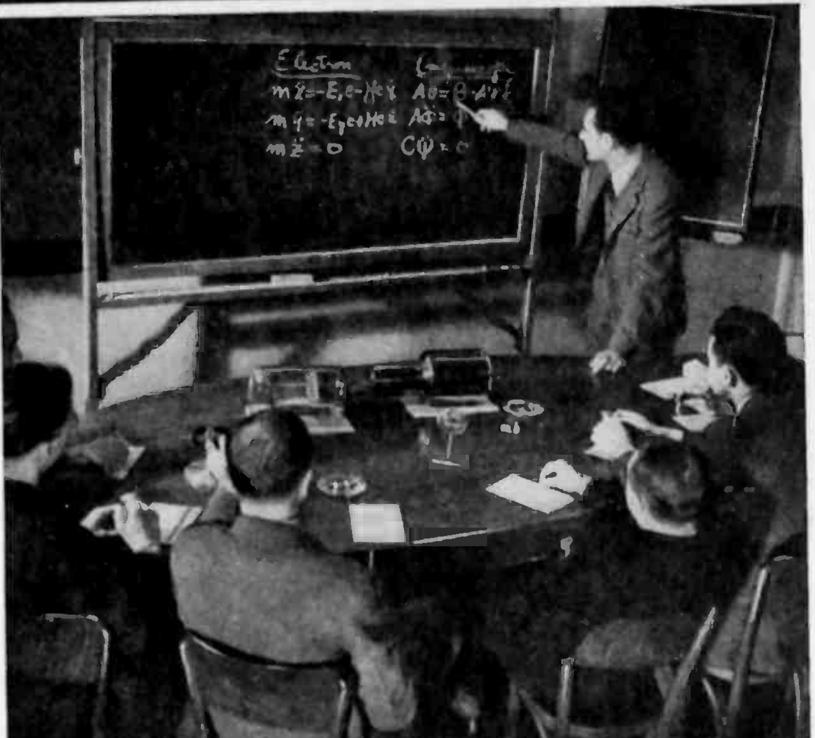
Attention Employers . . .

Announcements for "Positions Open" are accepted without charge from employers offering salaried employment of engineering grade to I.R.E. members. Please supply complete information and indicate which details should be treated as confidential. Address: "POSITIONS OPEN," Institute of Radio Engineers, 330 West 42nd Street, New York, N.Y.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal

Skillful Engineering Foundation of Success!

'ROUND THE WORLD in 3 days and 19 hours! This remarkable feat was accomplished less than a year ago by an intrepid young flyer whose skillful engineering of the entire flight—from the time the plan was conceived, until the swift air journey was over—was responsible for its amazing success.



TELEVISION on April 30th became an established, practical fact in New York. And Television, too, is the result of skillful engineering—much of it in the RCA tube plant at Harrison, N. J., where men who live, eat, and sleep radio, spent years in research and development perfecting the television tubes that have played so important a part in the solution of a complex problem.

RCA Radiotrons Offer You the World's Most Skillful Tube Engineering

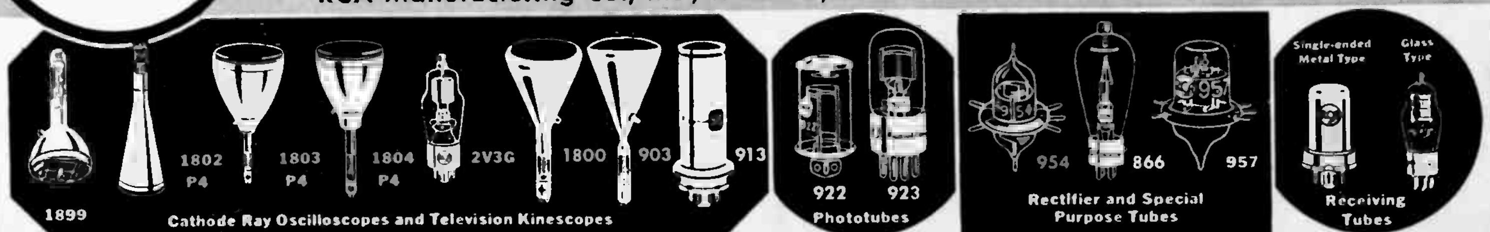
Just as skillful engineering and careful research were prime requisites in the record-breaking 'Round the World Flight and in the development of television, so skillful engineering is necessary for the creation of receiving set tubes.

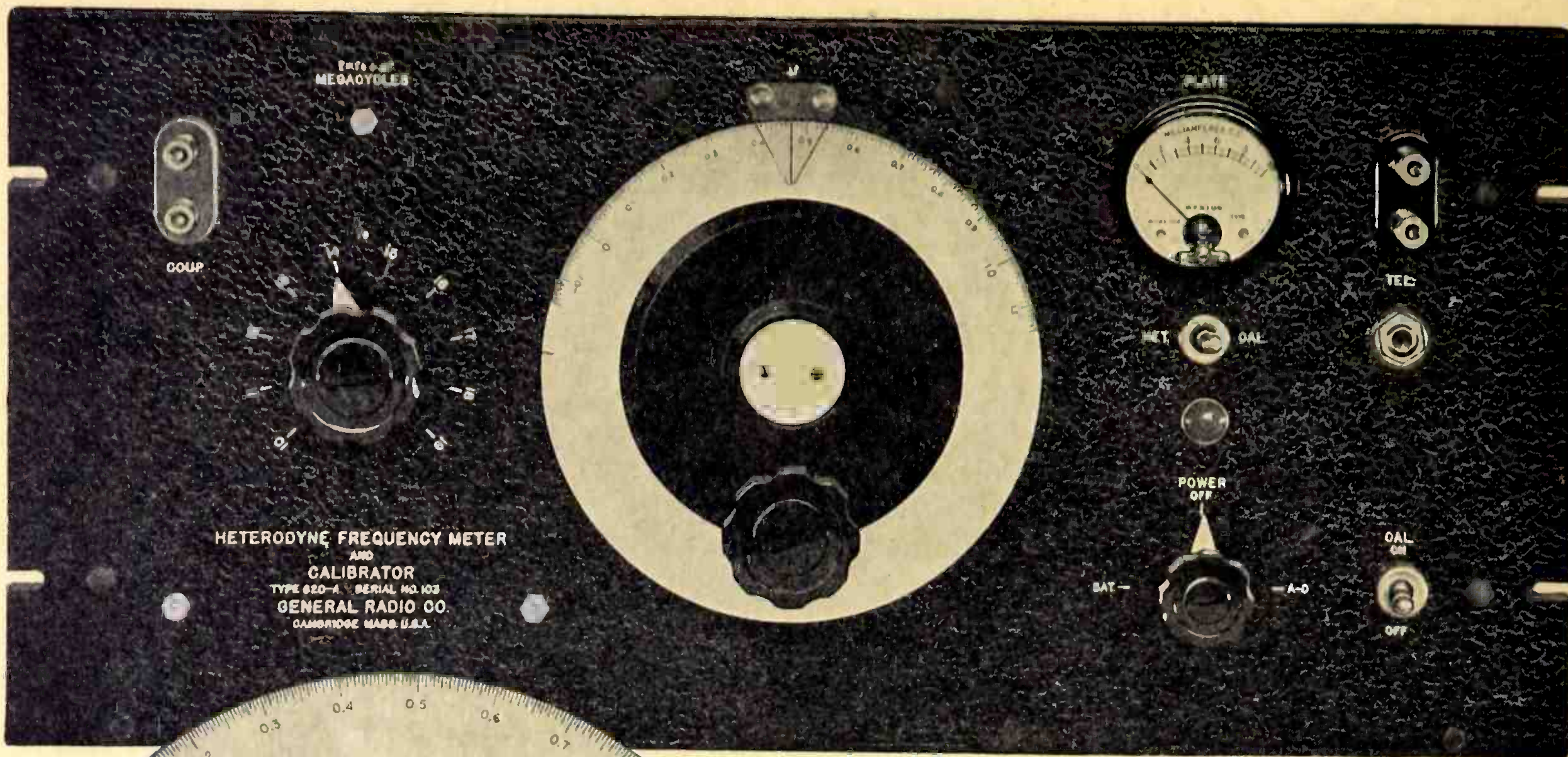
RCA is the only company whose activities cover all phases of radio. Hence RCA naturally builds truly fine radio tubes. And since one of the most important parts of the radios you build is tubes, it stands to reason that it will pay you to use RCA Radiotrons in your instruments. Give your own skillful engineering a "plus" by adding the product of the world's most skillful tube engineers—RCA Radiotrons!

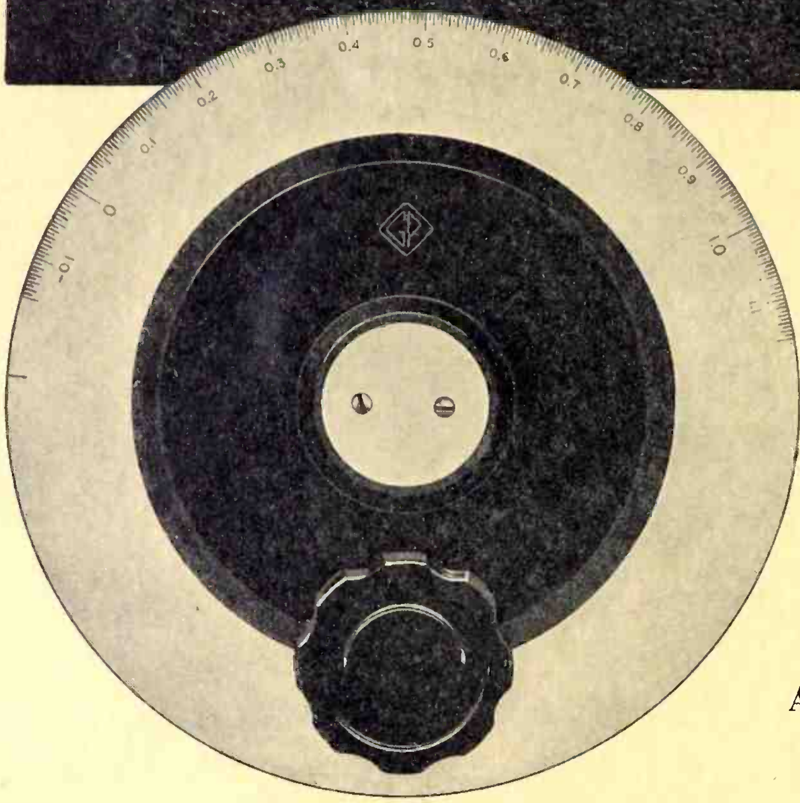
Trademark "Radiotron" Reg. U. S. Pat. Off. by RCA Manufacturing Company, Inc.

RCA Radiotrons

RCA Manufacturing Co., Inc., Harrison, N. J. • A Service of the Radio Corporation of America







for

TELEVISION *and* *other* Ultra-High-Frequency Services

ACCURATE · SIMPLE TO OPERATE

WIDE FREQUENCY RANGE

THE General Radio Type 620-A Heterodyne Frequency Meter and Calibrator offers a number of advantages for frequency measurements in the high- and ultra-high frequency bands.

1. **HIGH UPPER-FREQUENCY LIMIT—satisfactory to 300 Mc**
2. **VERY WIDE RANGE—300 kc to 300 Mc**
3. **DIRECT READING—no charts or curves**
4. **SELF-CALIBRATING—calibration can be checked instantly from built-in piezo-electric oscillator**
5. **NO COILS TO CHANGE—built-in coil system with 10-point panel selector switch**
6. **GOOD ACCURACY—over-all is 0.01% or better—can be used for monitoring frequencies of transmitters if allowable tolerance is 0.02% or better**
7. **FLEXIBLE POWER SUPPLY—built-in a-c operated power supply or external d-c source, either selected by panel switch**

This instrument is finding wide application in the communications field as well as in the laboratories of receiver and transmitter manufacturers where a very wide range of frequencies has to be measured simply, accurately and quickly.

TYPE 620-AR Relay Rack Model $490.00

TYPE 620-AM Portable Model $555.00

WRITE FOR BULLETIN 411 FOR COMPLETE DATA

GENERAL RADIO COMPANY, CAMBRIDGE, MASS.

Branches in New York and Los Angeles