Toth

VOLUME 25 SEPTEMBER, 1937 NUMBER 9

PROCEEDINGS
of
The Institute of Radio Engineers

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Institute of Radio Engineers
Forthcoming Meetings

ROCHESTER FALL MEETING
Sagamore Hotel
Rochester, New York
November 8, 9, and 10, 1937

SEATTLE SECTION
September 24, 1937

WASHINGTON SECTION
September 13, 1937

TOTH

PROCEEDINGS OF

The Institute of Radio Engineers

VOLUME 25 — September, 1937 — NUMBER 9

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The Institute of Radio Engineers

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Proceedings of the Institute of Radio Engineers
Volume 25, Number 9 *September, 1937*

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INSTITUTE NEWS AND RADIO NOTES

Institute Meetings

Buffalo-Niagara Section

On June 23 a meeting of the Buffalo-Niagara Section was held at the University of Buffalo with G. C. Crom, chairman, presiding. There were thirty-three present.

Two papers were presented by engineers at the National Broadcasting Company. The first on "N.B.C. Directional and Nondirectional Antenna Development," was by R. F. Guy. The development of antennas from those used by Marconi to present types was first presented. Field intensity patterns were described and it was pointed out that public response from the operation of stations indicated close agreement with the predictions resulting from field intensity measurements. Improvements have been mostly in the pattern of radiation rather than in the efficiency. The paper was closed with a description of the antenna to be operated with the new 500-Kilowatt transmitter to be installed at WJZ.

The second paper was by W. S. Duttera and covered "Some Factors in Design of Directive Broadcast Antenna." He outlined first an original theoretical method of analyzing the performance of a directional antenna system by determining the relative power radiated in equally distant zones at all angles above the earth. Directional systems composed of two antennas were considered although the analysis may be extended to more complicated arrays. It was shown that with short antennas, certain spacings and phasings accentuate the undesirable high angle radiation characteristics of a single element and other spacings and phasings materially improve the characteristics for broadcast purposes. Directional systems comprised of two radiators were considered in detail for 90- and 190-degree radiators having equal currents. A number of those present participated in the discussion of these papers.

Emporium

On June 25 and 26 the Emporium Section held its second annual summer meeting. Four technical papers were presented and over a dozen out-of-town guests were present.

The evening technical session on the 25th, presided over by M. I. Kahl, chairman, was attended by eighty and was opened with the showing of a talking motion picture describing the operation and manufacture of several types of Eveready batteries. L. M. Temple of the National Carbon Company then spoke on "The Importance of Power Efficiency in Battery Receivers." Numerous curves were shown of sensitivity and distortion as functions of battery life and power output for various typical battery operated receivers. Improvements in power efficiency could be obtained by more efficient tubes or the use of over-bias switches permitting reduction of power input when conditions permitted. Curves indicated that battery life could be increased as much as sixty-seven per cent by the use of over-bias. The paper was discussed by a number of those present.

The second speaker of the evening was Lincoln Walsh, consulting engineer, who discussed "The Design of Radio Receivers for High Fidelity Reproduction." Features discussed included separation of the automatic volume control circuit from the diode detectors to prevent overloading of the detector, variable intermediate-frequency coupling, and the use of ten-kilocycle audio-frequency filters to prevent inter-station heterodyning. A demonstration was then presented of a receiver capable of practically flat response up to sixteen kilocycles with means for narrowing this band when conditions demand. The most effective demonstration employed high fidelity phonograph records to modulate a signal generator, the output of which was coupled to the receiver. Several of those present participated in the discussion of the paper.

On the morning of the 26th, forty members and guests attended the meeting which was presided over by M. I. Kahl, chairman. E. F. Carter of the Hygrade Sylvania Corporation presented a paper on "An Introspection of Engineering Organization." In it an attempt was made to evaluate some of the human aspects of engineering. The superior results obtainable under a staff type of organization as compared with a strictly military type when dealing with highly trained men of varying temperament, ideals, and attitudes was pointed out. The value of co-operation was discussed as were methods of obtaining it. Personal deficiencies such as excessive pride, stubbornness, and destructive criticism were explored to show their disadvantages and methods needed to minimize their disastrous effects on intelligent co-operation. Compromise was considered a useful means of alleviating these conditions except where fundamental principles or policies were jeopardized. Other means discussed included an appreciation of the other fellow's problems, the creation of confidence in others in their own ability, and the use of frequent contacts between groups working on the same or similar lines. A discussion of this paper followed.

The final technical paper was presented by C. J. Franks of the Ferris Instrument Company, and was on "Recent Advances in Signal

Generator Design." It was demonstrated by means of a cathode-ray oscilloscope which showed the wave forms of the signal generator output. The signal generator demonstrated contained a regulated power supply. Its range of radio frequencies was from fifty kilocycles to twenty-five megacycles. Modulation was achieved by simultaneously varying the plate and screen grid voltages of the power amplifier. The tuning scale of the radio-frequency unit is equivalent to thirty feet in length and provision is made for direct reading calibration. By use of calibration charts, the frequency may be adjusted to within one-half of one per cent of a desired value.

The audio-frequency system uses a 400-cycle oscillator which through a switching arrangement may be supplanted by an external modulating source. When used with external variable frequency sources the amplifier is capable of one hundred per cent modulation and very low distortion values at frequencies up to 12,000 cycles. The generator output is transferred to the receiver under test by means of a concentric transmission line terminated in its characteristic impedance.

The afternoon was given over to outdoor sports at a camp site in Rich Valley. A picnic supper was served and the meeting was terminated by darkness.

Indianapolis

The first regular meeting of the Indianapolis Section was held at the Indianapolis Athletic Club on January 22nd. Because of the unfavorable weather, which sharply restricted transportation, there were only twenty-eight members present. I. M. Slater, temporary secretary, presided in place of A. D. Silver, temporary chairman. Because of the small attendance it was agreed not to hold the election for permanent officers.

H. P. Westman, national secretary, outlined the history, aims and activities of the Institute.

A paper on "Oscillators for Superheterodyne Receivers" was presented by V. C. MacNabb, chief engineer of Fairbanks-Morse Corporation, Home Appliance Division. Various types of oscillator circuits were described and the advantages and disadvantages of each outlined. The paper was discussed by Messrs. French and Mallory.

Pittsburgh

On June 29 the annual dinner meeting and election of officers of the Pittsburgh Section, which was attended by eighteen, was held at the Villa D'Este, with B. Lazich, presiding. In the election of officers R. T. Gabler of the Carnegie Institute of Technology was named chairman, W. P. Place, of the Union Switch and Signal Company, was elected vice-chairman, and A. F. Shreve of the Equitable Sales Company was named secretary-treasurer. After the short business session, and addresses by the newly elected and retiring officers, the meeting adjourned to a nearby room where several reels of motion pictures were projected.

San Francisco

The May 12 meeting of the San Francisco Section was held in the Downtown Association Meeting Rooms with Noel Eldred, vice chairman, presiding. This was a seminar meeting and H. E. Metcalf of Lippincott & Metcalf led a review of the paper by L. A. Kubetsky on "Television Multiple Amplifiers," which appeared in the April, 1937, Proceedings. Harry Greene, chief design engineer of the Remler Manufacturing Company led a discussion of the paper by D. E. Foster and S. W. Seeley on "Automatic Tuning, Simplified Circuits and Design Practice," which appeared in the March, 1937, issue.

On June 16 V. C. Freiermuth, chairman, presided at a meeting of the San Francisco Section which was held in the Pacific Telephone and Telegraph Company's auditorium and attended by thirty-four.

A paper on "The Production and Measurement of High Vacua and Its Relation to the Field of Electronics" was presented by H. W. Lindsay, vacuum technician and instrument designer for the Shell Development Company. In it he covered the historical experiments and devices for obtaining reduced gas pressure. He then described construction and operation of modern mechanical and diffusion pumps, and outlined current practice in this field.

Proceedings of the Institute of Radio Engineers
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TECHNICAL PAPERS

THE ORIGIN AND DEVELOPMENT OF RADIOTELEPHONY*

By

Lloyd Espenschied

(Bell Telephone Laboratories, Inc., New York City)

Summary—*Upon this, the Silver Anniversary of the Institute of Radio Engineers, it is appropriate to recall how there came into being the art of radiotelephony and, in turn, such services as overseas telephony and broadcasting. The Institute has seen the entire evolution within its relatively short life, with radiotelephony an unsolved problem in 1912 and today an accomplished fact of world-wide application.*

The pages of the Institute Proceedings *testify to much of the building of the art, but nowhere has there been given a unified account of the structure as a whole and the relation of its technical substance to electric communications generally. To do this objectively and while the development is still fresh in mind is the purpose of the present paper. Naturally, the story is limited by space and by the information available to the writer.† Most of the account pertains to America. If the contributions of other countries are not adequately presented, it is because the limitations of time, space, and language have not yet been entirely overcome.*

Background in the Physical Sciences

IT IS well to acknowledge, in the first place, the debt which radio owes to the more fundamental contributions from the physical sciences, the more pertinent ones of which, underlying as they do the entire art of electric communications, may be epitomized as three major waves of advance:

The great transition which occurred in the early 1800's from the electrostatic to the electric current and electromagnetic state of electrical science, which led to the telegraph and later to the telephone (to say nothing of electric power).

The conception and demonstration of electromagnetic wave propagation and electric oscillations, notably by Maxwell and Hertz. This advance applied to guided as well as unguided wave propagation, and is the basis of the transmission art of both wire and wireless communication.

The proof of the corpuscular nature of electricity and its identity with matter, the basis of twentieth century physics and of electronics.

* Decimal Classification: R094. Original manuscript received by the Institute, June 10, 1937. Presented at Silver Anniversary Convention, May 10, 1937.

† The story is told from the background of one who became acquainted with radio as an amateur wireless telegraphist, was associated with the founding of the Institute and, since then, as an engineer of the Bell System, has taken an active part in the development of both wire and radiotelephony.

Early Experiments

It happens that an early attempt at transmitting speech without wires was made by the inventor of the telephone himself, Alexander Graham Bell. Back in the 1880's he sent speech over a beam of light, using reflectors in much the same way that ultra-short waves are directed today.[1] He called the system the "photophone." Mercardier[2] rechristened it the "radiophone" because it employed frequencies not limited to the visible range, and here we have the earliest use of the word "radio" in the sense employed today.

Of course, the more direct forerunners of radiotelephony were wire telephony and wireless telegraphy. The transmission side of both these arts came out of the early work of Maxwell and Hertz, but they developed for many years quite independently, because of the great difference in the transmission frequencies involved. Early attempts at carrier-current telephony and telegraphy over wires,[29] involving frequencies of tens of thousands of cycles and utilizing modulation, frequency selecting circuits and detection, were unsuccessful because of the lack of suitable technique even for those frequencies, and were in general unknown to later wireless telegraph experimenters. The devices with which Marconi initiated practical wireless telegraphy were adapted to frequencies of the order of a million cycles, generated discontinuously by means of sparks. In time wireless telegraphy evolved toward the use of continuous waves and, by such means as the high-frequency alternator and the oscillating arc, bridged the gap between the radio and the wire frequency ranges.

In the period of 1906–1912 radiotelephony was an experimental fact but a practical nonreality. Many were the early experimenters who had succeeded in transmitting speech over distances of some miles, notably Fessenden and De Forest in America, and Majorana, Vanni, and Poulsen in Europe. In 1911 General Squier,[12] of the United States Signal Corps, brought widespread attention to the possible application of the then wireless instrumentalities to high-frequency transmission over wires.

But, radiotelephony remained for the radio experimenter a golden goal of attainment, for there were wanting practicable means for generating the high-frequency currents, for controlling them in accordance with the relatively weak waves of speech, and for renewing at the receiving end the waves so greatly weakened in transit. The story which follows of the successful meeting of these problems, principally by means of the vacuum tube, is broken by the incidence of the Great War into three periods.

[1] Numbers refer to bibliography.

The Formative Period of 1912–1916

In retrospect, it is now apparent that by about the time of the formation of the Institute the general front of technical advance had reached the point of almost inevitably yielding the solution of the radiotelephone problem. The two-element[4,5] and three-element[7,9,13] vacuum tubes existed, knowledge of thermionics and means for attaining higher vacuua were accumulating, coupled tuned circuits were well-known, and in wire telephony the basis had been laid in the loaded-line theory for the electric wave filter[20,21] and circuit network philosophy. But lest the attainment of radiotelephony seemed too easy, let us follow in a little more detail how the structure of the art was built. The scene is placed in America principally, for it was here that De Forest was experimenting with his three-element audion tube, that telephony generally was developing apace, and that certain research laboratories were working upon problems which needed the tube.

The High-Vacuum Tube

Dr. Lee De Forest invented the three-element tube in 1906–1907, but it was not until about 1912 that he succeeded in adapting it under some circuit conditions to operate as a true amplifier. In the Fall of that year he and an associate, John Stone Stone, demonstrated the audion to engineers of the telephone company* in the role of an audio amplifier, a candidate for the solution of the telephone repeater problem. The device was still a weak and imperfect thing, had in the grid circuit the familiar blocking condenser of the audion detector, and was incapable of carrying any considerable voice load without blue hazing; yet, it *was* capable of amplifying speech.

Among those in the telephone laboratory who witnessed De Forest's demonstration was one H. D. Arnold, then fresh from the study of electron physics in Dr. Millikan's laboratory of the University of Chicago. Whereas there had always been confusion of thought concerning the effect of gas upon the operation of the audion, Arnold immediately recognized that what was wanted was a pure thermionic effect, free of gas complications. He set to work to produce a higher vacuum tube, using evacuation methods then only recently available. He succeeded and, once and for all, took the three-element tube out of the realm of uncertainty and unreliability and made of it a definite, reliable, amplifying tool.

About the same time that Arnold was doing this in the laboratories

* By "telephone company" is meant the American Telephone and Telegraph Company and Associated Companies, including the Western Electric Company, and now also the Bell Telephone Laboratories, which comprise the Bell System.

of the telephone company, principally in 1913, Langmuir, in the laboratories of the General Electric Company, studying the problem of X-ray tubes and power rectifiers, arrived at substantially the same result.[17] In a patent contest lasting many years the Supreme Court of the United States gave to Arnold the credit of having been the first to attain the truly high-vacuum tube and agreed with Arnold's original viewpoint that this step, important though it was, did not constitute invention over the prior art.

By the time the high-vacuum tube was obtained several gaseous forms of tubes had appeared. One of these was of the mercury-vapor

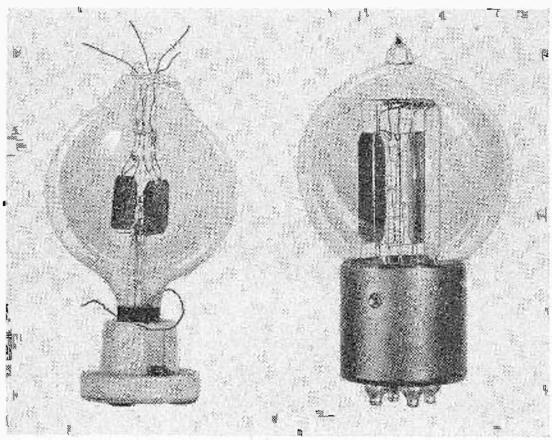

Fig. 1—Early De Forest audion and telephone repeater tube—about 1914.

type, employing magnetic control, which was being worked upon by Arnold as a telephone repeater at the time the audion was first called to his attention. Another was the tube of von Lieben and Reisz, of Austria and Germany, which employed a grid element. All such gaseous devices were soon eclipsed by the high-vacuum tube.

The high-vacuum tube was further improved in 1913 by the application to it of Wehnelt's oxide-coated cathode.[6,27] The filament electron emission was thereby increased, producing the dull-emitter type of long-life tube. The vacuum tube in this form, stable, with adequate filament emission and long life, set the pace in the amplifier art from that time forward, and was the practical basis of the succession of further developments which resulted in practical radiotelephony. One of De Forest's audions and one of the early high-vacuum telephone repeater tubes are pictured in Fig. 1.

Oscillator

One of the next developments was, of course, the conversion of the vacuum tube amplifier into a generator of high-frequency currents. This was accomplished first by De Forest in 1912, according to a decision of the United States Supreme Court. Others did it independently about the same time, notably Armstrong here and Meissner in Germany. Particular forms of oscillating vacuum tube circuits were developed by other investigators, including C. S. Franklin and H. J. Round, of the British Marconi Company, and Colpitts and Hartley, in the United States. Armstrong's 1915 I.R.E. paper[16] upon the subject was a notable one, as was evidenced by the demand for the issue of the Proceedings in which it appeared.

The earliest uses known to have been made of the oscillator in radiotelephony are the experiments of Meissner,[14,32] in Germany in 1913, between Berlin and Nauen, using the von Lieben-Reisz tube, and of H. J. Round,[19] of England, early in 1914 in experimental transmission between two ships.

Modulator

Another major step, the invention of the vacuum tube modulator, soon followed. This solved the problem of enabling low power voice energy to control the considerably higher power waves required for radiotelephone transmitting, and enabled this control to be exercised remotely over a telephone line, thereby giving through-transmission between wire and radio circuits. The earlier attempts at radiotelephony had depended for modulation upon the carbon microphone, usually worked directly in the antenna ground circuit. Here, again, we have a case of several investigators arriving at the invention at about the same time, 1913–1914, with Alexanderson, of the General Electric Company, and Colpitts, of the telephone laboratories, sharing the honors. Other modulating circuits followed. The telephone engineers had in mind doing the modulating at low power and then amplifying the modulated current by means of a high-frequency amplifier.

High-Frequency Vacuum-Tube Telephony

By the latter half of 1914 there was within grasp in the laboratories sufficient of the high-frequency technique, based upon the high-vacuum tube, to cause the telephone engineers to set about the development of high-frequency telephone systems. The first attempt was at the wire carrier-current problem. A two-channel multiplex system was set up using vacuum-tube oscillators, modulators, amplifiers, and detectors. The result was decidedly encouraging. Since the same instrumentalities were applicable to radiotelephony there was next undertaken the development of a vacuum-tube radiotelephone system.

These early wire carrier-current and experimental systems proved to be the precursors of our modern art. They mark a climax in what is perhaps the most rapid accretion of technique known in modern electric communications, from the condition, in the fore part of 1912, of there being no suitable generator nor modulator, to that of 1914–1915 where these essentials had become available and were being synthesized into operative high-frequency telephone systems.

Long-Distance Tests of 1915

Vacuum-tube radiotelephony was now to be taken out of the laboratory for a field trial. A vacuum-tube transmitter of a few watts output was developed and installed at Montauk Point, Long Island, and an amplifying receiver was located at Wilmington, Delaware, 200-odd miles distant. The distance was then stretched to some 600 miles by receiving the Montauk transmitter at St. Simons Island, off the coast of Georgia. These were one-way transmissions. For some of them the reception was brought back to New York by wire lines. The speech was itself clear, but was sometimes buried in noise due to the small transmitting power and the fact that it was the spring of the year. Wave lengths of 800 to 1800 meters were employed.

The success of these preliminary tests, together with the promise of laboratory developments for higher-power transmitting tubes, now led to a bold attempt on the part of the telephone engineers to overcome that great natural barrier of telephony, the oceans. Through the cooperation of the United States Navy Department, on the one hand, and the French Administration, on the other, appropriate field stations were made available for the tests. The large antenna of the naval station at Arlington, Virginia, was used for transmitting. A new vacuum-tube radiotelephone transmitter was developed, employing hundreds of tubes, each having a capacity of the order of fifteen watts, and installed at Arlington. For reception the Navy Department made available their stations at the Canal Zone, on the Pacific Coast, and in Hawaii. Through the kindness of General Ferrié, of the French Administration, use of the Eiffel Tower station was permitted the American telephone engineers for receiving purposes. Thus did the French collaborate in the interest of technical advance and international good will by accepting foreign engineers in their most important military station during the life-and-death struggle of the Great War.

By June all the distant receiving points were covered by engineers who had been dispatched from New York provided with the then latest receiving apparatus; the new telephone transmitter had been installed at Arlington; a great effort was being made in the laboratories to produce the necessary quantity and quality of power tubes. The tests continued on and off during the entire Summer on a reduced power basis, during which the difficult atmospheric conditions were studied by the receiving engineers. As the transmitting power was built up, and as the conditions improved with the coming of Fall, results began to be obtained, first, from Panama; next, from the Pacific Coast, representing transmission across the continent; then, from more distant Hawaii; and, finally, in November from Paris, where the receiving conditions had proven to be most difficult.

These were, of course, one-way transmissions. The reception was so uncertain and so subject to noise as to make it evident that the art would need to be advanced greatly before the requirements of a service could be met over such long distances.

Some of the technical features of the apparatus of these early tests were:

In the transmitting station, the use of the master-oscillator, power-amplifier type of circuit, operating in the 30 to 100-kilocycle range, with circuits designed to accommodate the "carrier and sideband" aspect of the modulated wave.

The development of power tubes of the order of fifteen watts, requiring new designs and more thorough pumping and degassing.

The operation of large numbers of tubes in parallel (as many as 500), in order to build up the necessary transmitting power. The problem of operating these tubes in parallel and preventing singing can well be imagined. An average power of two or three kilowatts was obtained in the antenna. A photograph of two banks of 250 tubes each is reproduced in Fig. 2, the tips of some of the tubes showing on the right.

Receivers employing a radio-frequency amplifying stage, plus two audio-frequency stages. Heterodyne detection was employed to find the carrier. Homodyne reception of the telephone signals was used at some of the receiving points.

The following year, 1916, tests of radiotelephony were made for the United States Navy, which included what is believed to have been the first attempt at tying together radio and wire lines for through two-way radiotelephony. The Secretary of the Navy, Josephus Daniels, talked from his desk in Washington, D. C., with the commanding officer of the *U.S.S. New Hampshire*, off the Chesapeake Capes. During this year the important subject of modulation was advanced and Heising devised his well-known "constant current" system of modulation.[30] The simplicity of this system led to its use in the radio-

Fig. 2—Power amplifier of the Arlington, Virginia, experimental transmitter, 1915.

telephone sets which were produced during the war and in the early radio-broadcast transmitters. Incidentally, early in this same year there was undertaken anew the problem of carrier-current telephony and telegraphy, this time looking toward commercial designs, utilizing the newly acquired instrumentalities of the vacuum-tube and the electric wave filter. The application of the vacuum tube amplifier to voice-frequency telephone circuits was also proceeding apace.

Radiotelephony was now progressing rapidly, building upon and, in turn, stimulating its antecedent and contemporary arts of wire telephony, wireless telegraphy, and electronics. Something of the content of these related arts is indicated below:

ARTS UNDERLYING RADIOTELEPHONY

Wire Telephony	Wireless Telegraphy	Electronics
Electroacoustics Transmitters, receivers, characteristics of sound, high-quality reproduction	Generators and receivers of high-frequency currents; selective circuits	Discovery and study of the electron (Crookes, J. J. Thomson and others)
Wire Transmission Propagation constant, characteristic impedance, transmission measurement, interference, carrier, wave filters, and network theory	Antennas Dipole (Hertz), grounded (Marconi), directive	Thermionics (Richardson, Wehnelt and others)
	Wave Propagation Spreading and absorption, ground and sky waves, effects of solar and meteorological phenomena	The Edison effect and the Fleming valve
Amplification Microphone, repeaters		De Forest 3-electrode tube
		High-vacuum, high-power, and multielectrode tubes

THE WAR PERIOD

The war came to Europe before the new vacuum-tube art and radiotelephony had been fully born. Vacuum tubes were employed in the war by the European countries for radiotelegraphy, but radiotelephony is not known to have played a part on the Continent. This may have been due in part to the lack of secrecy of this form of communication.

In the United States the normal development of radiotelephony continued, as we have seen, up to the time of this country's entry into the war. The new vacuum-tube radiotelephony had by then assumed real promise. The United States Government undertook to develop two-way radiotelephone sets on a large scale for dispatch purposes on submarine chasers and airplanes. In the short space of a year or so hundreds of thousands of tubes and thousands of sets were developed and manufactured. Several of the larger laboratories of the country were in effect taken over by the government for this purpose. The apparatus was featured technically[23] by:

The general use of the high-vacuum, oxide-coated filament type of tube.

Employment of the constant-current type of modulation.

The attempt to make the operation of the sets simple and foolproof, as by the elimination of filament rheostat and the standardizing of tubes and circuits to permit of ready interchangeability.

Such apparatus was used in some quantity on submarine chasers of the Navy, but the large production program for airplanes was not completed in time to enable radiotelephony to come into play in the Army on the battle front. From the technical standpoint the program stimulated apparatus design and gave a useful experience in the standardization and quantity production of tubes.

Through the many military training schools in the United States the new vacuum tube radiotelephone art was "broadcast" to the most likely young men of the country, many of whom developed a real interest and after the war helped to swell the peacetime development.

Post-War Developments

The peacetime development of the art was now immediately renewed, especially in the United States where the technical effort had been sustained throughout the war. It became evident to the several large companies, particularly the American Telephone and Telegraph Company and the General Electric Company, which were pursuing the vacuum tube art, that their inventions so interleaved as to require an exchange of patent rights. This took the form of an interlicensing agreement, entered into by these two companies in 1920. It enabled the telephone company to use tubes on its lines and to proceed with the development of two-way radiotelephony. The General Electric Company and its affiliates, including the R.C.A., were free to proceed in other fields, principally radiotelegraphy, and, as it turned out, in broadcasting.

Early Ship-to-Shore Telephone Experiments

The telephone company had by this time undertaken development work in marine radiotelephony, partly as a means of advancing the art and partly with an eye toward the eventual establishment of a mobile public telephone service connecting with the land line system. Experimental shore stations were provided (one of which, that at Deal Beach, New Jersey, is shown in Fig. 3), and ship apparatus capable of duplex operation was devised and tested on coastal vessels and on one of the transatlantic liners. This work was done in the frequency range then most available, that of the order of a million cycles. Shortly thereafter broadcasting preempted this range and because of this and of the post-war depression in shipping, these experiments did not then materialize into a service. Trial connections with the landline network extending across the Continent and to Catalina Island served to demonstrate the possibilities of combined wire and radio. Some of the technical attainments in this work[34] were:

The development of duplex systems for ship use.
The development of superheterodyne receivers.
Progress in placing radio transmission upon a quantitative basis by the measurement of received field strengths and the

overall circuit equivalent of radio links, and in the setting up of radio links as integral parts of long landline connections.

The beginning of the volume indicator, used to insure the voice loading of the radio transmitter and later employed extensively on wires, as well as in radio.

First Public Telephone Service by Radio

Another pioneering undertaking about this time, 1920, was the development of what has proven to be the first use of radiotelephony for public service, in the form of a point-to-point link on the Pacific

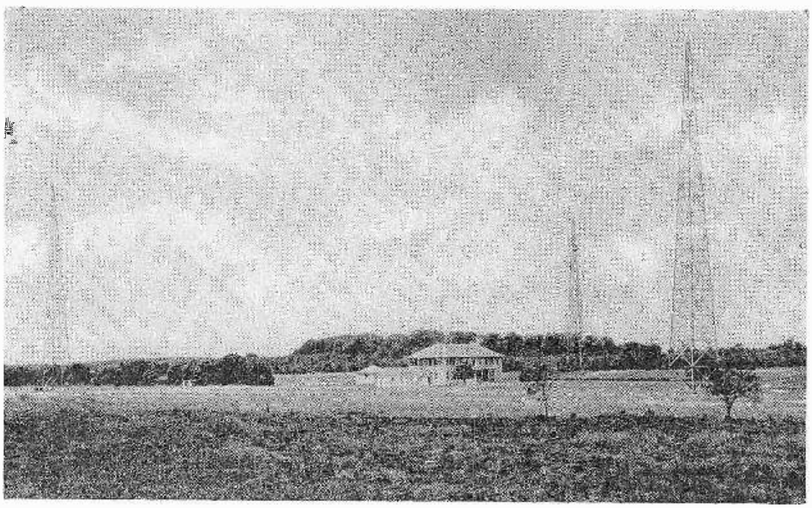

Fig. 3—Experimental Radiotelephone Station, Deal Beach, New Jersey

Coast between Catalina Island and Long Beach on the mainland, connecting thence to Los Angeles.[31] Service was given over this link for about a year, when the frequency band being used was wanted for the then newly developing service of broadcasting. The telephone service itself came near being a broadcast one, so extensively were the conversations listened to by amateur radio enthusiasts. The system was replaced by a submarine cable.

From the standpoint of technical progress, this installation included a number of interesting features:

Full-duplex operation in the sense of separate channels for the two directions of transmission, joined at the terminals to the two-wire telephone network by means of hybrid coils.

Through voice-frequency ringing, the first application to radio.
Superheterodyne receiving sets, incorporating wave filters in the intermediate-frequency stages which separated out:

Fig. 4—Long Beach, California, receiving terminal of the Catalina Island radiotelephone link, 1920.

A telegraph channel which was superimposed upon; i.e., multiplexed with, the telephone channel and used independently for telegraph service with the Island.
The provision toward the end of the period of means for rendering the telephone transmission private, comprising voice inverters, plus carrier-frequency wobbling, the first installation of this combination to have been made.

The picture of the Long Beach receiver in Fig. 4 shows at the top a portion of the loop receiving antenna; in center foreground, the circuit control desk; above, to the right, the speech inverter for privacy; and, in the left background, the apparatus of the superimposed telegraph channel.

Broadcasting

By 1920–1921 the stage was set in the United States for radio broadcasting. A radiotelephone technique was becoming available in the relatively empty portion of the frequency range centering about one megacycle. Something of an audience existed in the thousands of amateur radiotelegraphists spread throughout the country, a lively public interest in radiotelephony had been aroused during the war, and all that was needed to excite the public generally into providing itself with receiving apparatus was to have the experience of hearing speech and music on the air. These essential elements of an appropriate technique and of a widespread audience were lacking in the earlier years when De Forest and others broadcast speech and music upon a number of occasions with considerable success.

Public interest was first fanned by amateur listening to the experimental telephone transmissions being conducted by various people, amateur and professional. Engineers in making tests frequently availed themselves of reports written in by listeners, as a means of checking in a general way the effectiveness of their transmitters. For example, in the Fall of 1919, tests made between New York and Cliffwood, New Jersey, of a pair of 500-watt transmitters intended for shipment to China, were reported by many amateur listeners. Tests of ship-to-shore radiotelephony, which were being made on more or less regular schedules from Deal Beach, New Jersey, were listened to and reported by hundreds of amateurs throughout the eastern part of the country. In the vicinity of Los Angeles, California, listening to the radiotelephone link to Catalina Island was becoming enough of an indoor sport to be embarrassing to the public telephone service, as has been mentioned.

Of all the experimental activity at the time, it happened to fall to the personal efforts of Frank Conrad, an engineer of the Westinghouse Electric and Manufacturing Company, Pittsburgh, Pennsylvania, to give rise to broadcasting of a continuing nature. Starting with transmissions from his home, the activity was taken up by his company, which had been engaged during the war in making radio apparatus for the government, and the experimental emissions were evolved into a continuing program, accompanied by the entering of the company into the business of supplying receiving sets. The original transmitter of the now well-known station of KDKA is pictured in Fig. 5 as it appeared on the occasion of its first broadcasting, when it sent out the returns of the presidential election on November 2, 1920. Note that the room which housed the transmitter served also as the studio. Public interest mounted rapidly and within a few years transmitting stations were growing up throughout the country[41] and a boom was on in receiving sets. So great was the demand for transmitting station equipment that the telephone company was called upon to provide what proved to be most of the installations in these earlier days.[36]

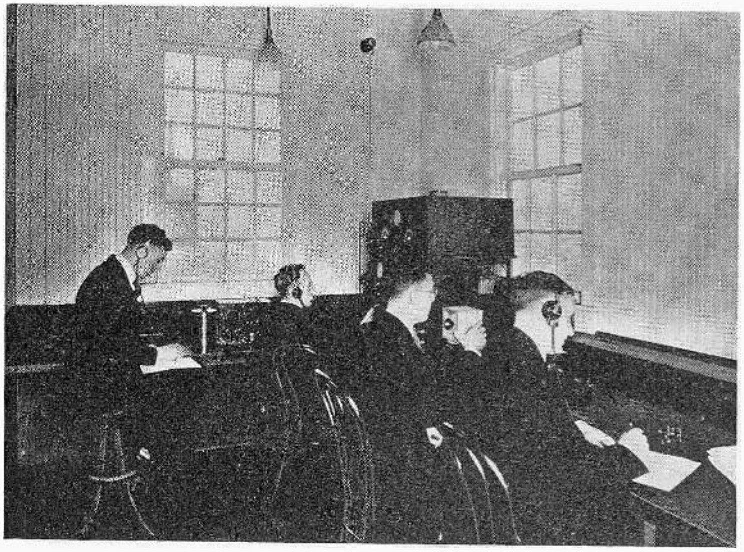

Fig. 5—First KDKA Transmitter, Pittsburgh, Pennsylvania, 1920.

This great burst of activity brought with it real concern as to the character which broadcasting might assume and as to how it could be supported as a continuing service. It being a form of telephony, the telephone company undertook to explore the field from the sending end by engaging in broadcast transmitting. There evolved the idea of putting the transmitter at the disposal of others for hire (toll broadcasting), the sponsored program, and arrangements for the syndication of programs over the wire telephone network. Thus was demonstrated the ability to support broadcasting from the sending end.

One of the first five-kilowatt, water-cooled transmitters, that used in Station WEAF of the telephone company in 1924, is shown in Fig. 6. Aside from representing an advanced design at the time, this transmitter is associated with an interesting bit of technical history. A 500-watt transmitter, which had been used just before it, had shown bad quality when received in certain outlying sections of the city on the far side of groups of skyscrapers. This gave rise to the making of one of the first studies of the broadcast transmission medium, including the element of fading and of coverage.[37] The trouble proved to be due to the effect of the tall buildings in attenuating the direct transmission and making apparent interference between multiple paths. The effect of the interference upon quality proved to be exaggerated by a degree of frequency modulation occurring in the transmitter. The latter

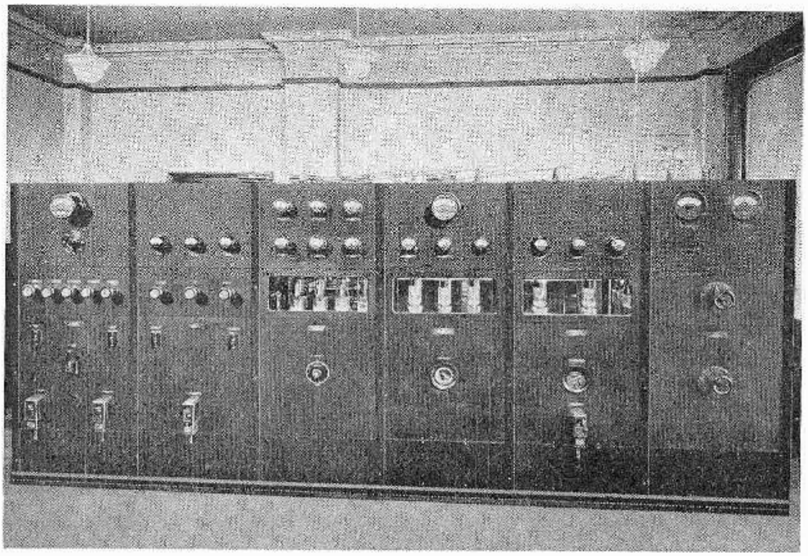

Fig. 6—Early Five-Kilowatts Transmitter of WEAF, Crystal Controlled, 1924.

trouble was removed by the adoption in the five-kilowatt transmitter of the master-oscillator type of circuit, employing piezoelectric crystal control, one of the first transmitters so provided.

The intimate interplay which existed technically between telephony broadly and broadcasting is shown also in the high quality side of broadcasting, involving studio acoustics, high-quality microphone pickup and high quality amplifiers. In the beginning of broadcasting the pickup and amplifying means were taken more or less bodily from the high-quality speech study work which had been going on in the telephone laboratories,[35] and from public address systems. In 1919 a great public address demonstration had been made in New York upon the occasion of a Liberty Loan; and in the Summer of 1920 such systems had played a prominent part in the two national political conventions. Addresses delivered over such systems from a distance emphasized the need for high quality lines. As a result of such experience and the considerable amplifier network technique which had been built up in the long-distance telephone field, it was possible at an early stage of broadcasting to adapt telephone lines to handle as wide a sound-frequency band as the economics of the situation justified.

Broadcast Receivers

A realization of the progress which has been made in broadcast receivers is had by contrasting the modern, stable, and selective loudspeaker set with the ticklish crystal or regenerative battery set with which listeners first heard whispers in headphones. One of the first advances was to the high-frequency amplifying set, whereby sensitivity was achieved together with simplicity of adjustment. The stabilizing of these sets against singing stimulated the art of tube-balancing circuits and is remembered by Hazeltine's neutrodyne. The superheterodyne, the indirectly-heated cathode tube permitting operation from the alternating-current supply mains, the screen-grid tube, automatic gain control, featured the rapidly evolving receiving-set technique. Loud speakers progressed from the old horn type to the armature driven cone, to the electrodynamic, and multiple-unit system. While many of these advances had their origin elsewhere than in broadcasting, certainly the quantity production of broadcast receiving sets has been a powerful leaven in advancing the weak-current technique generally.

Transoceanic Telephony

As broadcasting was getting started, continuing research in the laboratory gave promise of considerably greater transmitting powers, in the form of the copper-anode, water-cooled tube. This and the other advances which had occurred since the original transoceanic experiments of 1915 indicated that it might be timely again to undertake the problem of extending telephony overseas.

A powerful water-cooled amplifier, the first of its kind, was developed and in 1922, in cooperation with the R.C.A., was installed at the transatlantic transmitting station at Rocky Point, Long Island. It is pictured in Fig. 7. Success attended the first objective of developing an antenna power of the order of 100 kilowatts and the transatlantic project was vigorously pushed. This work being in the then relatively low frequencies, it was possible to adopt single-side-band, carrier-suppressed transmission by borrowing that feature more or less bodily from the wire art, whereby the transmitting effectiveness was multiplied by a factor of about ten. The transmitting path to England was studied by making measurements there, in collaboration with the engineers of the British Post Office, of the diurnal and seasonal variations received and of the noise levels. A further improvement was obtained by borrowing from the wireless telegraph art the newly-developed directive antenna known as the Beverage wave antenna.[33] There were, of course, other problems in getting started, and these and the manner in which service was established, beginning in 1927, and, in

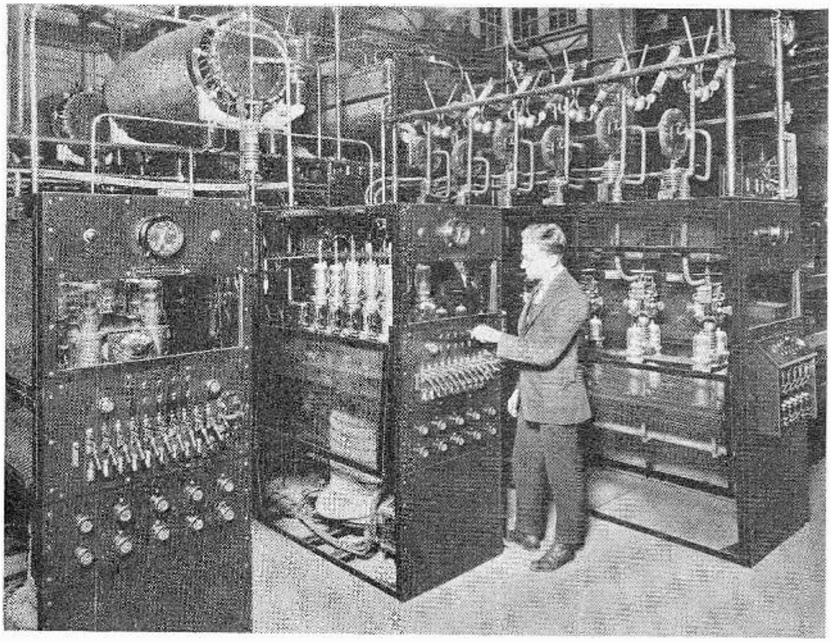

Fig. 7—Power Stage, in two units, with rectifier, of the first transatlantic radiotelephone transmitter at Rocky Point, Long Island, 1923.

time, extended by the use of high frequencies, are told in a companion paper entitled "Transoceanic telephone development," by Ralph Bown.

Higher Frequencies and Mobile Services

The extension of radio to the higher frequencies, or shorter waves, gave new opportunity for the development of radiotelephone services because of the greater message-carrying capacity of the higher frequencies and the greater transmission range.

Following the introduction of short waves to transatlantic telephony, the ship-to-shore problem was undertaken anew on a short-wave basis[40] and service was initiated on the North Atlantic in 1929. On the shore end the essential facilities comprised a duplicate of one of the transatlantic point-to-point installations, including directive antennas pointing out along the transatlantic shipping route, and means for effecting two-way operation and for connecting into the wire network. The ship installation included a transmitter of 500 watts capacity, employing a new screen-grid power tube. As a result of "stay noise," there was adopted the "cut carrier" method of transmission. As it has turned out, no American ships yet have been equipped permanently for service, but most of the larger foreign vessels are so equipped and marine telephone service is being given from both sides of the Atlantic.

A related form of marine telephone service is that to small boats. In the United States this started somewhat as a continuance by the Coast Guard of the submarine chaser installations of the war. In Europe fishing trawlers have been provided with simple radiotelephone sets in considerable numbers. In these installations the intention has been to enable the boats to talk with each other and with certain land stations; not with the landline telephone users. Small-boat telephony linked with the landline network is a more difficult matter. It is now under active development in the United States on both the East[44] and West Coasts, and on the Great Lakes, and in some countries in Europe. The installations in the United States are of crystal-controlled sets, designed to be used directly by the officer of the ship without technical attendance. Many of the ships are equipped to be "rung" individually as wanted by the shore station. The small boat telephony works generally in the medium-frequency range of two to three megacycles.

Another type of mobile service is that being used throughout the airways of the United States in maintaining contact between the planes and the ground stations. Telephony has proven particularly useful here because of the facility it offers the pilots of communicating directly on a two-way basis with ground stations. The service is operated generally in the three to six megacycle portion of the spectrum. The apparatus is crystal controlled, of special design for lightness, simplicity of operation, and reliability. This type of service was well started about 1929.[38,39]

A third type of mobile radiotelephone service, and one which has become quite important in the United States, is that of the various city and state police departments, used to direct patrol cars. Most of these services are limited to one-way talking to the cars, and operate on intermediate frequencies. Now that ultra-high frequencies are be coming available, some of these systems are being extended to two-way service. The apparatus is generally similar to that employed in the aviation service.

Ultra-High Frequencies

The recent extension of the radio technique to ultra-high frequencies brings new opportunities and also new problems for radiotelephony. One of the earliest practical trials of these frequencies for telephony, and one representing at the time a very large jump in frequency, was the seventeen-centimeter wave propagation across the English Channel in 1931, accomplished by the system developed by the laboratory of Le Matériel Téléphonique of Paris, using the Barkhausen type of oscillator.[42] Further experience has shown frequencies as high as this to be susceptible under some circumstances to rather serious transmission instability, resulting, it has been suggested, from changing moisture content of the air or from turbulent atmospheric conditions. A number of short radiotelephone links are now being operated in various parts of the world on somewhat lower frequencies, generally in the range of 40 to 100 megacycles.

It appears that as rapidly as the message-carrying capacity of radio is enlarged by extension to the ultra-high frequencies, the demand increases on the part of older services and of entirely new services, such as television. How much of the spectrum may be available for telephony will naturally be influenced by relative usefulness and economics. One problem is how to obtain radiotelephony sufficiently economically to "prove it in" for the shorter distances which characterize the useful range of these waves. Another is the one of preserving the privacy of communication by relatively simple means. It may be that the principle use of these waves for telephony will be for mobile services, thereby helping telephony keep pace with our increasingly mobile way of living.

Leaven of the Art

In describing the rise of radiotelephony we have spoken principally of physical things such as the vacuum tube, the filter circuit, etc. Another cross section of the art would be the leaven of ideas which gave rise to it, the analyses and the reductions to measurement which enabled results to be obtained by design. That radiotelephony is particularly rich in this respect will be evident from the following citation of some of the more outstanding analytical contributions.

One of the first is that of van der Bijl's early study of the operation of the vacuum tube. In 1913–1914 he derived approximate expressions for the plate current in terms of plate and grid voltage, and presented the concept of the amplification factor μ. This work was published[22] toward the end of the war and was the forerunner of his 1920 book[23] on the vacuum tube, an authority for many years.

One of the earliest elucidations to be published of the operation of the audion as detector was the 1914 paper of Armstrong.[15]

Next, there is the more exact mathematical solution of the plate current in terms of the tube constants and grid voltage variations, given in an I.R.E. paper in 1919;[24] and the treatment of the vacuum tube as a part of a circuit network published the same year.[25]

A potent factor has been the growing appreciation throughout the

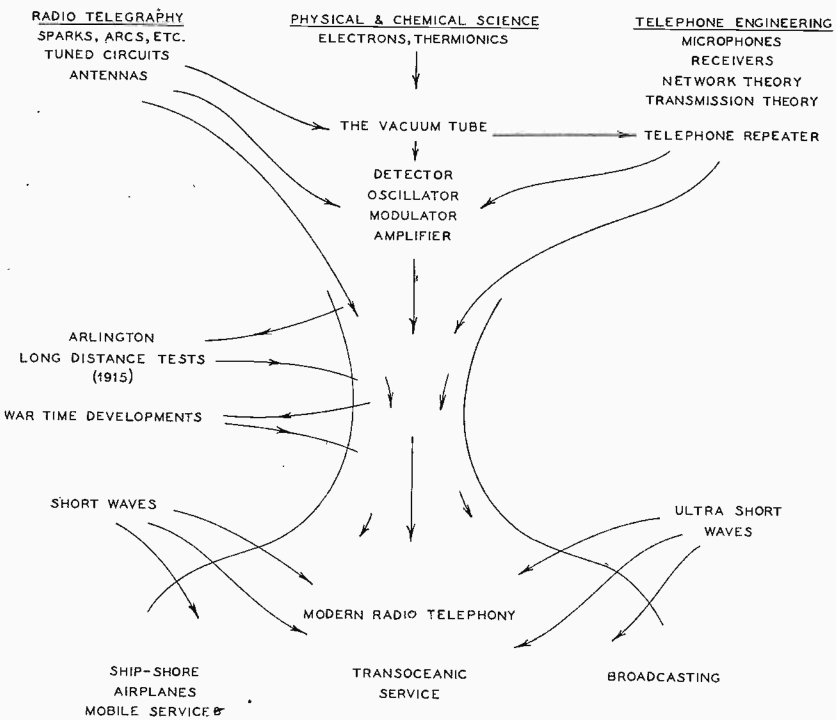

Fig. 8—An attempt to diagram the flow of the art.

electric communication art generally of Fourier's theorem and the steady-state concept of transient phenomena. Related thereto is the band idea of wire telephone transmission which developed out of the frequency-band nature of speech itself, the characteristics of lines, and the necessity of suppressing reflections at circuit junctions. Campbell, using the loaded-line theory, combined the band idea of telephony with the sharp selectivity feature of radiotelegraphy to secure the wave-filter characteristic of a uniform transmission band, plus a sharp cutoff. This was a milestone in the development of circuit network theory.

Related to both vacuum tubes and the band conception were Carson's analysis of the modulated wave into the component carrier and sidebands and his invention of single-side-band transmission,[18] made as far back as 1915, and the general extension of the signal-band idea to high frequencies, which has meant so much to both wire carrier-current telephony and radiotelephony.

In the field of measurement and standardization there are the technique of making single-frequency measurements throughout a band, the decibel unit of attenuation, the volume indicator and the concept of volume range, and the measurements of the field strength of desired signals and of noise.

Fig. 8 represents an attempt to diagram the flow of the art as a whole.

Future

Radiotelephony may be said to have "arrived" and to still be young. Looking toward the future, the writer likes to think of radio and wire telephony as increasingly dovetailing together to form one general front of advance. The principles and technical tools being fundamentally the same for both, a technical advance in one is likely to help the other. Thus radio has led the way to the higher frequencies and this has benefited wire communication, as we see in the carrier-current and wide-band coaxial cable development. As regards services, we observe that radio and wire telephony are one and the same thing in respect to the over-all result, that of the delivery of sound messages. Where one or the other is to be used will be a matter of natural adaptability of the medium of transmission and of the economics of the situation, with the meeting line shifting from time to time, but with the main emphasis upon integration rather than differentiation. Thus does radiotelephony become an integral part of telephony and of the whole field of electric communications.

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TRANSOCEANIC RADIOTELEPHONE DEVELOPMENT*

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Summary—*Marking the tenth anniversary of the beginning of commercial long-distance radiotelephone service, the paper reviews: (1) Technical advances over the older radiotelegraph and short distance radiotelephone arts which made the initiation of long-distance service possible. These included water-cooled tubes, single-side-band suppressed-carrier transmission, and voice operated switching devices designed to prevent singing in combined wire and radiotelephone connections. (2) Engineering developments without which the growth of such services would have been severely restricted. The more important are short-wave transmission, accurate frequency stabilization, and privacy methods. (3) Certain further improvements of less universal application. (4) Present outlook for future development. In most immediate prospect are the extensive application of single-side-band reduced carrier in short-wave commercial operation and the commercial use of a new and improved method of receiving known as a multiple unit steerable antenna. The possibilities of applying grouped channel and multiplex methods to radio transmission are considered. The treatment is brief but comprehensive, being intended to interest the general engineering reader. A bibliography is attached.*

TEN years have elapsed since the opening to public use on January 7, 1927, of the first long-distance radiotelephone circuit. This form of intercontinental communication has now come into practical business and social use. A network of radio circuits interconnects nearly all the land wire telephone systems of the world. The art has passed through the pioneering stage and is well into a period of growth. The present paper reviews the technical side of this development and considers the outlook for future advances. The brief treatment given is intended for those who desire a comprehensive but not detailed acquaintance with the features unique to this field of radio engineering activity. The appended bibliography furnishes adequate leads for an exploration of the English literature of the subject.

A graphic view of the point at which the establishment of radiotelephone facilities now stands is given by the map in Fig. 1, and by the accompanying tables which list further data relating to the circuits illustrated by the map. The circuits shown permit interconnection of about ninety-three per cent of the world's telephones and enable United States telephone subscribers to reach sixty-eight other countries.

* Decimal classification: R412. Original manuscript received by the Institute, June 10, 1937. Presented before Silver Anniversary Convention, May 10, 1937.

506	London– Bombay	British Post Office Indian R.T. Co.	4,474	5/ 1/33
508	London– Tokio	British Post Office Japanese Government	5,941	3/13/35
510	Berlin– Manila	German Post Office Phil. L.D. Telephone Co.	6,124	3/ 1/33
511	Berlin– Tokio	German Post Office Japanese Government	5,532	3/13/35
	No. America–Asia–Oceania			
601	San Francisco– Honolulu	American Tel. and Tel. Co. Mutual Telephone Co.	2,393	12/23/31
602	San Francisco– Manila	American Tel. and Tel. Co. Phil. L.D. Telephone Co.	6,969	3/30/33
603	San Francisco– Tokio	American Tel. and Tel. Co. Japanese Government	5,133	12/ 8/34
605	San Francisco– Bandoeng	American Tel. and Tel. Co. Neth. Indies Tel. Admin.	9,101	2/ 1/34
	North America			
701	New York– Hamilton (Bermuda)	American Tel. and Tel. Co. Cable and Wireless, Ltd.	795	12/21/31
702	Miami Tegucigalpa (Honduras)	American Tel. and Tel. Co. Tropical R.T. Co.	919	4/23/35
703	Miami– Managua (Nicaragua)	American Tel. and Tel. Co. Tropical R.T. Co.	1,002	6/ 7/33
704	Miami– San Jose (Costa Rica)	American Tel. and Tel. Co. Tropical R.T. Co.	1,120	3/20/33
705	Miami– Panama	American Tel. and Tel. Co. Tropical R.T. Co.	1,161	2/24/33
706	Miami– Nassau	American Tel. and Tel. Co. Bahamas Government	175	12/16/32
707	Miami– Guatemala	American Tel. and Tel. Co. Tropical R.T. Co.	1,017	4/17/33
708	Miami– Trujillo	American Tel. and Tel. Co. Dominican Tel. Co.	833	10/31/35
709	Miami– Kingston	American Tel. and Tel. Co. Jamaica Tel. Co.	576	4/ 3/36
710	Miami– San Juan	American Tel. and Tel. Co. Puerto Rico Tel. Co.*	1,034	2/20/36
711	Miami– San Salvador	American Tel. and Tel. Co. Salvador Government	1,013	6/10/36
712	Trujillo– San Juan	Dominican Tel. Co. Puerto Rico Tel. Co.	251	9/ 4/36
	South America			
801	Buenos Aires– Rio de Janeiro	Transradio Internacional Companhia R.T. Brasileira	1,227	/ /31
802	Buenos Aires– Rio de Janeiro	Compania Int. de Radio* Co. Radio Int. do Brasil*	1,227	12/12/31
803	Bogota– Santiago	All American Cables, Inc.* Co. Int. de Radio S.A.*	2,644	8/ 1/31
804	Lima– Santiago	Co. Peruana de Tel. Ltd.* Co. Int. de Radio S.A.*	1,537	12/15/32
805	Lima– Bogota	Co. Peruana de Tel. Ltd.* All American Cables, Inc.*	1,166	2/22/35
806	Lima–	Co. Peruana de Tel. Ltd.*		

	Buenos Aires	Compania Int. de Radio*	1,952	12/ 1/34
807	Bogota– Buenos Aires	All American Cables, Inc.* Compania Int. de Radio*	2,902	12/10/34
808	Santiago– Antofagasta	Co. Int. de Radio S.A.* Co. Int. de Radio S.A.*	690	11/ 1/35
809	Lima– Antofagasta	Co. Peruana de Tel. Ltd.* Co. Int. de Radio S.A.*	1,004	6/16/36
	Asia–Oceania			
901	Sydney– Wellington	Australian Post Office New Zealand Government	1,375	11/25/30
902	Bandoeng– Bangkok	Neth. Indies Tel. Admin. Siamese Government	1,469	4/15/31
903	Bandoeng– Sydney	Neth. Indies Tel. Admin. Australian Post Office	3,392	12/23/30
904	Bandoeng– Medan (Sumatra)	Neth. Indies Tel. Admin. Neth. Indies Tel. Admin.	875	9/16/31
906	Bandoeng– Makassar (Celebes)	Neth. Indies Tel. Admin. Neth. Indies Tel. Admin.	854	7/ 1/33
907	Bandoeng– Manila	Neth. Indies Tel. Admin. Phil. L.D. Telephone Co.	1,742	2/23/33
908	Bandoeng– Tokio	Neth. Indies Tel. Admin. Japanese Government	3,609	10/26/34
909	Tokio– Taihoku (Formosa)	Japanese Government Japanese Government	1,338	6/20/34
910	Tokio– Hsinking	Japanese Government Manchukuo T. and T. Co.	965	8/ 1/34
911	Tokio– Manila	Japanese Government Phil. L.D. Telephone Co.	1,867	9/26/34
914	Bandoeng– Kuala Lumpur	Neth. Indies Tel. Admin. Fed. Malay States Gov't.	720	3/19/34
915	Kuala Lumpur– Bangkok	Fed. Malay States Gov't. Siamese Government	1,018	9/ 2/35
916	Tokio– Shanghai	Japanese Government Chinese Government	1,100	2/15/36
917	Tokio– Saigon	Japanese Government Companie Generale de T.S.F.	3,375	5/ 1/36
918	Saigon– Bangkok	Companie Generale de T.S.F. Siamese Government	458	5/ 4/36
919	Bandoeng– Saigon	Neth. Indies Tel. Admin. Companie Generale de T.S.F.	1,198	10/19/36
	Europe			
1001	Barcelona– Palma (Majorca)	Co. Tel. Nac. de Espana* Co. Tel. Nac. de Espana*	112	10/24/31
1002	Nice– Calvi (Corsica)	French P.T.T. French P.T.T.	105	6/ 1/32
1003	Reykjavik– Copenhagen	Icelandic Tel. Admin. Danish Post Office	1,305	8/ 1/35
1004	Reykjavik– London	Icelandic Tel. Admin. British Post Office	1,175	8/ 1/35
1005	Moscow– Paris	U.S.S.R. Administration French P.T.T.	1,544	1/21/35

* Indicates International Tel. and Tel. Co. subsidiary or affiliate.

Note: Circuit Group 502 is so arranged that each transmitter has two independent telephone circuits, making a total of four circuits between Amsterdam and Bandoeng.

The technical matters to be considered fall into four categories. The first covers those factors which made possible the beginning of commercial radiotelephony. In the second are the things without which its rapid growth and wide expansion could not have occurred. In the third are a few incidental but interesting or valuable technical features. The fourth considers future improvements now in view.

A description of the early years of radiotelephone development preceding extensive commercial application, together with a discussion of the origins of the whole art, will be found in a companion paper "The origin and development of radiotelephony," by Lloyd Espenschied. This early work included the experiments in 1915 during which the voice was first sent overseas by radio, from Arlington, Virginia, in the United States, to Paris, France, to Panama, and to Hawaii. Also noteworthy were the wartime development of radiotelephony to small vessels and airplanes, and in 1920, the first practical tests of high seas ship-shore radiotelephony. These gave a basis of engineering experience for establishing the first commercial radiotelephone circuit, which operated between Santa Catalina Island and the California Coast, a distance of about twenty-five miles. But all these developments were, comparatively speaking, of such low power as to give no immediate promise of practical use in long-distance commercial communication. Long-distance radio service up until the middle twenties was entirely by telegraph and employed such generating agencies as high-frequency alternators, spark gaps, and arcs, which had not been found practically usable for telephony.

Essential Initial Developments

Radiotelephony presented difficulties in addition to those existing in radiotelegraphy because: (1) The communication is two-way, and the radio system must be linked in with the wire telephone systems and available to any telephone instrument; (2) the subscriber cannot deliver himself of his message until the connection is actually established, and on this account delay due to unfavorable transmission conditions is less tolerable; (3) the grade of transmission required to satisfy the average telephone user is higher than that tolerable in aural tone telegraph reception by an experienced operator.

These requirements emphasized the need for accurate and quantitative knowledge of radio transmission performance as a basis for engineering radiotelephone systems. There was at the same time a similar need for transmission data in the engineering of early radio-broadcast installations. The effort brought to bear on these twin problems resulted in the development of practical field methods[1] of measuring radio signal strength and radio noise. The employment of long-distance radiotelephony in commercial use was preceded by experimental operation and tests[2] which gave a considerable fund of statistical information covering the cyclical changes characteristic of overseas radio transmission.[3]

The realization that a relatively high degree of reliability was essential to success discouraged any attempt at commercial service until

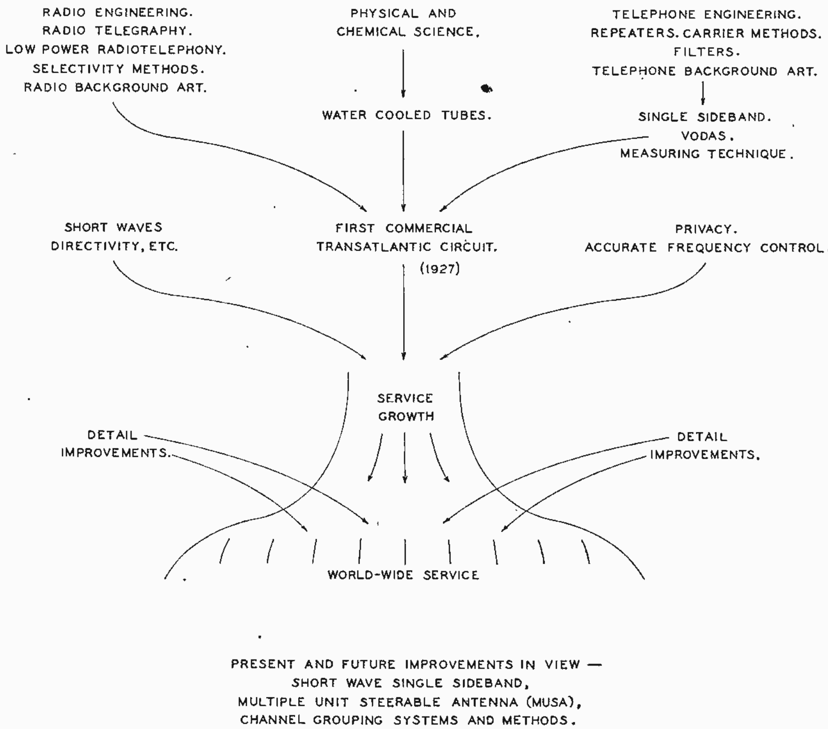

Fig. 2—One way of diagraming technical factors contributing to the development of radiotelephone service.

high power transmission on a practical basis was assured. The invention of a method of making vacuum tight seals between metal and glass envelopes opened the door to higher power. The development of water-cooled[4] tubes became the starting point of successful long-distance radiotelephony.[5]

In searching for the most efficient way of applying the power made available by water-cooled tubes, telephone engineers were led to the

[1] Numbers refer to bibliography.

employment of a method which had already been successfully used in high-frequency wire telephony. This method, now well-known to radio engineers, is called single-side-band suppressed-carrier transmission.[6] As compared with the ordinary modulated carrier transmission, it increases the effectiveness of a radiotelephone system by about ten-to-one in power. This accrues partly because none of the power capacity of the transmitter is used up in sending the noncommunication bearing carrier frequency and partly because the narrower band width permits greater selectivity and noise exclusion at the receiver.

A very important final element was then necessary to permit effectively utilizing in a combined wire and radio system the potential power capacity provided by single-side-band water-cooled tube transmitters.

In a radiotelephone conversation the talking back and forth in the two directions requires two separate one-way radiotelephone circuits, one for each direction. In the local wire telephone plant the oppositely directed halves of the conversation travel both on the same pair of wires. The ordinary wire telephone practice for joining a two-way circuit with two one-way circuits is the balanced bridge or "hybrid coil" circuit employed in telephone repeaters. This could be used in the radio wire connection only at great sacrifice. In these bridge circuits a real wire line is matched against an artificial line and the balance cannot be perfect. In order to prevent voice frequency singing through the residual unbalance and around the entire radio link, it was necessary at times to reduce the amplification between the subscribers telephone set and the radiotelephone transmitter to the point where the percentage modulation was low and the side-band power output of the transmitter was much less than its full capacity.

To overcome this major obstacle, recourse again was had to a device newly worked out for wire telephone transmission. By associating and electrically interlocking several of the voice current operated switching devices which had been developed for suppressing echoes on long wire lines, an arrangement now commonly known as a "vodas"* was developed.[7] When the subscriber talks, his own speech currents, acting on the vodas, cause it to connect the radio transmitter to the wire line and at the same time to disconnect the radio receiver. When the subscriber listens the connection automatically switches back to the receiver. No singing path ever exists. The amounts of amplification in the two oppositely directed paths can be adjusted substantially independently of each other, and constant full load output from the

* This word, "vodas," is synthesized from the initial letters of the words voice operated device, antisinging.

radio transmitters is secured. With this device it became possible to connect almost any telephone line to a radio system and to adjust amplification so that a weak talker over a long wire line could operate the radio transmitter as effectively as a strong local talker.

Summarizing the above, advances over the older radiotelegraph and telephone arts which promoted the initiation of long distance commercial radiotelephony were: (1) the development of high power water-cooled tubes; (2) the application of the single-side-band suppressed-carrier method of employing them efficiently for telephony; (3) the development of quantitative methods of measuring radio transmission and noise, and the collection of information on radio transmission over the projected path which enabled engineering a system for sufficient reliability; and (4) the development of the vodas for interconnecting wire and radio circuits to give constant volume operation in the radio portion.

Developments Essential to Growth

The first long-distance radiotelephone circuit[8] operated (and still operates) between the United States and England with long wave transmission at about 5000 meters. We did not then, and we do not today, know how any considerable amount of intercontinental radiotelephony could have been accomplished with circuits of this kind. The frequency space available in the long-wave range would accommodate comparatively few channels. The high attenuation to overland transmission and the high noise level at these wave lengths precludes their satisfactory use for very great distances or in or through tropical regions. The discovery that short waves could be transmitted to the greatest terrestrial distances and could be satisfactorily received in the tropics came at a most opportune time.

Short-wave transmission not only released the limitations on distance and location inherent to long waves but also opened up such a wide range of frequency space as to give opportunity for an extensive growth in numbers of both radiotelegraph and radiotelephone circuits. In addition to removing these physical barriers, short waves further encouraged the growth[9] of radiotelephony by making it cheaper to carry on. It became possible to make directive antenna structures of moderate size which increased the effectiveness of transmission many times, thereby reducing the transmitter power required for a given reliability of communication. Short waves were the indispensable element without which material growth could not have occurred, but there were other significant things.

An important desideratum in telephony is privacy. The all-wave receiver makes no distinction between broadcast channels and commercial telephone channels. Radiotelephone communication would have been severely hampered if privacy systems had not been developed to convert speech into apparently meaningless sounds during its radio transit.

Another item of great aid in promoting growth was the development of methods of accurate stabilization of transmitted frequencies. The first effect of this was to eliminate the extreme distortion which characterized early short-wave telephone transmission and which was found[10] to be due to parasitic phase or frequency modulation effects in the transmitters. As the number of radio communication facilities, both telegraph and telephone, grew, accurate stabilization of frequency became a necessity in order to permit effective utilization of the available frequency space without mutual interference between stations.

Later Technical Advances

Among more recent technical advances a few may be mentioned.

There have been developed improved and much cheaper antenna systems. The "rhombic" antenna,[11] for example, is mechanically simple and electrically nearly aperiodic, covering a wide wave length range efficiently. It has radically changed the character of the physical plant and investment necessary to the employment of directivity in short-wave transmitting and receiving.

In Hawaii and the Philippines on circuits to the United States the "diversity"[12] method of reception is used wherein three individual separated antennas and receivers with interlocked automatic gain controls are combined to produce a common output having less distortion and noise than a single receiver.

The effects of distortion in short-wave circuits are avoided to some extent by an arrangement called a "spread side-band system,"[13] which has been used on circuits between Europe and South America. By raising the speech in frequency before modulation the speech side bands are displaced two or three kilocycles from the carrier and many of the product frequencies resulting from intermodulation fall into the gap rather than into the sidebands.

On the Holland-Java route a system is being used whereby more than one side band is associated with a single carrier or pilot frequency, each such side band representing a different communication. No publication on this system is known.†

† Since this paper was sent to the printer the author has had the privilege of reading a description of this system prepared by Professor Ir. N. Koomans of the Radiolaboratory of the Netherlands States Telegraphs. This description has been submitted for publication in the PROCEEDINGS.

An improved signal-to-noise ratio is given by a device called a "compandor"* employed[14] on the New York-London long-wave circuit. It raises the amplitude of the weaker parts of the speech previous to transmission. In depressing these raised parts to their proper relative amplitude, after reception, the compandor also depresses the accumulated radio noise.

Present Outlook

The foregoing should be enough to make it evident that many fundamental engineering problems have been solved and that the pioneering stage of the service, when its possibility of continued existence may be in doubt, has definitely been passed. In looking toward the future we find that the greatest needs are for improvement in reliability and in grade of service, accompanied by reduced costs.

Improving the reliability is hindered by the fact that short-wave transmission varies through such a wide range of effectiveness, and seems to be so much influenced by the sun. We have not only a daily cycle in the transmission of a given frequency but also an annual cycle and beyond this an eleven-year cycle associated with the change in sunspot activity. Superimposed upon these are erratic and occasionally large variations associated with magnetic storms.

A statistical study[15] of the data secured from operation of transoceanic radiotelephone circuits over the past several years has given valuable help in engineering circuits to meet a given standard of reliability. This study has shown that the percentage of lost time suffered on a circuit appears to follow a probability law and that its relation to the transmission effectiveness of the circuit in decibels is given by a straight line when plotted to an arithmetic probability scale. Such a relation tells us, for example, that if a circuit as it stands suffers fifteen per cent lost time, the lost time can be reduced to a selected lower value, say five per cent, by improving the circuit a definite amount, in the assumed case ten decibels. It then becomes possible, by making engineering cost studies of the various available ways of securing the necessary number of decibels improvement in performance, to choose the most economical one. This approach is being applied to study of the radiotelephone circuits extending outward from the United States. Some of the technical possibilities which are being considered for improving these circuits are discussed below.

The performance of a radiotelephone circuit may be changed by dynamically modifying the amplification or other characteristics of the

* The synthetic word "compandor" is a contraction of the compound word compressor-expander which describes the effects the device has on the volume range of speech.

circuit in accordance with the speech transmitted. The compandor already mentioned is an example of this kind of improvement on long waves. Further developments particularly suited to the vagaries of short-wave transmission are possible.

The operation of the vodas, or voice operated switching device linking the wire and radio circuits, is adversely affected by noise. Methods are being investigated for using single frequencies, called "control tones," transmitted alongside the speech band and under the control of speech currents, to give more positive operation of the switching devices and reduce the adjustment required.

The transmission improvement of about nine decibels (about ten-to-one in power) offered by single-side-band suppressed-carrier transmission has been delayed in its application to short-wave transmission partly because of the high degree of precision in frequency control and selectivity necessary to its accomplishment. In recent years successful apparatus[16] has been developed and proved satisfactory in trials. The introduction of single-side band into commercial usage is already in progress.

For several years past an intensive study[17] has been made of the characteristics of short-wave radiotelephone transmission, viewed from the receiving end, to find whether there may be any laws or principles which can be invoked to give improvement against fading, distortion, and noise. One fundamental way to reduce noise in radiotelephony is to employ sharper directivity. It has been found by observation that there is a limit to which directivity, as ordinarily practiced, can be carried to advantage. It is easy to design antennas so sharp that at times very large improvements in signal-to-noise ratio are secured. But it is found that at other times these antennas are actually poorer than are much less sharply directive systems. Such observations also indicate a wide variation in the performance of antennas as regards selective fading, and the signal distortion accompanying it.[18]

The result of all this work has been the development of a system based on an entirely new approach to the problem of sharp directivity and of telephone receiving. This system is called a MUSA System, the word MUSA being synthesized from the initial letters of the descriptive words Multiple Unit Steerable Antenna. Since a complete description is in process of publication,[19] only an outline of the principles and methods is given below.

In the usual method of measuring the height of the ionosphere, extremely short spurts of short-wave radiation are emitted from a transmitter and the echo reflected back from the ionosphere is picked up on a nearby receiver. The interval between the outgoing pulse and the returning echo is timed with a cathode-ray oscillograph, the time required for the round trip being taken as a measure of the distance traversed. When such spurts or pulses are sent from one side of the Atlantic, and received on the other side several echoes are usually observed. It has been found that these echoes do not arrive like successive bullets from the same gun, all following the same path. They come slanting down to the receiver from different angles of elevation or different vertical angles. This shows that with continuous wave sending, the energy comes down to the receiver in distinct streams from different vertical angles. These vertical angular directions remain comparatively stable. While the signal received at each of the individual directions may be subject to fading, the fading is somewhat slower and is not very selective as to frequency. It is found that the signal component coming in at a low angle has taken less time in its trip from the transmitter than a high angle component. Evidently the low angle paths are shorter. All these facts fit in well, on the average, with the ideal geometrical picture of waves bouncing back and forth between the ionosphere and the ground and reaching the receiver as several distinct components which started out at different angles, have been reflected at different angles and have suffered different numbers of bounces.

The ordinary directive antenna is blunt enough in its vertical receiving characteristic to receive all or nearly all of these signal components at once. Because of the different lengths of time the various components have taken in their flights over different paths from the transmitter they are at cross purposes and do not mix well but clash and interfere with one another. This shows up as the selective fading and distortion which characterizes short-wave reception much of the time. The MUSA method remedies this trouble.

The MUSA system provides extremely sharp directivity in the vertical plane. By its use a vertical angular component can be selected individually. It consists of a number of rhombic antennas stretched out in a line toward the transmitter and connected by individual coaxial lines to the receiving apparatus. The apparatus is adjustable so that the vertical angle of reception can be aimed or "steered" to select any desired component from the others, as a telescope is elevated to pick out a star. The antennas remain mechanically fixed. The steering is done electrically with phase shifters in the receiving set. By taking several branch circuits in parallel from the antennas to different sets of adjusting and receiving apparatus the vertical signal components may be separated from each other.

Nature breaks the wave into several components and jumbles them together. The first function of the MUSA system, as just described, is to sort the components out again. Its second function is to correct their differences so that they may be combined smoothly into a replica of the original signal. To do this the received wave components are separately detected and passed through individual delay circuits to equalize their differences in transit time. They are then combined to give a single output. As compared with a simple receiver the MUSA receiving system gives: (1) improvement in signal-to-noise ratio, as a result of the sharp directive selectivity of the antenna; (2) improvement against selective fading distortion, by virtue of the equalization of the time differences between the components before they are allowed to mix; and (3) improvement against noise and distortion, because of the diversity effect of combining the several components.

It is found that the directive selection and the delay compensation adjustments correct for one frequency are satisfactory for a considerable band of frequencies adjacent thereto. Thus there is offered the possibility of receiving a number of grouped channels through one system and the prospect appears not only of improved transmission but also of reduced cost per channel.

The possibility of grouping channels at the transmitting station may be conceived on the basis of either "multiple" or "multiplex" transmission. In the multiple arrangement each channel has its own antenna and its individual transmitter whose frequency is closely spaced from and coordinated with the adjacent channels of the group. In "multiplex" transmission, the channels are aggregated into a group at low power and handled *en bloc* through a common high power amplifier and radiating system. Particularly in the multiplex case, there are possibilities of important economies if the technical problems are satisfactorily solved. Passing a multiplicity of channels simultaneously through a common power amplifier involves interchannel interference due to modulation products which is not met with when only one channel is present. Severe requirements are thereby placed on the distortion characteristics of the power amplifier.

It seems a fair conclusion that the tendency in the engineering solution of the problems of economy and growth in radiotelephone development (and perhaps also radiotelegraph development) will be toward channel grouping methods, especially for backbone routes between important centers where large traffic may develop. This will be a considerable departure from past practice which has resulted in the existing system of scattered frequency assignments. It is to be hoped that the obvious difficulties in rearranging frequency assignments will not prove so unyielding as to preclude putting new engineering developments into service.

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NOTES ON SOME PRACTICAL COMPARISON TESTS MADE BETWEEN SEVERAL ACOUSTIC MEASUREMENT METHODS*

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Summary—*In considering the various methods used in making over-all acoustic measurements on radio receivers, from the standpoint of possible standardization, it was felt desirable to make a comparison of the results obtainable by the several principal methods now in use in this country. Accordingly a radio receiver was shipped successively to five different laboratories equipped to make such measurements. It was checked between each shipment by a sixth laboratory to guard against changes caused by the shipments.*

The test equipment setup and method used by each of the co-operating laboratories is described. The results obtained are discussed briefly and typical curves made by each laboratory are shown.

The author has refrained from drawing any arbitrary conclusions regarding the relative accuracy or effectiveness of the various test methods employed by the co-operating laboratories. Instead, the comments, which the engineers of each laboratory cared to make on the curves obtained and the methods used, have been given. The reader is thus provided with data and a valuable series of comments from some of the most expert engineers in this field, from which he may arrive at his own conclusions.

In general it is felt that the curves show a greater degree of similarity than might have been expected considering the rather fundamental differences in the test methods used, and that certain possibilities in the direction of useful standardization in this field have been indicated.

A STANDARD method of taking over-all acoustic measurements on radio receivers has long been felt desirable. Some time ago in connection with some standardization work, this desire took concrete form since it was necessary to draw up rather quickly an outline of measurements methods which would be widely acceptable. The radio-frequency measurements were quite well standardized and accepted as outlined in the then current report of the I.R.E. Standards Committee. When the subject of over-all acoustic measurements came up, however, there was no standard available which covered the ground and it appeared necessary to decide upon a measurement method from among those already in use in the several laboratories making measurements of this kind. Brief study of these showed marked differences in the methods used and no available data to tell which might have the greatest accuracy or even whether any of the methods could reasonably be expected to give equivalent results.

* Decimal classification: R261.3. Original manuscript received by the Institute, April 7, 1937.

It occurred to the author that possibly some useful data could be obtained on this subject by having a receiver circulated to the various laboratories who were set up for making over-all acoustic measurements so that each might measure it. The curves obtained could be compared and a possible answer to the problem might then be indicated. Even if this did not point the way to an immediate standard method, it was felt that at least some interesting and valuable data would thus be obtained concerning variations between methods, since the device under test would be the same in all cases. The variables would thus be reduced to only two; i.e., the test equipment used and the acoustic surroundings in which the test was made.

Accordingly a console model receiver was selected and all laboratories who might be interested were contacted as to whether they would like to make measurements on it. The receiving set selected was an RCA Model 280. This was not designed as a high fidelity model but was believed to represent reasonably good fidelity for its price class at the time it was selected. It was selected partly because of being easily available at the time. It was felt that it covered a sufficient range of high and low frequencies to be useful for the circulation tests contemplated.

The groups who co-operated in the test were the Ballantine Laboratories, Inc., Boonton, N. J., the General Electric Company, Bridgeport, Conn., the Hazeltine Service Corporation Laboratory, Bayside, L. I., N. Y., the Philco Radio and Television Corporation, Philadelphia, Pa., and the Stromberg Carlson Telephone Manufacturing Company, Rochester, N. Y.

From the RCA Victor Division, of the RCA Manufacturing Company in Camden two groups co-operated; i.e., the research department and the receiver department.

The circulation procedure was briefly as follows:

The receiver was first measured by the RCA receiver department. It was then shipped to one of the co-operating laboratories who would measure it and return it to Camden. The RCA receiver department would then remeasure the receiver before shipping it out to the next co-operating laboratory. In this way a check was kept continually on the receiver to make certain that it did not change its over-all acoustic characteristics due to tube, circuit, or other changes resulting from the frequent shipments. It might be mentioned at this point that the receiver remained sufficiently unchanged in its over-all acoustic characteristics throughout these tests so that no tube replacements, circuit realignments, or other changes were necessary.

Measurement Methods

In order to minimize the variables that might be introduced by the use of different test signal and receiver control settings in the various laboratories the test conditions given below were recommended. Where it was impossible to adhere to these because of peculiarities of the test equipment or other conditions, the differences in test conditions were recorded and the necessary corrections were made in the curves to make them comparable with those taken as outlined below:

Input radio-frequency signal.—Ten millivolts at 1000 kilocycles, modulated thirty per cent and impressed on a standard dummy antenna circuit.

Setting of receiver controls.—Sensitivity control set for maximum sensitivity and tone controls were set for maximum fidelity. Since the manual volume control was of the automatic tone compensation type it was set at a definite point to avoid fidelity changes.

The other test arrangements were left to each of the co-operating laboratories, but each was requested to describe its equipment and test conditions. Below will be found the essential items of these descriptions:

Measurement Conditions Used by the Ballantine Laboratories, Inc., Boonton, N. J.

Measurement location.—Outdoors.

Microphone, test equipment, and methods.—Specially constructed pressure operated condenser microphone fixed in position. Other equipment consisted of the automatic logarithmic recorder which has been described elsewhere.[1]

Microphone location versus receiver.—First position with receiver on the ground with no back wall. Microphone distance was four feet and height about 45 inches above ground and in the vertical plane passing through the axis of the loud speaker.

Second position with receiver at a sufficient distance above ground to avoid modification of the curves by earth reflection. Microphone distance two feet and directly in front of the loud speaker and on its axis.

Ballantine also took curves in a sound room and a living room but since, as will be noted from his comments near the end of this paper, he favors the outdoor measurements, we have not included these indoor curves.

[1] Stuart Ballantine, *Jour. Acous. Soc. Amer.*, vol. 5, pp. 10–24; July, (1933). See also Proc. I.R.E., vol. 23, pp. 618–652; June, (1935).

Measurement Conditions Used by the General Electric Company, Bridgeport, Conn.

Measurement location.—This was a "soundproofed" room of irregular shape, the two long walls being respectively 16 feet 6 inches and 21 feet 6 inches and parallel at a distance 14 feet apart. One end wall is at right angles to the two long walls and the other end wall is at the necessary angle to connect the two long walls. The walls, ceiling and floor of the room are lined with five inches of rock wool.

Microphone, test equipment, and method.—A Brush crystal microphone is used, calibration of which was supplied by the manufacturer and substantiated by checks which the General Electric Company have made on it. The output of the microphone is fed into an automatic sound pressure recorder, the recording drum of which is linked with the frequency control of the oscillator which supplied the modulation for the carrier.

Microphone location versus receiver.—The receiver is placed on a table 30 inches high, whose top is 30 inches long and extends 24 inches from the wall. The table is placed at the center of the 14-foot room wall. The microphone is placed on the axis of the loud speaker and at a distance three feet from the front wall of the cabinet. The back of the receiver is placed two inches from the room wall.

Measurement Conditions Used by the Hazeltine Service Corporation Laboratory

Measurement location.—Room of about 2000 cubic foot volume, having acoustic characteristics approximating those of a typical living room.[2]

Microphone, test equipment, and method.—A nondirectional two-element crystal microphone is used at a height of three feet above the floor (approximately the average ear height of a seated listener). The frequency was wobbled by an amount not exceeding plus or minus (ten cycles + five per cent).

Microphone location versus receiver.—Out of a series of several curves taken by Wheeler on the test receiver, the author has chosen a group taken with the receiver located at the center of a side wall. The distance from microphone to that wall was ten feet for the center curve and eight feet for each of the side curves.

Measurement Conditions Used by the Philco Radio and Television Corporation, Philadelphia, Pa.

Measurement locations.—Two locations were used, the first being

[2] See paper by Wheeler and Whitman, Proc. I.R.E., vol. 23, pp. 610–618; June, (1935), for further details on room, measuring equipment, and method.

outdoors on the roof of the Engineering Building. The second was a "soundproofed" room of 20 by 30 feet by 9½ feet high. The walls of this room are covered with two two-inch thicknesses of rock wool with a two-inch air space between them. The ceiling is of similar construction. The floor consists of expanded metal placed on top of four inches of loose rock wool. The room is constructed so as to float on felt and the ceiling beams hang from felt. An artificial back wall is used made of wood 1½ inches thick and six feet square.

Microphone, test equipment, and method.—A two-cell crystal microphone is used associated with a semiautomatic sound pressure recorder. The recorder drum is linked with the frequency control of the audio-frequency oscillator which supplies the modulation for the carrier.

Microphone locations versus receiver.—In the outdoor test the receiver was set on the 28 by 48-inch floor of an elevator and raised 20 feet above the roof. The microphone was on the axis of the loud speaker 5½ feet from the front of the loud speaker baffle. No back wall was used.

In the indoor test the receiver was placed on the expanded metal floor with its back two inches from the back wall. The microphone was 45 inches above the floor in the vertical plane through the axis of the loud speaker and five feet in front of the loud speaker baffle.

Measurement Conditions Used by the Stromberg-Carlson Telephone Manufacturing Company, Rochester, N. Y.

Measurement location.—"Soundproofed" room 22 by 15 feet by 11 feet high. All surfaces covered with one-inch hair felt and with additional hangings of the same felt. Computed reverberation time (Eyring formula) 0.1 second.

Microphone, test equipment, and method.—Western Electric condenser microphone calibrated by Rayleigh disk. The microphone is rotated at about 25 revolutions per minute in a circular orbit eight feet in diameter and inclined 45 degrees to the axis of the loud speaker. The center of the orbit is 53 inches above the floor and four feet six inches from one end of the room. The axis of the microphone is maintained parallel to the axis of the loud speaker at all times.

The method is essentially as described by Bostwick.[3] Low-pass filters are used up to 5000 cycles. The loud speaker is fed from approximately twice its own nominal impedance. Measurements are made at about ten frequency points per octave. The sound pressure curve is added to the output voltage curve taken across five ohms using a modulated signal at the input of the receiver. At frequencies of 70

[3] *Bell Sys. Tech. Jour.*, vol. 8, pp. 135–158; January, (1929).

cycles and below the loud speaker curves were corrected for measuring room effects by predetermined factors obtained by free space measurements. A check which was later made on the receiver indicated a slight discrepancy in the curves because of the particular impedance values which had been used in the above-mentioned tests. It was found that under the conditions above described the loud speaker and receiver output circuit performance characteristics were quite critical to the particular impedance conditions used, and a correction has been made on the curves provided by Stromberg-Carlson to compensate for this, and make the curves more comparable in loud speaker operation conditions to the other curves given in this article.

Microphone location versus receiver.—A back board and floor board were used extending approximately one foot beyond the receiver. The back of the receiver was placed four inches from the back board and the front of the receiver was placed ten feet from the center of the microphone orbit. The axis of the speaker was 53 inches from the floor and pointed at the center of the microphone orbit.

Measurement Conditions Used by the Research Department of the RCA Manufacturing Company, Camden, N. J.

Measurement location.—Outdoors on roof of engineering department building.

Microphone, test equipment, and method.—RCA Type 44A velocity microphone with a free wave calibration. Recorder was of semiautomatic type with audio-frequency oscillator fed directly to the loud speaker and the voltage across the voice coil recorded. Knowing the voltage characteristic under these conditions, the necessary correction was applied for the voltage obtained with standard modulated signal, the latter characteristic being obtained by a separate test.

Microphone location versus receiver.—One test was made with the receiver standing on the surface of the roof and with no back wall. The microphone distance was four feet and the microphone height 45 inches above the roof and in the vertical plane passing through the axis of the loud-speaker. The microphone was not tipped to eliminate "floor reflection."

A second test was made with the receiver raised ten feet above the roof. In this case the microphone distance was two feet directly in front of the loud speaker and on its axis. The cabinet was in its normal upright position.

Measurement Conditions Used by the Receiver Department of the RCA Manufacturing Company, Camden, N. J.

Measurement location.—This is a "soundproofed" room 50 by 32 feet by 11 feet high. The measurements are made near one end of the room but the back wall is an artificial one made of hard wood approximately six feet square. The room itself is covered on ceiling and all four sides with one-inch thick balsam wool behind which is one-half inch thick Celotex and behind this a double Pyrobar wall, the two sections of which are spaced one foot apart. The floor is of hard wood built up on the regular building floor and covered with one-half-inch Celotex.

Microphone, test equipment, and method.—The microphone is a General Electric Company condenser microphone with a free wave calibration. The output of the microphone is fed into a semiautomatic sound pressure recorder, the recorder drum being linked with the frequency control of the audio-frequency oscillator which supplies the modulation for the carrier.

The microphone is rotated approximately 25 revolutions per minute in a circular orbit approximately six feet in diameter and inclined at an angle of 30 degrees to the horizontal. The center of the orbit is 54 inches above the floor. The axis of the microphone is maintained parallel to the axis of the loud-speaker at all times.

Microphone location versus receiver.—The receiver was placed on a flooring made of hard Masonite four by six feet. The back of the receiver was two inches from the artificial back wall and at a point equidistant from the two ends of that wall. The back wall was approximately eleven feet from the center of the microphone orbit and the center of the orbit was in the vertical plane through the axis of the loud speaker.

Results Obtained

A rather large number of curves were received from the various cooperating laboratories, some of which were taken under conditions that made it difficult to compare them with those of the other laboratories. A selection was made of those curves taken under conditions which were sufficiently similar to warrant comparison. These curves were replotted to the same scales and are shown in Figs. 1 to 8 inclusive. The ordinates are 1.0 decibels per division. For convenience the conditions under which each curve was taken are given in brief form in the accompanying table. Curves which were felt to show interesting degrees of similarity have been superimposed for convenient comparison.

In particular the several outdoor curves shown in Figs. 1 and 2 are of interest because of the very close similarity in the results obtained. This was, of course, to be expected since the taking of the curves outdoors removes the troublesome effect of room characteristics. There seems to be little doubt but that this method is desirable where a close check of the curves is wanted. Difficulties of taking sound pressure measurements outdoors such as, the local noise level, weather condi-

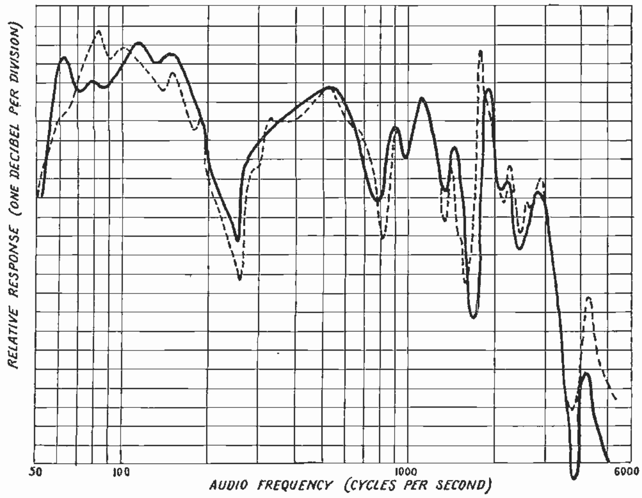

Fig. 1

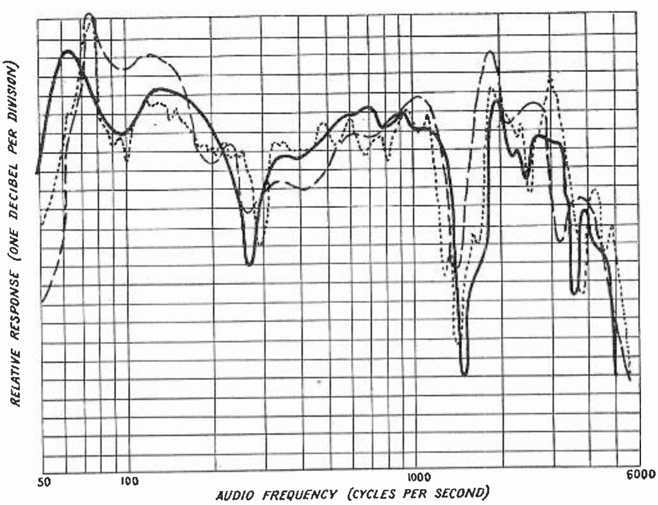

Fig. 2

tions, etc., make this method undesirable for routine acoustic measurement work.

Of the curves taken indoors, the two which showed the greatest similarity were those taken by the Hazeltine Laboratories and the RCA receiver division. For this comparison the solid line (central microphone position) curve of the three Hazeltine curves shown in Fig. 4 was taken. This results in somewhat too much high-frequency response since this curve is taken on the axis of the loud speaker and

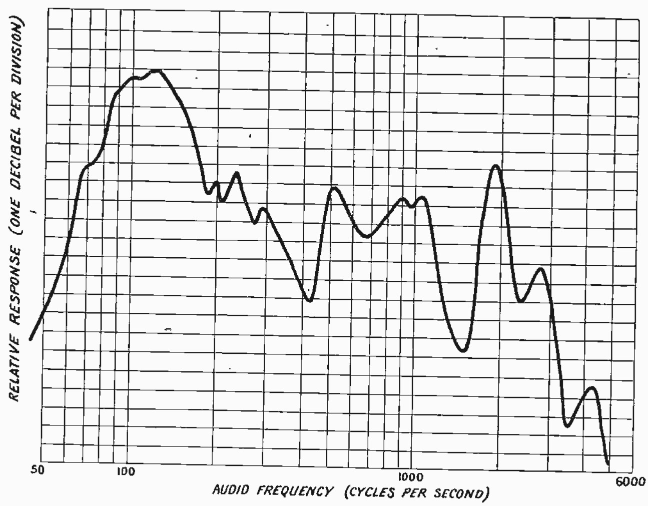

Fig. 3

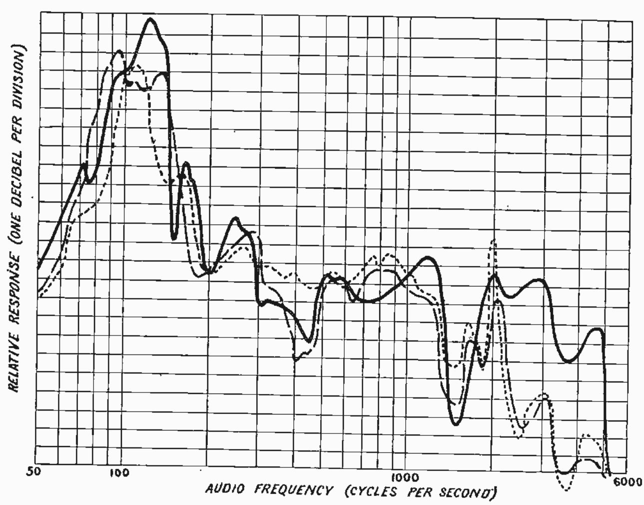

Fig. 4

this explains the discrepancy in the two curves of Fig. 7 at the high-frequency end of the range. It is possible that the discrepancy in the low-frequency end of the range is due to the differences between the two rooms since these were quite dissimilar.

Comments

The author has received comments on these curves from some of the co-operating laboratories and these are included below since it is

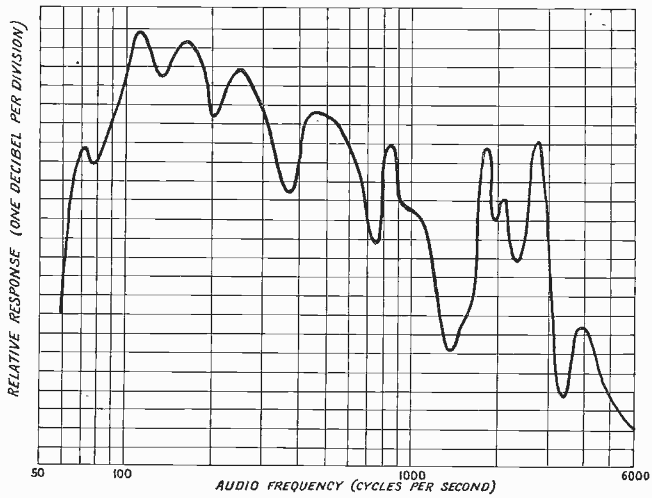

Fig. 5

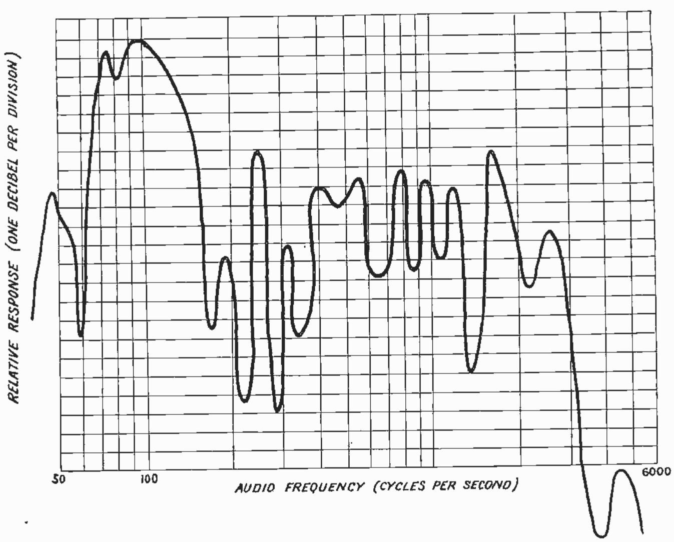

Fig. 6

felt that they contribute points of definite importance to this whole subject.

Comments from Stuart Ballantine of Ballantine Laboratories, Inc.:

"We have been interested in this subject for a number of years and believe that this interlaboratory comparison is a valuable step toward the formulation of a standard testing procedure.

"The technique employed in our measurements of this receiver has been described in detail in the literature.[4,5,6,7,8,9,10,11]

"A pressure operated fixed microphone was employed whose free wave calibration had been performed by means of a Rayleigh disk. The receiver was supplied with a signal of thirty per cent modulation and variable frequency at 900 kilocycles. The method of registering the

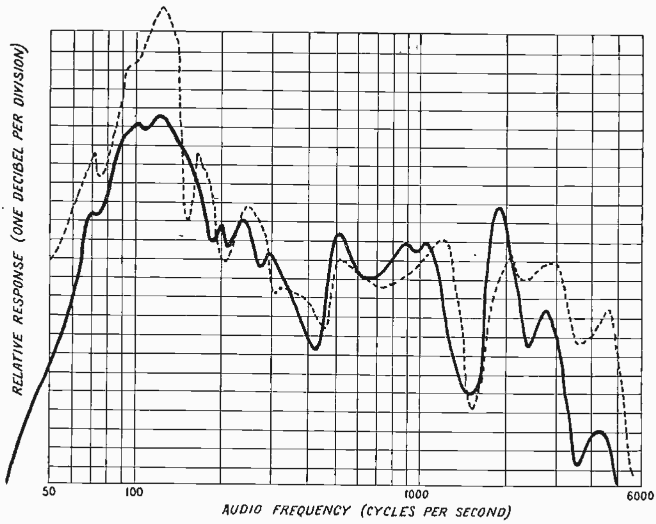

Fig. 7

sound pressure frequency characteristic is fully automatic and has been described.[7,9,11]

"Our experience has been that the best conditions for this kind of testing are to be found outdoors and this seems to be in concordance with the results of this intercomparison. I would expect pretty close agreement between fixed, properly calibrated pressure operated microphones under the same conditions outdoors, that is to say, with the same relative microphone positions and distances from the ground. In comparing curves taken by velocity and pressure operated microphones

[4] *Phys. Rev.*, vol. 32, p. 988; December, (1928).
[5] Proc. I.R.E., vol. 16, pp. 1639–1644; December, (1928).
[6] Proc. I.R.E., vol. 18, pp. 1206–1215; July, (1930).
[7] *Electronics*, vol. 1, p. 472; January, (1931).
[8] *Jour. Acous. Soc. Amer.*, vol. 3, p. 319; January, (1932).
[9] *Jour. Acous. Soc. Amer.*, vol. 5, p. 10; July, (1933).
[10] Proc. I.R.E., vol. 22, p. 564; May, (1934).
[11] Proc. I.R.E., vol. 23, pp. 618–652; June, (1935).

it should be borne in mind that at distances where reflection from the ground is of importance the reflected wave and direct wave are additive algebraically for the pressure operated microphone whereas they are additive vectorially for the velocity operated microphone (assuming vertical ribbon); also close to the loud speaker the pressure and velocity may not bear the same simple relationship to each other which is found in an approximately plane wave at greater distances.

"Due to the inclemency of the weather it is often necessary to work indoors. With regard to measurements carried out in specially treated

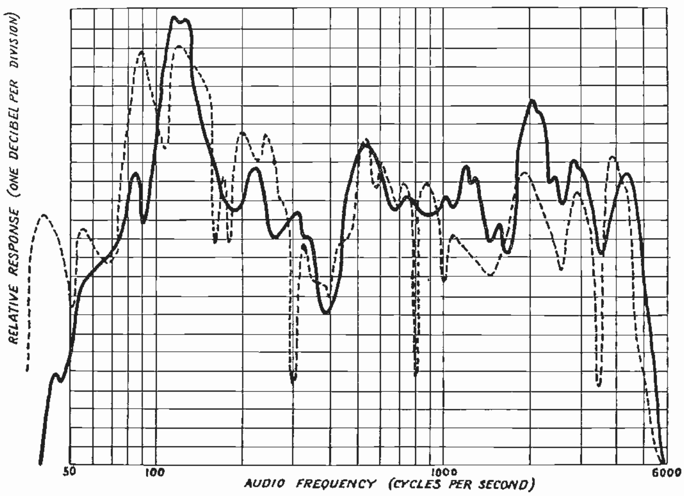

Fig. 8

sound rooms we have found that measurements sufficiently close to the loud speaker, say at two feet, agree pretty well with the outdoor curves.[11] I think this is the next best standard condition that can be adopted. Measurements made in such rooms under other conditions are usually too much influenced by the room to give comparable results.

"Our experience with measurements in standard living rooms has been pretty fully outlined.[11] In general the low-frequency picture is considerably influenced by the room and by the location of microphone and loud speaker. At high frequencies the room has less effect and the results depend more on the position of the microphone with respect to the axis of the loud speaker. A difference will also be observed between curves taken with a microphone which shows diffractive directivity and a nondirective microphone such as a single small sound cell, especially at the higher frequencies. I would not expect any but fortuitous agreements between living room curves taken by different laboratories. This does not of course mean that curves taken in a standard living room are of no absolute value; on the contrary they are of considerable engineering utility when properly taken and interpreted.

"I would recommend that the best standard condition would be, in the order of desirability, as follows:

(a) Outdoors with a fixed microphone with distances between receiver and microphone and receiver and ground sufficient to avoid the effect of reflection from the earth.

(b) A fixed pressure operated microphone, preferably with some diffractive directivity, in an acoustically treated room, the dimensions of the room being large enough and the microphone close enough to the loud speaker to reduce reflection effects to the order of plus or minus two decibels."

Comments from I. J. Kaar and H. Roder of General Electric Company, Bridgeport, Conn.:

"As far as our measurements are concerned we are pleased to note that you could duplicate the result very nearly after having arranged for the same condition of test. (See Fig. 8.) The difference in the low-frequency peak I believe is due to the difference in room volume and room geometrics because you probably could not provide a room which is an exact duplicate of the room we are using. Rooms of different volume and configuration, however, have different "Eigen" frequencies and this probably is the reason why the curves taken by you and by us differ in the low-frequency measurement."

Comments from Harold A. Wheeler of the Hazeltine Service Corporation Laboratory, Bayside, L. I., N. Y.:

"At the lower frequencies we do not regard any single curve as sufficiently representative and therefore we usually estimate the average of the three curves in different directions as indicative of the low-frequency performance. We have no regular procedure for drawing the average of the curves although a line drawn as the average of the three curves would probably be adequate.

"At the higher frequencies each individual curve is reliable for what it purports to show. Probably the only objection to showing only one curve is the fact that the information is incomplete. This is not a serious objection if the main purpose is to make comparisons with the observations of other laboratories."

Comments from David Grimes and R. S. Fisher of the Philco Radio and Television Corporation, Philadelphia, Pa.:

"Examination of the curves made under similar acoustical conditions shows fair agreement even though the microphones were different

SUMMARY OF MEASUREMENT CONDITIONS

Figure No.	Name of Laboratory	Type Microphone	Mike Calibration	Position of Mike Relative to Speaker	Speaker to Mike Distance	Test Location	Surroundings		Distance from Cabinet to Back Wall	Other Conditions
							Back	Floor		
1 Solid	RCA Research Department	Velocity	Free wave	45″ up in plane of speaker axis	4′	Outdoors	No	Yes		Sound pressure measured and combined with overall *r-f* voltage curve by calculation
1 Dashed	Ballantine Laboratories	Condenser	Free wave	45″ up in plane of speaker axis	4′	Outdoors	No	Yes		
2 Solid	RCA Research Department	Velocity	Free wave	On speaker axis	2′	Outdoors	No	No		Cabinet raised above roof. Sound pressure measured and combined with over-all *r-f* voltage curve by calculation.
2 Short Dash	Ballantine Laboratories	Condenser	Free wave	On speaker axis	2′	Outdoors	No	No		Cabinet raised above ground
2 Long dash	Philco	2-element crystal	Free wave	On speaker axis	$5\frac{1}{2}$′	Outdoors	No	Small floor 28″×48″		Cabinet raised above roof.
3	RCA Receiver Division	Condenser	Free wave	Center of mike orbit in vertical plane of speaker	11′	Sound-proof room	Yes	Yes	2″	Mike rotating through 6′ diameter orbit inclined 30° to floor.
4 Solid	Hazeltine Laboratories	2-element crystal nondirectional	Free wave	On plane of axis of speaker and 36″ up	10′	Standard living room	Yes	Yes	2″	Modulator frequency wobbled
4 Two Dashed	Hazeltine Laboratories	2-element crystal nondirectional	Free wave	36″ up and 37° each side of center line	8′	Standard living room	Yes	Yes	2″	Modulator frequency wobbled.
5	Stromberg Carlson	Condenser	Free wave	Center of mike orbit on speaker axis	10′	Sound-proof room	Yes	Yes	4″	Mike rotated through 8′ diameter orbit inclined 45° to floor. Over-all *r-f* measured separately and combined with acoustic curve.
6	Philco	2-element crystal	Free wave	45″ up in plane of speaker	5′	Sound-proof room	Yes	No	2″	
7 Solid same as Fig. 3	RCA Receiver Division	Condenser	Free wave	Center of mike orbit in vertical plane of speaker	11′	Sound-proof room	Yes	Yes	2″	Mike rotating through 6′ diameter orbit inclined 30° to floor.
7 Dashed same as solid curve of Fig. 4	Hazeltine Laboratories	2-element crystal nondirectional	Free wave	On plane of axis of speaker and 36″ up	10′	Standard living room	Yes	Yes	2″	Modulator frequency wobbled.
8 Solid	RCA Receiver Division	Condenser	Free wave	On speaker axis	2′	Sound-proof room	Yes	Yes	2″	Cabinet on 30″ high table.
8 Dashed	General Electric company	Crystal	Manufacturer's Calibration	On speaker axis	3′	Sound-proof room	Yes	Yes	2″	Cabinet on 30″ high table.

and probably calibrated independently. Outdoor curve Fig. 2 is an example—three different microphones were used: velocity, condenser, and crystal. This curve indicates uniformity of microphone calibration and also that consistency can be expected if measurements are made in surroundings with negligible reflection, as is the case out of doors. If some standard of measuring conditions is to be decided upon, it seems logical at the present time to select the outdoors for surroundings, but a measurement of this type is only an indication of the response at one point in the sound field. It must be remembered, however, that a single outdoor measurement is not sufficient to analyze completely the acoustic output of a receiver and that the final criterion is its performance under actual working conditions."

Comments from Benjamin Olney of the Stromberg-Carlson Telephone Manufacturing Company, Rochester, N. Y.:

"The fact that our measurements were the only ones to employ a filter in the microphone circuit may have some bearing on the differences, especially when compared with the measurements made by your receiver division by the rotating microphone method. It is probable also that the great difference in the size of the measuring rooms in this latter case had considerable influence on the results.

"While it is, of course, highly desirable that the curves of different laboratories agree, lack of such agreement does not materially detract from the value of indoor response measurements to the engineer who uses them as a development tool. Although no single curve can completely define the performance of a loud-speaker it is, however, believed to be true that by a large amount of comparative measuring and listening under fixed conditions one eventually acquires experience which enables him to extract a great deal of useful information from a single curve made by the method with which he is familiar. If such a method is capable of disclosing response irregularities and of determining the significant frequency range of a loud-speaker, it will serve a useful purpose, and the adjustment of the *trend* of the curve may well be left to listening tests for decision. It is believed that indoor measurements made by any method are of dubious value in determining the latter characteristic because of the influence of directivity and of selective absorption at the walls of the room."

Comments from H. F. Olson of the research division of the RCA Manufacturing Company, Inc.:

"The results of these tests indicate that outdoor tests taken under the same conditions can be duplicated. For this reason it seems that outdoor response curves should be used for comparison between laboratories. As pointed out in the report, there are many difficulties connected with outdoor measurements, and therefore routine and acoustic testing must be carried on indoors. Since it is not always convenient to obtain outdoor response curves it seems that one way to obtain better correlation between laboratories would be for each laboratory to calibrate its indoor measuring equipment using outdoor measurements as a standard. This calibration could be used when comparisons are made between different laboratories.

"With reference to the amount of disagreement found in the curves taken outdoors we feel that the difference to be expected between a velocity and a pressure microphone when ground reflection is present would be less than the difference found between the curves shown and that it can therefore be disregarded. In using the velocity microphone no attempt was made to eliminate the ground reflection by suitable microphone orientation."

Comments from C. O. Caulton of the receiver division RCA Manufacturing Company, Inc.:

"I believe it is significant that there is the agreement that we find, considering the great variety of conditions of measurement. The fact that the outdoor curves checked as well as they do is to be expected and any *primary* standard set up between laboratories should undoubtedly use the outdoor measurement with other conditions specified properly.

"The indoor curves of a given laboratory could then be corrected against these outdoor measurements. However, the fact that certain of the indoor measurements checked closely even when no attempt was made to duplicate exactly the conditions, indicates that if conditions of setup were chosen having in mind maximum reduction of room effects (working, for instance, as close as possible to the cabinet and using equivalent floor and wall conditions), simplicity of setup and similar microphone calibration and placement, that very close agreement could be expected. I believe such a standard would be very desirable."

Comments from S. V. Perry of the receiver division, RCA Manufacturing Company, Inc.:

"In setting up a proposed standard for over-all acoustic measurements, many of the conditions chosen are necessarily quite arbitrary. However, there are other conditions which should always be observed, particularly those acoustic factors which tend to affect the motion of the reproducer diaphragm. It is to be noted that the various laboratories used entirely different conditions of measurement, particularly with regard to the immediate surroundings of the receiver under test. These factors could easily be standardized, and such standardization should be effected as a first step toward more complete agreement between different laboratories. We refer particularly to the location of the receiver with respect to the wall behind it and to the floor under it. In the great majority of applications, radio receivers are used standing on a hard, unpadded floor pushed back to a hard wall parallel to the back of the cabinet and located a relatively short distance (of the order of two inches) away from it. We believe these conditions should be adopted as standard for all such acoustic measurements, whether indoor or outdoor and regardless of whatever other conditions of measurement may obtain. The extent of the back wall and floor to be used for the measurement are of some importance and should be given due consideration."

The author would appreciate very much receiving any additional comments on these curves which may occur to readers of this article. The difficulties caused by differences in measurement method and surrounding conditions have prevented the adoption of a standard over-all acoustic measurement method for the radio art and it is hoped that the above compilation of data may be of some assistance in the direction of determining upon methods which are capable of giving comparable results in different laboratories.

Acknowledgment

The author wishes to express his sincere appreciation of the efforts and time contributed to this investigation by the laboratories and engineers who co-operated in it.

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FREQUENCY MULTIPLICATION AND DIVISION*

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Summary—*Constantly increasing demand for synchronizing devices in telecommunication techniques has involved extended use of devices for the production of harmonics and subharmonics of an input master frequency. This article describes a new circuit for a generator by means of which frequency multiplication and division can be obtained The principle of feedback is applied in a new way to a valve circuit having a nonlinear response characteristic. The conditions for optimum output power are deduced and the advantages of the new circuit are discussed.*

I. Introduction

In modern telecommunication techniques the use of devices for synchronizing electrical and mechanical processes of different kinds is constantly increasing. In installations for multiple telegraphy or telephony on overhead lines or cables, where it is important that the carrier frequencies of different channels are fixed in relation to each other, different kinds of synchronizing apparatus are used. Such apparatus is particularly needed when the carrier is suppressed at the transmitter and thus must be added locally at the receiver. The same is true in wireless techniques, both for long-wave and short-wave operation, as, e.g., for double modulation used in transatlantic telephony or telegraphy. Synchronization is of paramount importance for wireless transmitters which use the same carrier frequencies, and finally in television a sharp definition of the picture and the suppression of flutter depend to a great extent on perfect synchronization.

Synchronizing devices are of many different kinds according to the different requirements imposed in each case. Thus all descriptions are found, from the simple synchronizing of two generators or motors connected to alternating-current mains to the complicated installations for the generation of multiple frequencies used in laboratories, where a tuning fork or a quartz crystal produces a time-controlled master frequency which governs a number of multiple generators generating one or several measuring frequencies. The relative accuracy of each of these frequencies is exactly the same as that of the master frequency.

When a synchronizing frequency is to be transmitted by wire or wireless, it is often advantageous to locate this frequency in a band of

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the frequency range, which is not taken up by the transmission of the communication proper, as, e.g., telegraph code, telephone calls, wireless programs, or television signals. A synchronizing frequency transmitted in this way, however, may not be used directly for its purpose but is used in the form of a master frequency for controlling apparatus from which suitable frequencies for modulating or demodulating the signals are tapped. There are several apparatuses of this kind; viz., multiple generators, multivibrators etc. The functioning of a multiple generator is explained most easily if compared with a harmonic generator, although it must be remembered that it is no oscillator in the proper sense of this word. Considering this similarity it is evident, however, that a multiple generator may only produce oscillations having frequencies which are whole-number multiples of the master frequency. This latter must, therefore, be located lower in the frequence range than all the frequencies produced. There are, however, no difficulties in synchronizing oscillations having lower frequencies than the master frequency. For this purpose a harmonic of the output frequency is synchronized with the master frequency. A typical apparatus of this kind is the multivibrator.

The requirements of an arrangement for generating multiple or submultiple frequencies controlled by a given master frequency may be briefly summed up as follows:

1. The device must function without inertia and without loading the feeding circuit for the master frequency; this limits the number of possible solutions to devices containing electronic valves.

2. In each individual case it must be possible to draw energy from the output circuit of the arrangement, which may have tapping circuits for either one or several frequencies.

3. The frequencies of the oscillations produced must follow the master frequency, even if this varies within wide limits; this condition must be fulfilled without necessitating any alteration of the connections or tuning of the output circuits.

4. In each output circuit an oscillation of pure sinusoidal form must be produced.

A circuit which fulfills the above conditions very satisfactorily has been introduced by the author. This circuit permits the production of multiple as well as submultiple frequencies and differs distinctly from that of the multivibrator.

II. Fundamental Principles

The fundamental principles of the method for the production of oscillations with multiple or submultiple frequencies to a master frequency will first be explained with reference to the circuit diagram Fig. 1.

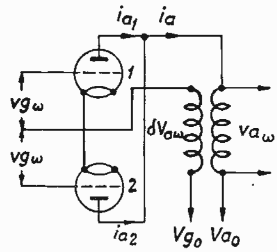

Fig. 1. Circuit diagram of compounded generator.

Compoundation

We designate by $V_c = V_g + V_a/\mu$ the voltage which controls the electronic current through the valves 1 and 2 and get thus the anode currents

$$\left.\begin{aligned} i_{a1} &= k_0 + k_1 V_{c1} + k_2 V_{c1}^2 + \cdots \\ i_{a2} &= k_0 + k_1 V_{c2} + k_2 V_{c2}^2 + \cdots \end{aligned}\right\} \tag{1}$$

where k_0, k_1, k_2, $\cdots$, are constants for the type of valves used. We then split the voltages V_c, V_g, and V_a, in direct- and alternating-current components with the indexes 0 and ω respectively and obtain

$$\left.\begin{aligned} V_{c1} &= V_{g0} + \frac{V_{a0}}{\mu} + v_{g\omega} + \frac{v_{a\omega}}{\mu} + \delta v_{a\omega} \\ V_{c2} &= V_{g0} + \frac{V_{a0}}{\mu} - v_{g\omega} + \frac{v_{a\omega}}{\mu} + \delta v_{a\omega} \end{aligned}\right\}. \tag{2}$$

The generator is characterized by the following conditions

$$\left.\begin{aligned} V_{g0} &= -\frac{V_{a0}}{\mu} \\ \delta &= -\frac{1}{\mu} \end{aligned}\right\}. \tag{3}$$

We then get

$$\left.\begin{aligned} V_{c1} &= v_{g\omega} \\ V_{c2} &= -v_{g\omega} \end{aligned}\right\} \tag{4}$$

and therefrom

$$i_a = i_{a1} + i_{a2} = 2k_0 + 2k_2 v_{g\omega}^2 + 2k_4 v_{g\omega}^4 + \cdots$$

or simplified

$$i_a = m_0 + m_2 v_{g\omega}^2 (+ m_4 v_{g\omega}^4 + \cdots). \tag{5}$$

The anode current is thus independent of each reaction from the anode circuit and its intensity is determined only by even powers of the grid alternating voltage. In the first approximation the anode alternating current is illustrated by a parabolic curve, Fig. 2.

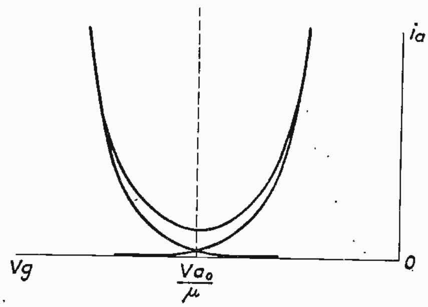

Fig. 2. Response characteristic of compounded generator.

It will now be interesting to discuss the significance of (3) and (4). For any three-electrode valve the anode alternating voltage

$$v_{a\omega} = -i_a Z = -\frac{\mu v_{g\omega}}{R_i + Z} Z \tag{6a}$$

or

$$\frac{v_{a\omega}}{\mu} = -\frac{1}{1 + \dfrac{R_i}{Z}} v_{g\omega} \tag{6b}$$

if Z is the external anode impedance and R_i the internal anode impedance of the valve.

If the reaction of the external anode impedance on the control voltage of this valve has to be neutralized, $v_{a\omega}/\mu$ must disappear. This will happen when R_i/Z becomes infinitely large; i.e., when either $R_i = \infty$ or $Z = 0$. The factor $1/(1+R_i/Z)$ will then become zero and independent of the external anode impedance Z. By introducing the voltage $\delta v_{a\omega} = -(v_{a\omega}/\mu)$ into the grid circuit, we obtain the same result as if the relation R_i/Z had been made infinitely large, a condition which of course cannot be obtained by direct means.

From the point of view of the external circuit the generator, discussed, thus will work as if the anode impedance were infinitely large or in other words the generator will supply the load with constant current.

As a valve generator connected as above and a direct-current rotary compound generator work in a very similar manner the term "compound generator" will be used for the circuit in question. The voltage $\delta v_{a\omega}$ will correspondingly be termed the "compounding voltage."

In addition, according to (3), the grid bias V_{g0} has been made equal to $-(V_{a0}/\mu)$ which means that an operating point has been chosen for which theoretically no anode current will flow. At the same time the alternating grid voltage range from the operating point to the point

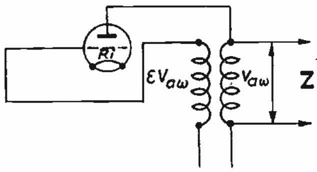

Fig. 3. Feed-back circuit.

where grid current starts to flow is made use of in the most economical way.

To summarize, the advantages of the compound generator are as follows:

1. The generator works with an apparently infinite internal resistance.
2. The external anode impedance (Z) thus has no effect on the anode current (i_a).
3. The static and dynamic characteristics of the valve coincide (Fig. 2).
4. The control voltage (V_c) is always equal to the grid alternating voltage ($v_{g\omega}$).

The condition for compounding is, according to (3), $\delta = -(1/\mu)$. Thus the compounding voltage must be in phase opposition to the anode alternating voltage. This is the same phase relationship as in an ordinary valve oscillator with feedback. It is therefore appropriate to study the difference between "compoundation" and reaction a little more fully. We assume in Fig. 3 a grid alternating voltage of $v_{g\omega}$ and then get the anode alternating voltage.

$$v_{a\omega} = \frac{\mu v_{g\omega}}{R_i + Z} Z. \tag{6a}$$

Of this voltage a certain fraction ϵ may now by reaction be applied to the grid. In order that the valve may start to oscillate, obviously $\epsilon v_{a\omega}$ must be $\geqq v_{g\omega}$. This gives the condition for oscillation; viz.,

$$\epsilon \leqq -\frac{R_i + Z}{Z} \frac{1}{\mu} \cdot \tag{7}$$

If now for instance $R_i = Z$ the condition will be

$$\epsilon \leqq -\frac{2}{\mu}.$$

For compoundation the corresponding condition is

$$\delta = -\frac{1}{\mu}.$$

Thus with $R_i = Z$ or correct matching between the internal anode impedance and the external load impedance the feed-back voltage must be twice the compounding voltage.

If on the other hand the external load impedance Z is very large $(Z \to \infty)$ the condition (7) will become

$$\epsilon \leqq -\frac{1}{\mu}$$

or equal to the condition for compoundation.

III. Reintroduction

We shall now study a compound generator with a circuit diagram according to Fig. 4.

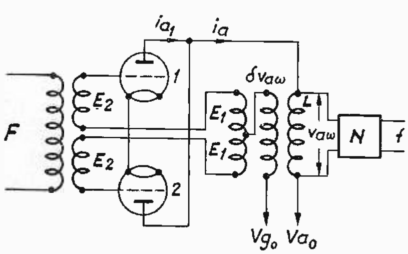

Fig. 4. Circuit diagram of compound generator with reintroduction.

An oscillation having the frequency F and the amplitude E_2 is applied to the grids in opposite phase. The anodes of the valves are connected in parallel and feed an external load Z of a frequency-selective network N, as, e.g., a resonance circuit or a filter. A transformer having the primary winding L is connected in parallel with the network N, L having an inductance either high enough to permit its reactance being neglected in comparison with the input impedance Z of the network, or being included in the network as the first parallel coil. We suppose that the input impedance Z of the network N with the parallel inductance L is real and equal to R ohms for the frequency f, being, however, very nearly equal to zero for low and high frequencies. The three secondary windings of the transformer are designed and connected in such a way that each of the two grids receives the voltage $\delta v_{a\omega}$, where $v_{a\omega}$ is the anode alternating voltage across the network N, as well as the voltages $+E_1$ and $-E_1$ which depend on $v_{a\omega}$.

The characteristic equation of the generator is according to the previous deduction

$$i_a = m_0 + m_2 v_{g\omega}^2. \tag{5}$$

The alternating grid voltage will be, according to Fig. 4,

$$v_{g\omega} = E_2 \sin \Omega t + E_1 \sin (\omega t + \phi). \tag{8}$$

The voltage E_2 fed to the grid has been chosen as zero phase and the voltage E_1 is supposed to have a phase shift ϕ, unknown as yet, in relation to E_2. Equations (5) and (8) give

$$\left.\begin{aligned} i_a = m_0 &+ \frac{m_2}{2} E_1^2(1 - \cos (2\omega t + 2\phi)) \\ &+ \frac{m_2}{2} E_2^2(1 - \cos 2\Omega t) + \\ &+ m_2 E_1 E_2 \cos ((\Omega - \omega)t - \phi) \\ &- m_2 E_1 E_2 \cos ((\Omega + \omega)t + \phi) \end{aligned}\right\}. \tag{9}$$

This anode current i_a produces a voltage drop $v_{a\omega}$ across the anode impedance Z

$$v_{a\omega} = i_a Z. \tag{10}$$

If an oscillation shall be maintained in the anode circuit the voltage $\alpha v_{a\omega}$, which constitutes a certain fraction α of the anode alternating voltage $v_{a\omega}$, must have the same frequency and phase shift and at least as great an amplitude as the oscillation $E_1 \sin (\omega t + \phi)$, which we have assumed to exist across the secondary winding of the anode transformer. The voltage $\alpha v_{a\omega} = \alpha i_a Z$ depends partly on the value of α and partly on the frequency-dependent anode impedance Z. As i_a contains four alternating-current terms having different frequencies, four cases may occur depending on the value of Z for these frequencies. We then suppose that α is a positive number, and introduce positive as well as negative signs before the right member of the equation. These signs are included in the angular functions and determine the value of ϕ.

1. $Z = R$ for $2F$

   $\cong 0$ for other frequencies

$$\mp \alpha R \frac{m_2}{2} E_2^2 \cos 2\Omega t \geqq E_1 \sin(\omega t + \phi) \tag{11}$$

and thus

$$\left.\frac{m_2}{2}\alpha R E_2^2 \geqq E_1 \quad \begin{array}{|l} \sin\left(\dfrac{\pi}{2} \pm 2\Omega t\right) = \sin(\omega t + \phi) \\ \text{or} \\ \sin\left(\dfrac{3\pi}{2} \pm 2\Omega t\right) = \sin(\omega t + \phi) \\ \omega = \underset{(-)}{+} 2\Omega \qquad \phi = \dfrac{\pi}{2} \\ \omega = \underset{(-)}{+} 2\Omega \qquad \phi = \dfrac{3\pi}{2} \end{array}\right\} . \tag{12}$$

Frequency doubling is obtained even if $E_1 = 0$, i.e., $\alpha = 0$, as m_2, R and E_2 are known and >0.

2. $Z = R$ for $2f$

   $\cong 0$ for other frequencies.

$$\mp \alpha R \frac{m_2}{2} E_1^2 \cos(2\omega t + 2\phi) \geqq E_1 \sin(\omega t + \phi) \tag{13}$$

and thus

$$\left.\frac{m_2}{2}\alpha R E_1 \geqq 1 \quad \begin{array}{|l} \sin\left(\dfrac{\pi}{2} \pm (2\omega t + 2\phi)\right) = \sin(\omega t + \phi) \\ \text{or} \\ \sin\left(\dfrac{3\pi}{2} \pm (2\omega t + 2\phi)\right) = \sin(\omega t + \phi) \\ \omega = 0 \qquad \phi = -\dfrac{\pi}{2}, \quad \dfrac{\pi}{6} \\ \omega = 0 \qquad \phi = -\dfrac{3\pi}{2}, \quad \dfrac{\pi}{2} \end{array}\right\} . \tag{14}$$

This solution is rejected as it is not practically realizable.

3. $Z = R$ for $F + f$

   $\cong 0$ for other frequencies

$$\mp \alpha R m_2 E_1 E_2 \cos((\Omega + \omega)t + \phi) \geqq E_1 \sin(\omega t + \phi) \tag{15}$$

and thus

$$m_2 \alpha R E_2 \geqq 1 \quad \left.\begin{array}{|l}
\sin\left\{\dfrac{\pi}{2} \pm ((\Omega + \omega)t + \phi)\right\} = \sin(\omega t + \phi) \\
\text{or} \\
\sin\left\{\dfrac{3\pi}{2} \pm ((\Omega + \omega)t + \phi)\right\} = \sin(\omega t + \phi) \\
\Omega t = -\dfrac{\pi}{2} \\
\omega = -\dfrac{\Omega}{2} \qquad \phi = \dfrac{\pi}{4} \\
\Omega t = -\dfrac{3\pi}{2} \\
\omega = -\dfrac{\Omega}{2} \qquad \phi = \dfrac{3\pi}{4}
\end{array}\right\} \qquad (16)$$

This solution is rejected as it is not practically realizable.

4. $Z = R$ for $F - f$

$\cong 0$ for other frequencies

$$\pm\, \alpha R m_2 E_1 E_2 \cos((\Omega - \omega)t - \phi) \geqq E_1 \sin(\omega t + \phi) \qquad (17)$$

and thus

$$m_2 \alpha R E_2 \geqq 1 \quad \left.\begin{array}{|l}
\sin\left\{\dfrac{\pi}{2} \pm ((\Omega - \omega)t - \phi)\right\} = \sin(\omega t + \phi) \\
\text{or} \\
\sin\left\{\dfrac{3\pi}{2} \pm ((\Omega - \omega)t - \phi)\right\} = \sin(\omega t + \phi) \\
\omega = \dfrac{\Omega}{2} \qquad \phi = \dfrac{\pi}{4} \\
\Omega t = \dfrac{\pi}{2} \\
\omega = \dfrac{\Omega}{2} \qquad \phi = \dfrac{3\pi}{4} \\
\Omega t = \dfrac{3\pi}{2}
\end{array}\right\} \qquad (18)$$

This solution gives *frequency splitting*; i.e., the oscillation in the anode circuit has half the frequency value of the input oscillation ($\omega = \Omega/2$). The phase angle of the oscillation is $\phi = \pi/4$ in relation to the input oscillation which we have chosen as zero phase. The solution giving $\phi = 3\pi/4$ produces the same results if both voltages E_1 fed back to the grid are shifted 180 degrees.

The impedance of the anode circuit ($Z = R$) and the modulator constant m_2 being known, the condition for the generation of this oscillation is at given α

$$E_2 \geqq \frac{1}{\alpha R m_2} \tag{19a}$$

at given amplitude E_2

$$\alpha \geqq \frac{1}{E_2 R m_2}. \tag{19b}$$

IV. Frequency Multiplication

As shown in Part II we obtain *frequency doubling* in the anode circuit of Fig. 4 even if E_1 or $\alpha = 0$. In this case (8) and (9) must be written.

$$v_{g\omega} = E_2 \sin \Omega t \tag{8a}$$

$$i_a = m_0 + \frac{m_2}{2} E_2{}^2(1 - \cos 2\Omega t) \tag{9a}$$

to find the conditions for optimum output power $P_{2\Omega}$ we have

$$\begin{aligned} P_{2\Omega} &= v_{a2\Omega} i_{a2\Omega} = \frac{Z}{2} (i_{a2\Omega})^2 \\ &= \frac{Z}{2} \left(\frac{m_2}{2} E_2{}^2 \right)^2 \end{aligned} \tag{20}$$

as $v_{a2\Omega}$ and $i_{a2\Omega}$ are peak values.

The output power $P_{2\Omega}$ is limited by the condition that the fixed grid bias $V_{g0} = -(V_{a0}/\mu)$ does not allow the oscillation E_2 to reach great amplitudes. Under these conditions grid current would flow through the valves and the fundamental equations (1) and (2) would no longer be valid. We shall therefore calculate the optimum output on that condition. The voltage E_2 and the compound voltage $\delta v_{a2\Omega} = (m_2 E_2^2/2\mu) Z$ are fed to the grids. These voltages have, as shown before, a certain phase shift in relation to each other, and the maximum amplitude of the vector sum of them will be less than the algebraic sum of their amplitudes. However, to make sure of being on the same side, we put

$$E_2 + \frac{m_2 E_2{}^2}{2\mu} Z \leqq | V_{g0} | . \tag{21}$$

Introducing (21) into (20) we obtain

$$P_{2\Omega} = \frac{\mu m_2}{4} E_2{}^2(| V_{g0} | - E_2) . \tag{22}$$

By derivation with respect to E_2 we find the condition for optimum output power to be

$$E_{2\text{opt}} = \frac{2}{3} | V_{g0} | = \frac{2}{3} \frac{V_{a0}}{\mu} \text{ (peak)} \tag{23a}$$

and thus by means of (20), (21), and (22)

$$\left.\begin{aligned} P_{2\Omega\text{opt}} &= \frac{\mu m_2}{27} | V_{g0} |^3 \\ Z_{\text{opt}} &= \frac{3\mu}{2m_2} \frac{1}{| V_{g0} |} \\ v_{a2\Omega\text{opt}} &= \frac{\mu | V_{g0} |}{3} = \frac{V_{a0}}{3} \text{ (peak)} \end{aligned}\right\} \tag{23b}$$

The circuit which gives *frequency doubling* is shown in Fig. 5.

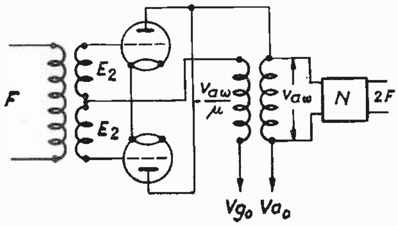

Fig. 5. Simplified circuit diagram of frequency doubler.

This circuit is only one example of frequency multiplication circuits. Suppose that we introduce into the grid circuit oscillations with the two frequencies ω_1 and ω_2. According to the fundamental equation (5) we get alternating-current terms in the anode current with the following frequencies $2\omega_1$, $2\omega_2$, $\omega_1+\omega_2$, and $\omega_1-\omega_2$ and if we insert frequency selective networks or resonant circuits in the anode circuit for these frequencies we might draw power at these frequencies from the generator.

The principle of *reintroduction* is of particular importance in certain types of multiple generators. If we make $\omega_1 = \omega$ and $\omega_2 = 2\omega$ the frequencies in the anode circuit will be ω, 2ω, 3ω, and 4ω. Among these we find the frequency 2ω and it is now of no need any longer to take the primary oscillation $\omega_2 = 2\omega$ from a separate oscillator. Instead we take this oscillation 2ω from the anode circuit and reintroduce it into the grid circuit. The circuit according to this principle is shown in Fig. 6.

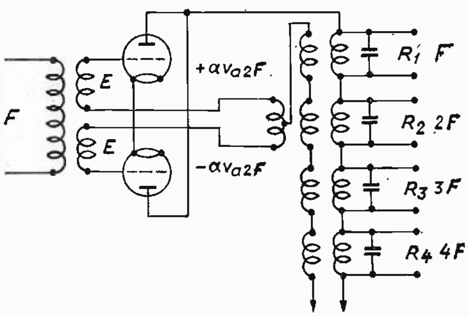

Fig. 6—Circuit diagram of multiple generator.

For this circuit it is also possible to perform a computation in order to ascertain the conditions for equal and optimum power in the anode circuits with the load resistances R_1; R_2, R_3, and R_4. This is done in the same way as previously and we find

$$\left.\begin{array}{l}
E_{\text{opt}} = \dfrac{1}{3}\left| V_{g0} \right| = \dfrac{1}{3}\,\dfrac{V_{a0}}{\mu} \text{ (peak)} \\[2ex]
\alpha = \dfrac{3}{\mu} \\[2ex]
P_1 = P_2 = P_3 = P_4 = \dfrac{\mu m_2}{324}\left| V_{g0} \right|^3 \\[2ex]
\begin{array}{l|l}
\text{for } f \text{ and } 3f & \text{for } 2f \text{ and } 4f \\[1ex]
R_{1\text{opt}} = R_{3\text{opt}} = \dfrac{\mu}{2m_2 \left| V_{g0} \right|} & R_{2\text{opt}} = R_{4\text{opt}} = \dfrac{2\mu}{m_2 \left| V_{g0} \right|} \\[2ex]
v_{a\text{opt}} = \dfrac{\mu}{18}\left| V_{g0} \right| & v_{a\text{opt}} = \dfrac{\mu}{9}\left| V_{g0} \right|
\end{array}
\end{array}\right\}. \quad (24)$$

V. Frequency Division

In Part II we deduced the conditions for the production of an oscillation having half the frequency of the master oscillation.in the anode circuit. We shall now briefly touch on the physical process in this case. The generation of an oscillation with *twice* the master frequency is in no way peculiar since it is a logical result of the modulating properties of the circuit with its characteristic parabolic curve. But to explain the production of an oscillation with *half* the master frequency we have to make the same assumptions as those made for the *de Forest-Meissner* feed-back circuit. As in this case we must assume that an oscillation existed from the beginning in the anode circuit, having been produced

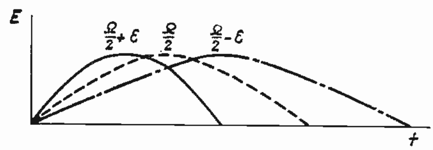

Fig. 7—Curves showing functioning of frequency splitter.

either through the shock when closing the anode circuit, or through shot effect in the valve, through thermal agitation, or in any other way. Let us assume that the frequency of the oscillation is $\Omega e/2 \pm \epsilon$ as shown in the full-drawn line of Fig. 7. The oscillation which is produced as a modulation product of the master oscillation Ω and the oscillation $\Omega/2 \pm \epsilon$ has according to (9) the frequency $\Omega-(\Omega/2 \pm \epsilon)=\Omega/2 \mp \epsilon$, as shown in the dot-dashed line. An oscillation having too high a frequency in the anode circuit thus receives immediately a contribution in oscillation having too low a frequency, the resultant of these being the correct oscillation having the fequency $\Omega/2$ as shown by the dotted curve. This is shown clearly in Fig. 7 or by assuming that both oscillations $\Omega/2 \pm \epsilon$ and $\Omega/2 \mp \epsilon$ are vectors rotating with different speeds and calculating their resultant. The oscillation in the anode circuit obtains thus inevitably a frequency which is exactly $\Omega/2$.

The conditions at which the *frequency splitter* will give optimum anode output, are easily computed in the same way as for the frequency doubler. Taking into account that $i_{a\Omega/2}$ and $v_{a\Omega/2}$ are amplitude values, the anode output is according to (9) and (10)

$$P_{\Omega/2} = i_{a\Omega/2} \cdot v_{a\Omega/2} = \tfrac{1}{2} R(m_2 | E_1 | | E_2 |)^2 \qquad (25)$$

with $Z=R$ for the frequency $\omega=\Omega/2$.

As previously the anode output is limited by the fixed grid bias $V_{go}=-(V_{ao}/\mu)$ and we put

$$E_1 + E_2 + \frac{Rm_2E_1E_2}{\mu} \geqq | V_{g0} | . \tag{26}$$

Introducing (26) into (25) and forming $(\delta P_{\Omega/2})/(\delta E_1)$ respectively $(\delta P_{\Omega/2})/(\delta E_2)$ we find

$$E_{1\text{opt}} = E_{2\text{opt}} = \frac{| V_{g0} |}{3} = \frac{1}{3} \frac{V_{a0}}{\mu} \text{ (peak)}$$

and thus

$$\left. \begin{aligned} P_{\Omega/2\text{opt}} &= \frac{\mu m_2}{54} | V_{g0} |^3 \\ R_{\text{opt}} &= \frac{3\mu}{m_2 | V_{g0} |} \\ \alpha_{\text{opt}} &= \frac{1}{\mu} \\ V_{a\Omega/2\text{opt}} &= \frac{\mu | V_{g0} |}{3} = \frac{V_{a0}}{3} \end{aligned} \right\} . \tag{27}$$

These values do not give abnormal dimensions of the anode impedance R for ordinary valves. As an example we choose a valve having $\mu = 5$ and $m_2 = 3 \cdot 10^{-2}$ ma/v². At $V_{ao} = 240$ volts we get according to the above

$$E_{1\text{opt}} = E_{2\text{opt}} = 16 \text{ volts, peak value}$$

$$\alpha_{\text{opt}} = \frac{1}{\mu} = 0.2$$

$$\delta = -\frac{1}{\mu} = -0.2$$

and may, the external anode impedance being

$$R_{\text{opt}} = 10{,}400 \text{ ohms,}$$

draw an output power from the anode circuit of

$$P_{\Omega/2\text{opt}} = 307 \textit{ milliwatts.}$$

The result of this calculation is remarkable in that the optimum value of the feed-back factor $\alpha = 1/\mu$ has the same numerical value as the compound factor $\delta = -1/\mu$. This signifies that the diagram, Fig. 4, which should have been transformed into the diagram, Fig. 8, may in fact be simplified as shown in Fig. 9. It is thus not necessary to provide the anode transformer with one feedback and one compounding winding, but instead these might be combined into one winding, one end of which is connected to the grid-bias source and the other to the secondary winding of the grid transformer for valve 1. For valve 2, on the other hand, the secondary winding of the trans-

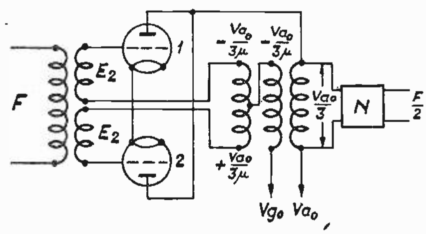

Fig. 8—Circuit diagram of frequency splitter for optimum output.

former is connected direct to the grid-bias source. It must be observed that the circuit, peculiar at first sight, shown in Fig. 9 for a completely compounded *frequency splitter* is valid only for optimum output and on the conditions stated above.

The function of the frequency splitter depends, as shown above, on maintaining in the anode circuit an oscillation having the desired fre-

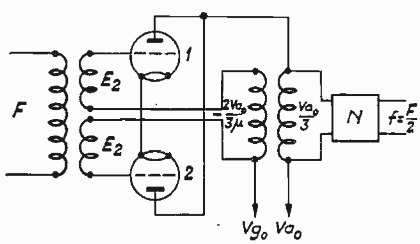

Fig. 9—Simplified circuit diagram of frequency splitter for optimum output.

quency $f = F/2$ by modulation with the master frequency F, in such a manner that $F - f = F/2$. We are thus led to the supposition that a division of the frequency of an input oscillation in three or n parts will be possible.

A mathematical examination shows that this is the case and experiments made with apparatus for the division by three of frequencies have fully confirmed the exactness of the calculations.

The circuit diagram, Fig. 10, demonstrates that compounding windings are introduced in both anode circuits, and that voltages of the frequencies $1/3F$ as well as $2/3F+\delta f$ are fed back to the grid circuit. The functioning of the arrangement is shown in Table I.

TABLE I

Frequencies Fed to the Grid Circuit		Frequencies Produced in the Anode Circuit		
		Double frequencies	Sum frequencies	Difference frequencies
F	master frequency	$(2F)$	$(4/3F)$	$2/3F$
$1/3F$	fed back from the	$2/3F$	$(5/3F+\delta f)$	$1/3F-\delta f$
$2/3F+\delta f$	anode circuit	$(4/3F+2\delta f)$	$(F+\delta f)$	$1/3F+\delta f$

Assuming that both frequency-selective networks in the anode circuit have very small impedances for all other frequencies than $1/3F$ and $2/3F$, respectively, we may neglect the frequencies F, $4/3F$, $5/3F$,

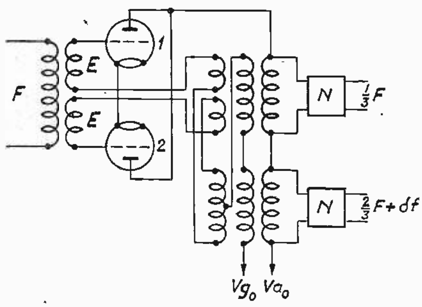

Fig. 10—Circuit diagram of submultiple generator for division by three of the frequency

and $2F$ (enclosed in brackets in the table). We then find that the oscillation having the frequency $1/3F$ may be maintained in two manners, either as $1/3F-\delta f$ by intermodulation between F and $2/3F+\delta f$, or as $1/3F+\delta f$ by intermodulation between $2/3F+\delta f$ and $1/3F$. The somewhat too low and somewhat too high part frequencies have the same amplitude and are combined into a single oscillation having the correct frequency $1/3F$, in the same way as indicated above for the frequency splitter.

VI. Experimental Results

The frequency multiplication circuits have already been in practical use for carrier current installations in different parts of the world. Thus the Telefonaktiebolaget L. M. Ericsson in Sweden has delivered to Mexico a number of four-channel carrier telephone terminals which are fed with carrier currents from multiple generators producing, e.g., the frequencies 10, 20, 30, and 40, or 15, 25, 35, and 45 kilocycles from an input master frequency of 5 kilocycles. These multiple generators have been in operation since 1929 with excellent results.

To verify the practical applications of the frequency splitter, the function of which has caused considerable doubt among expert and patent officials an experiment according to Fig. 11 was set up. The circuit diagram is self-explanatory and it is only appropriate to add that the inductance L consisted of an air-core coil, wound with

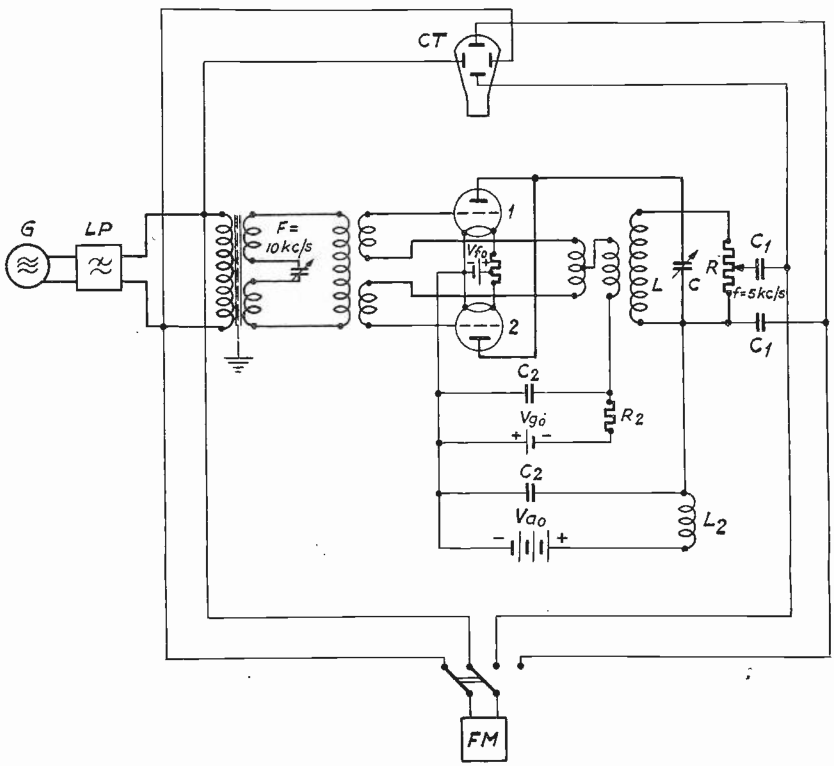

Fig. 11

G = Generator 10 kilocycles
LP = Low-pass filter
CT = Cathode-ray tube
FM = Frequency meter
1-2 = Valves, Marconi-Osram LS 5, $\mu = 5$
$V_{f0} = 6$ volts; $V_{g0} = -48$ volts, $V_{a0} = 240$ volts
$L = 6.63$ millihenrys $C = 0.151$ microfarad
$R = 100{,}000$ ohms $C_1 = 0.06$ microfarad
$R_2 = 70{,}000$ ohms $C_2 = 2$ microfarads $L_2 = 8.6$ henrys

stranded wire and completely shielded in a copper box. The capacitance C was formed by a variable mica condenser in parallel to an air condenser for vernier adjustment.

During the experiments the input and output frequencies F and f respectively were measured by means of the frequency meter. It was found that f was about equal to $F/2$ and that the oscillation in the anode circuit disappeared as soon as the input oscillation was disconnected. In order to investigate the exact relation between F and f the oscillations were fed to a cathode-ray tube using one pair of electrodes for each frequency. The curve on the screen of the tube showed that both oscillations were present and in exact synchronism. Then the two oscillations F and f were alternatively disconnected from the cathode-ray tube electrodes, a time-controlled voltage replacing them. This test showed that $F=2f$ and that the two oscillations were practically sinusoidal.

By changing the capacitance C within $\pm$ 3 per cent of its initial value no effect on the frequency splitting properties of the device was observed. For larger changes the device ceased to work because of the very large decrease in the effective resistance of the resonant circuit when the tuning was off resonance (equations (19a) and (19b)). A much better result was obtained if the resonance circuit (LCR) in the anode circuit was replaced by a band-pass filter. As long as the characteristic impedance of the filter within the pass band is nearly equal to the desired value R_{opt} (equation (27)) the input frequency F can be varied within wide limits. If a good filter circuit with sharp cutoffs is used the band width may be taken quite large, e.g., $0.2F$ making it possible to vary F between $0.8F$ and $1.2F$. Thus, e.g., any oscillation from a variable generator with frequencies between 800 and 1200 cycles might be split into oscillations with frequencies between 400 and 600 cycles without changing any other tuning than that of the input oscillator. This property is very useful for laboratory measurements of all kinds.

VII. Experimental Determination of the Constants μ, m_0, and m_2

The theoretical discussion has shown that both the power and the anode alternating voltage of the generator will depend on the constants μ, m_0, and m_2 which are characteristics of the valves. For the practical design of the multiple generator it is therefore necessary to have reliable information regarding the magnitude of these constants for the different types of valves.

Determination of μ.

The amplification factor is a characteristic quantity which is practically independent of the anode, grid, and filament voltages. Several methods for measuring this amplifying factor are now known. It may be ascertained by determining a number of static valve characteristics for various anode voltages, or else by direct measurements in alternating-current bridges.

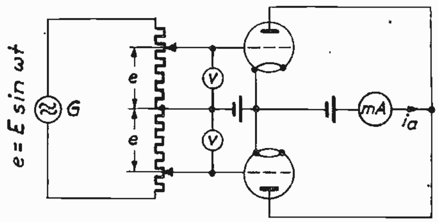

Fig. 12—Circuit for measuring constants of generator.

Generally, μ varies within fairly wide limits in different valves of a given type. Variations of up to ten per cent are not uncommon. When determining μ for a certain type of valve, the mean value of a great number of valves should therefore be taken. If necessary, valves for the multiple generator could be selected in pairs of approximately the same amplification factor value.

Determination of m_0 and m_2

The above discussion of the multiple generator theory has furnished the relation of grid voltage to anode current.

$$i_a = m_0 + m_2 V_g \omega^2. \tag{5}$$

This equation also provides us with a simple means of determining the constants m_0 and m_2. The circuit diagram for measuring these is shown in Fig. 12.

G is an alternating-current generator giving a sinusoidal voltage of an arbitrarily selected frequency. It is best to choose a low frequency, 25 or 50 cycles, as there will then be no difficulty in measuring the voltages E, which may be done with an ordinary electrodynamic voltmeter. Voltages, $e = E \sin \omega t$, are impressed on each of the grids. The anode current in the anode circuit common to both valves is measured by the milliammeter mA.

The measuring is now done by first observing i_a at a certain grid bias V_{go} when $E = O$. We then have

$$m_0 = i_a. \tag{28}$$

The two identical voltages E are then varied by means of the double potentiometer, and the corresponding anode direct current is observed. The value of the constant m_2 may be either computed by the formula

$$m_2 = \frac{i_a - m_0}{E^2} \tag{29}$$

or else obtained as the tangent of the slope of a curve the abscissa of which is E^2 and the ordinate i_a. The former method gives a mean value for m_2 corresponding to the amplitude E impressed at the time, while the latter gives the theoretical value for m_2.

If m_2 be plotted as a function of the amplitude E^2, a straight line parallel to the abscissa axis should be obtained. Sometimes, however, this function becomes a curved line, as in Fig. 13.

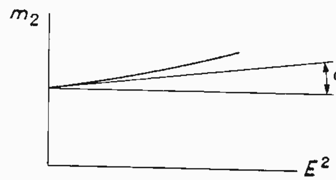

Fig. 13—m_2 as function of E^2.

The curvature is caused by the functional relationship of i_a and $v_{g\omega}$ being in reality expressed by the following:

$$i_a = m_0 + m_2 v_{g\omega}^2 + m_4 v_{g\omega}^4 + m_6 v_{g\omega}^6 \cdots . \tag{5}$$

Hence the curvature of this curve is also a measure of the size of the coefficients m_4, m_6, and so on.

By determining m_2 and m_4 for several values of the grid bias V_{go}, that value of V_{go} may be graphically determined which distorts the parabola the least, or in other words for which m_4 is a minimum. This value may then be used as grid bias for the generator. However, the desire to secure a minimum of distortion is frequently not the determining factor in the choice of the grid bias V_{go}. This is often selected as large as possible to give a greater available power in the anode circuit of the multiple generator, as we know that the power will increase with the cube of V_{go}.

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CHARACTERISTICS OF THE IONOSPHERE AT WASHINGTON, D. C., JANUARY TO MAY, 1937[1]

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JANUARY

THE critical frequency and virtual height data for January, 1937, are shown in Fig. 1. Of these data all of the virtual heights, and the critical frequencies between 2500 and 7800 kilocycles are based on automatic ionosphere records made twenty-hour hours per day every day of the month. The remaining critical frequencies are based on manual measurements made continuously each Wednesday. When the F layer is not stratified the symbols for F layer and F_2 layer will frequently be used interchangeably.

The ionosphere during January, as in other winter months, was marked by great regularity of behavior and high values of daytime f_{F_2}. The graphs require very little discussion, as the average values plotted represent the values for any individual day within 10 per cent except on a few days at hours represented by the very steep portions of the $f_{F_2}{}^x$ graph near sunrise and after sunset. Both f_E and $f_{F_2}{}^x$ were greater than in January, 1936, thus indicating the continuation of the effect of the advancing sunspot cycle.

January was a quiet month ionospherically. The two most disturbed days magnetically were January 10 and 29 but the ionosphere, which is normally very stable during the winter, was not disturbed on these days.

The fade-outs[2] observed at Washington are shown in Table I.

TABLE I

Date	Beginning of fade-out	Beginning of recovery	Recovery complete	Location of transmitter	Intensity
Jan. 19	1901	1907	1921 G.M.T.	Ohio	(0.1)
Jan. 27	1907	1920	1950 G.M.T.	Mass. and D. C.	(0.0)

* Decimal classification: R113.61. Original manuscript received by the Institute, June 9, 1937. Publication approved by the Director of the National Bureau of Standards of the U. S. Department of Commerce.

[1] For definition of symbols and other explanations see T. R. Gilliland, S. S. Kirby, N. Smith, and S. E. Reymer, "Characteristics of the ionosphere and their application to radio transmission," *Nat. Bur. Stand. Jour. Res.*, vol. 18, pp. 546–668; June, (1937), and PROC. I.R.E., vol. 25, pp. 823–840; July, (1937).

Emissions made from 0630 to 0200 E.S.T. by station W8XAL, Mason, Ohio, at 6060 kilocycles over a distance of 650 kilometers were propagated daily by the F layer[3] on the average from 0717 to 0819 E.S.T., principally by the E layer from 0819 to 1614 E.S.T., and by the F layer after 1614 E.S.T. The change of layer occurred when the normal incidence f_E halfway between the transmitter and receiver was about 2450 kilocycles. This transmitting station did not operate regularly from January 22 to February 5 on account of flood conditions in the Ohio River valley.

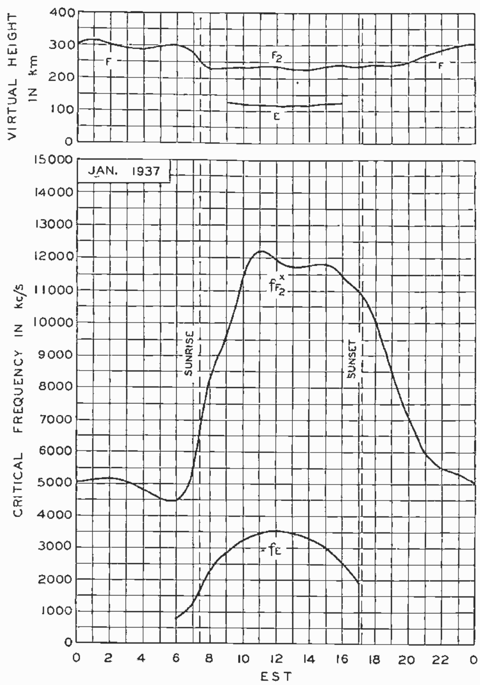

Fig. 1—Average virtual heights and critical frequencies of the E, F, and F_2 layers of the ionosphere at Washington for January, 1937.

Emissions made from 0600 to 0100 E.S.T. by station W1XK, Millis, Mass., at 9570 kilocycles over a distance of 600 kilometers were propagated by the F_2 layer from about 0740 to 1920 E.S.T. These transmissions were received when the normal incidence $f_{F_2}{}^x$ was about

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[2] The fade-out notation is that used by J. H. Dellinger in a paper "Sudden disturbances of the ionosphere," *Nat. Bur. Stand. Jour. Res.*, vol. 19, pp. 111–141; August, (1937). The intensities are fractions of normal.

[3] *Phys. Rev.*, vol. 51, p. 890; May 15, (1937).

8000 kilocycles or greater. The beginnings and endings of reception, caused by changing critical frequencies, were abrupt.

February

Fig. 2 shows the critical frequency and virtual height data for February, 1937. The data were taken in the same manner as during January.

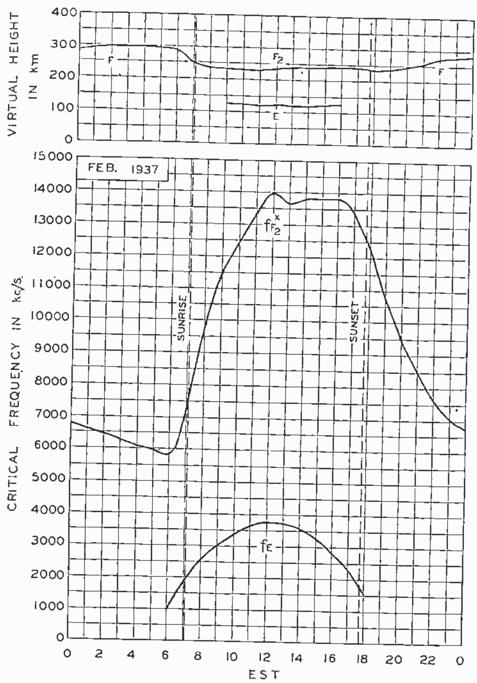

Fig. 2—Average virtual heights and critical frequencies of the E, F, and F_2 layers of the ionosphere at Washington for February, 1937.

The ionosphere during February continued to be marked by great regularity of behavior and high values of daytime f_{F_2}. The night values of f_F as well as the day values of f_{F_2} were greater than during January. Again the graphs of average values represent the values for any individual day within 10 per cent except on February 3 and 19, and on a few other days at hours represented by the very steep portions of the $f_{F_2}{}^x$ graph near sunrise and after sunset. Both f_E and $f_{F_2}{}^x$ were greater than in February, 1936, thus indicating a continuation of the effect of the advancing sunspot cycle. The average values of $f_{F_2}{}^x$ during February, 1937, were greater than for any other month for which such data are available.

February was, for the most part, a quiet month magnetically. The two most disturbed days were February 3 and 19. On February 3 a magnetic storm occurred during which h_F rose decidedly during the early morning but f_{F_2} was not appreciably affected. The latter effect is unusual, as magnetic storms of this intensity are ordinarily accompanied by low f_{F_2}.[4,5,6] This storm appeared to be of a different nature from previous storms studied in this work. On February 19 a slight disturbance occurred which lowered $f_{F_2}{}^x$ somewhat.

The fade-outs observed at Washington are shown in Table II.

TABLE II

Date	Beginning of fade-out	Beginning of recovery	Recovery complete	Location of transmitter	Intensity
Feb. 1	1920	—	1942	D. C.	—
Feb. 17	1555	1606	1640	Ohio, Mass., D. C.	(0.0)

The emissions from W8XAL at 6060 kilocycles described in the January report were propagated principally by the E layer from about 0805 to 1709 E.S.T., and by the F layer before 0805 and after 1709 E.S.T. These transmissions were propagated by the F layer when the transmitting station began emissions at about 0630 E.S.T.

The emissions from W1XK at 9570 kilocycles were propagated by the F_2 layer from about 0721 to 2108 E.S.T. The beginning of F_2 layer propagation on the magnetically disturbed day of February 3, was thirty-six minutes later than the average. This lag of reception was caused mainly by the abnormally great virtual heights rather than a slow rise of $f_{F_2}{}^x$. The beginning of F_2 layer propagation on February 19 was three hours and twenty-four minutes later than the average and the time of failure was three hours and twelve minutes earlier than the average. These abnormalities were caused chiefly by low $f_{F_2}{}^x$ and partly by high h_{F_2}.

March

Fig. 3 shows the critical frequency and virtual height data for March, 1937. The data were obtained in the same manner as during preceding months but the F_1 and F_2 layer data for magnetically quiet and for disturbed days are plotted separately. The solid-line graphs represent the magnetically quiet days and the dotted-line graphs represent the magnetically disturbed days. No disturbances of the E layer have been observed during magnetic storms.

The behavior of the ionosphere during March was fairly regular ex-

[4] *Phys. Rev.*, vol. 48, p. 849; (1935).
[5] *Phys. Rev.*, vol. 50, p. 258; Aug. 1, (1936).
[6] *Phys. Rev.*, vol. 51, p. 992; June 1, (1937).

cept that the F_2 layer seemed to be more subject to disturbances during magnetic storms as the season advanced toward spring.

The average values plotted represent the values for any individual magnetically quiet day within 10 per cent except on a few days for the hours represented by the steep portions of the graphs. The maximum values of f_E were about the same as in February instead of increasing, as might have been expected, with the usual seasonal change. This was

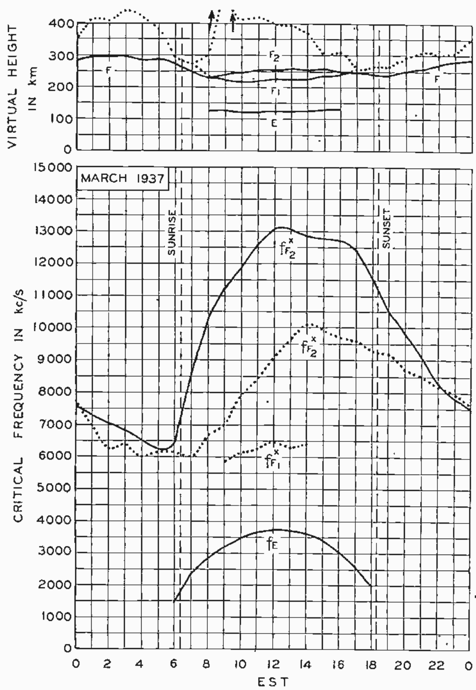

Fig. 3—Virtual heights and critical frequencies of the E, F_1, F_2, and F layers of the ionosphere for March, 1937. The solid-line graphs represent averages for the magnetically quiet days and the dotted-line graphs are for the magnetically disturbed day of March 31.

associated with a sharp decrease in sunspot numbers from February to March. The effects of the longer days were indicated by the broader graph. A slight separation of h_{F_1} and h_{F_2} was observed on magnetically quiet days but f_{F_1} was not well defined except on magnetically disturbed days. The daytime f_{F_2} was slightly lower and the night f_F was higher on quiet days than in February. The magnetically disturbed days in the order of the severity of the ionosphere disturbances were March 15, 31, 5, 27, and 28. There was a slight disturbance of the ionosphere on March 16 following the magnetic storm of March 15. March 15, 31, and 5 were marked by daytime stratification of the F layer, sharp f_{F_1}, high h_{F_2}, and low f_{F_2}. The fade-outs observed at Washington are shown in Table III.

TABLE III

Date	Beginning of fade-out	Beginning of recovery	Recovery complete	Location of transmitter	Intensity
Mar. 1	1908	—	1915 G.M.T.	Ohio	(0.1)

The emissions from W8XAL at 6060 kilocycles described in the January report were propagated principally by the E layer from about 0726 to 1739 E.S.T. These transmissions were propagated by the F layer only, before 0726 E.S.T. and after 1739 E.S.T. The E layer transmissions were unaffected by magnetic storm disturbances of the ionosphere.

The emissions from W1XK at 9570 kilocycles were propagated by the F_2 layer from about 0633 to about 2148 E.S.T. on magnetically quiet days. These times of beginning and ending of propagation, which were abrupt, were regular during the magnetically quiet days. For a given intensity of magnetic disturbance these transmissions were more highly disturbed than during the preceding winter months. See report for April, 1937. This is an indication of the decreased stability of the F_2 layer during the approaching spring transition period. The emissions from W1XK at 9570 kilocycles were not received on March 15 and were received only during the hours shown in Table IV on the other disturbed days.

TABLE IV

Date	Beginning	Ending	Remarks
	E.S.T.	E.S.T.	
March 31	1248	2020	Poorly defined and irregular
March 5	0800	1844	Intermittent
March 27	1015	2018	Regular
March 28	0800	2303	Regular
March 16	0727	2200	Regular

These irregularities were caused by the high h_{F_2} and low f_{F_2} during these disturbed periods.

April

Fig. 4 shows the critical frequency and virtual height data for April, 1937. The data were plotted in the same manner as for March.

The ionosphere during April was marked by increased irregularity of behavior of the upper layers, by a further separation of the F_1 and F_2 layers, by an increase of h_{F_2} and a large decrease of f_{F_2}. The irregular behavior of the upper layers, especially of the F_2 layer, seemed to be caused in part by an innate instability, but principally by severe disturbances of these regions associated with magnetic storms. The magnetic storms were both numerous and severe, especially during the latter part of the month. All of these effects may be regarded as seasonal to a large extent. The behavior of the E layer was regular. The values of f_E were greater than in March but the midday values were slightly less than in April, 1936.

A series of severe magnetic storms began on April 24 and while they were not continuous the remainder of the month was very much dis-

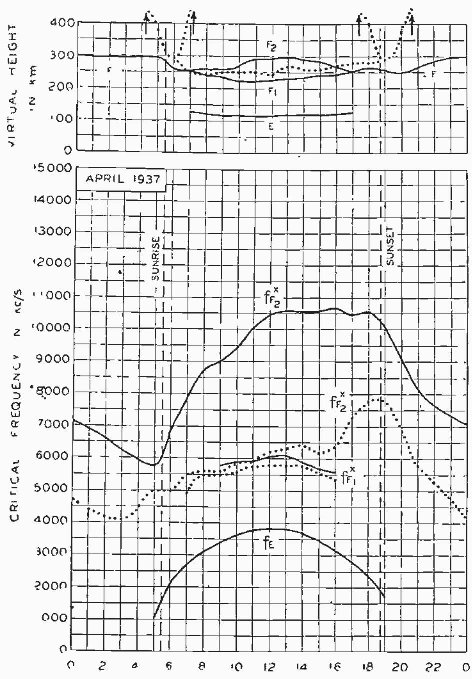

Fig. 4—Virtual heights and critical frequencies of the E, F_1, F_2, and F layers of the ionosphere for April, 1937. The solid-line graphs represent averages for the magnetically quiet days and the dotted-line graphs are for the magnetically disturbed day of April 27.

turbed. The principal effects in the ionosphere were abnormal increases of the virtual heights and abnormal decreases of the critical frequencies and increased absorption of reflections from the upper layers of the ionosphere. All of these effects made for poor F layer transmission. No effects of these disturbances were observed for the E layer or E layer transmission.

Table V gives data for magnetically disturbed days, the days being listed in the order of the intensity of the ionosphere disturbance.

TABLE V

Date	h_F before sunrise	Max. fF_2^x during day (near sunset)	Average mag. character[7]	
			0000–1200 G.M.T.	1200–2400 G.M.T.
Apr. 28	390 km	7000 kc	1.9	1.9
Apr. 27	446 km	8000 kc	1.5	1.4
Apr. 30	346 km	7600 kc	0.6	0.7
Apr. 25	452 km	7700 kc	1.3	1.9
Apr. 26	400 km	above 8400 kc	1.3	1.8
Apr. 24	no data	9100 kc	0.2	1.6
Apr. 3	352 km	9000 kc	0.9	0.8
Apr. 29	342 km	near normal	0.6	0.7
Undisturbed average	296 km	10,670 kc	0.0	0.0

A list of the fade-outs observed during April is shown in Table VI.

TABLE VI

Date	Beginning of fade-out	Beginning of recovery	Recovery complete	Location of transmitter	Intensity
			G.M.T.		
Apr. 1	1727	—	1743	Ohio	(0.2)
Apr. 21	1359	1404	1413	Ohio	(0.5)
	1453	1520	1535	Ohio, Mass., D. C.	(0.0)
	1808	1814	1821	Ohio	(0.0)
	1842	1852	1907	Ohio	(0.0)
	2006	2037	2116	Ohio, Mass., D. C.	(0.0)
	2132	—	2155	Ohio, D. C.	(0.0)
Apr. 22	1710	1721	1726	Ohio, D. C.	(0.05)
	1808	1816	1829	Ohio, D. C.	(0.02)
	1848	—	1943	Ohio, Mass., D.C.	(0.0)
	1948	2002	2033	Ohio	(0.0)
Apr. 24	1341	—	1355	Ohio	(0.2)
	1958	2012	2025	Ohio, Mass.	(0.0)
Apr. 25	1352	1403	1415	Ohio	(0.0)
	1547	1602	1614	Ohio	(0.0)
	1645	1729	1800	Ohio, D. C.	(0.0)
	2121	2133	2146	Ohio, D. C.	(0.0)
Apr. 26	1603	1636	1720	Ohio, Mass., D. C.	(0.0)
	2248	2313	2323	Ohio, D. C.	(0.0)
Apr. 27	1544	—	1624	Ohio	(0.0)
	1642	1708	1720	Ohio, D. C.	(0.0)
	1830	—	1901	Ohio, D. C.	(0.1)
	2102	—	2130	Ohio, D. C.	(0.1)
Apr. 28	1124	—	1130	Ohio	(0.1)
	1132	—	1143	Ohio	(0.1)
	1205	1217	1222	Ohio, D. C.	(0.0)
	1342	—	1406	Ohio	(0.1)
Apr. 29	2107	2115	2132	Ohio, D. C.	(0.0)
Apr. 30	1805	—	1810	Ohio	(0.5)
	1935	—	2119	Ohio	(0.0)

The emissions from W8XAL at 6060 kilocycles were propagated by the E layer from about 0637 to 1759 E.S.T. These emissions were propagated by the F layer before 0637 E.S.T. except on the magnetically disturbed days of April 26, 27, 28, and 30, when $f_F{}^x$ was too low. These emissions were also propagated by the F layer after

[7] These magnetic character figures were compiled by the Department of Terrestrial Magnetism, Carnegie Institution of Washington, from data supplied by their own observatories and by the observatories of the U. S. Coast and Geodetic Survey.

1759 E.S.T. except that the transmissions were weak and erratic on the nights of April 24, 25, 26, and 27. The absorption of the midday transmissions was much greater than during the winter.

The emissions from W1XK at 9570 kilocycles were propagated by the F_2 layer from about 0704 to 2058 E.S.T. except on the disturbed days of April 2, 3, 15, 24, 25, 26, 27, 28, and 30. The times of beginning and ending of these transmissions, even on magnetically quiet days, was not as regular as during the winter. The transmissions failed all day on April 25, 27, 28, and 30. Table VII gives the times of beginning and ending of transmission on other disturbed days, the days being listed in the order of the severity of the ionosphere disturbance.

TABLE VII

Date	Transmission begins	Transmission ends
April 3	1630 E.S.T.	1945 E.S.T.
April 26	1500 E.S.T.	1700 E.S.T.
April 24	1626 E.S.T.	uncertain
April 2	1510 E.S.T.	1854 E.S.T.
April 15	1515 E.S.T.	2021 E.S.T.
April 14	1003 E.S.T.	2010 E.S.T.

May

Fig. 5 shows the critical frequency and virtual height data for May, 1937. The data were plotted in the same manner as for March and April.

The ionosphere during May was marked by a continued increase of irregularity in the behavior of the upper layers, by a further separation of the F_1 and F_2 layers, by a further increase of the daytime h_{F_2} and a further decrease of the daytime f_{F_2}. As in April, the irregular behavior of the upper layers, especially of the F_2 layer, seemed to be caused in part by an innate instability and in part by disturbances of these regions associated with magnetic storms. All of these effects may be regarded as seasonal to a large extent. The behavior of the normal E layer was regular. The values of f_E were greater than in April but about the same as in May, 1936. Strong sporadic E reflections at normal incidence were observed up to 11 or 12 megacycles on the evening of May 14. The sporadic E layer appeared to be responsible for 56-megacycle transmission over 700 or 800 miles on this evening.[8]

May was less disturbed magnetically than April. On May 5, however, there occurred a magnetic storm which was accompanied by $f_{F_2}{}^x$ lower than $f_{F_1}{}^x$ during about eight hours of the day. This effect, which has been observed to be associated with severe magnetic storms, was

[8] The 56-megacycle transmissions were reported to the Bureau by Elmer H. Conklin, Assistant Editor of *Radio*.

also observed during the severe storms in the latter part of April, 1937. The most disturbed days are listed in Table VIII in the order of the intensity of the ionosphere disturbances.

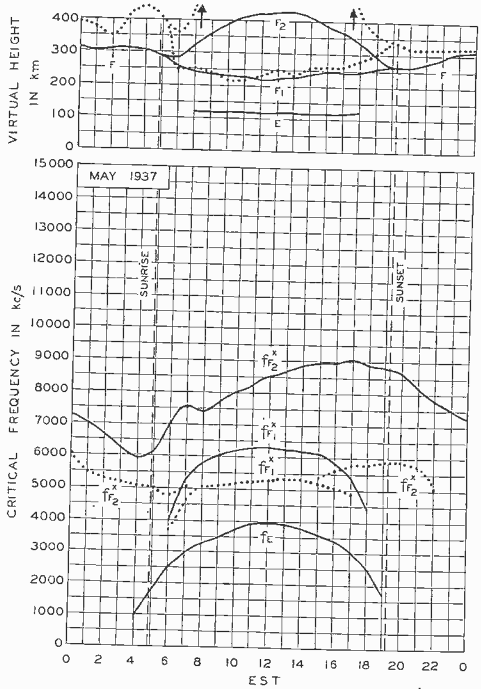

Fig. 5—Virtual heights and critical frequencies of the E, F_1, F_2, and F layers of the ionosphere for May, 1937. The solid-line graphs represent averages for the magnetically quiet days and the dotted-line graphs are for the magnetically disturbed day of May 5. Note F_2 layer reflections obscured by F_1 layer from about 0700 to 1500 E.S.T. on May 5.

TABLE VIII

Date	h_F before sunrise	Max. $f_{F_2}^x$ during day (near sunset)	Magnetic character	
			0000–1200 G.M.T.	1200–2400 G.M.T.
5	402 km	5900 kc	1.8	1.2
29	372 km	6800 kc	no data	no data
1	349 km	6800 kc	0.8	0.4
28	317 km	8100 kc	1.0	0.9
Average of undisturbed days	307 km	9080 kc	0.0	0.0

The f_{F_1}, which was normally well defined in May, was observed to be decreased during magnetic storms. On May 5 this decrease amounted to about 800 kilocycles below normal.

Fade-outs observed during May are shown in Table IX.

TABLE IX

Date	Beginning of fade-out	Beginning of recovery	Recovery complete	Location of transmitter	Intensity
			G.M.T.		
May 1	1902	1915	1940	Ohio, D. C.	(0.0)
	2022	2040	2110	Ohio, Mass., D. C.	(0.0)
May 3	1600		1620	Ohio	(0.05)
	1640		1730	Ohio	(0.02)
May 5	1434		1450	Ohio	(0.1)
	1512		1521	Ohio	(0.05)
	1643	1651	1705	Ohio, D. C.	(0.01)
May 17	1855	1910	1925	Ohio, Mass.	(0.01)
May 19	1601	1622	1631	Ohio, Mass., D.C.	(0.0)
	1640		1707	Ohio, Mass.	(0.05)
May 23	2140		2150	Ohio, D. C.	(0.05)
May 25	1313	1415	1505	Ohio	(0.0)
	1520	1528	1534	Ohio	(0.0)
	1649	1716	1727	Ohio	(0.0)
May 27	2052	2129	2200	Ohio, D. C.	(0.01)

Emissions from station W8XAL, 6060 kilocycles, were propagated regularly by the F layer at night except May 28 and on every morning except May 1, 4, 5, 14, and 29. E layer transmission began on the average at 0604 and continued until 1818 E.S.T. The high midday absorption continued and was especially marked on May 3, 20, and 22.

Emissions from station W1XK, 9570 kilocycles, were propagated very erratically during May, mostly by sporadic E at irregular times and occasionally by F_2 layer in the late afternoon. There was no F_2 transmission on May 1, 5, 8, 11, 12, 15, 19, 22, 23, 24, 25, and 30. There was no forenoon transmission on May 2, 3, 4, 6, 9, 16, 18, 20, 26, 27, 28, 29, and 31.

Transmission from DJB, Germany, 15,200 kilocycles, distance 6700 kilometers, operating hours 1700 to 2300 E.S.T., were recorded beginning May 7. Transmissions were good on the nights of May 7, 8, 9, 11, 12, 14, 15, 16, 17, 18, 19, 20, 23, and 26, fair on the nights of May 10, 13, 21, 22, 24, 25, 29, 30, 31, and poor on the evenings of May 27 and 28.

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CHARACTERISTICS OF THE IONOSPHERE AT WASHINGTON, D. C., JUNE, 1937*

By

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S. E. Reymer
(National Bureau of Standards, Washington, D. C.)

The critical frequency and virtual height data for June, 1937, are shown in Fig. 1.

The ionosphere during June behaved in a typically summer fashion. The F_1 and F_2 layers were well separated during the day, f_{F_1} was sharp and well defined, h_{F_2} great and f_{F_2} low compared with winter values. Night values of f_F were higher than during the winter. The behavior of the F_2 layer was more regular than during April and May.

The behavior of the normal E layer was fairly regular. Variations of f_E occurred, amounting to as much as 200 kilocycles from day to day. The values of f_E were considerably greater than those of May, 1937, and approximately 300 kilocycles greater than for June, 1936. The values of f_E for June, 1937, were the highest ever recorded at Washington. Sporadic E reflections were frequently observed at frequencies as high as 6200 kilocycles. Most daytime radio transmissions over moderate and long distances were propagated by way of the E layer.

June was much less disturbed ionospherically and magnetically than April and May. The greatest ionospheric disturbance of the month occurred during and following the moderate magnetic disturbance of June 5 and 6. The ionospheric disturbance began about 0200 E.S.T. June 6 and ended about sunrise June 7. The data are plotted in Fig. 1. The $f_{F_2}{}^x$ was lower than $f_{F_1}{}^x$ for about nine hours during the day of June 6 and therefore could not be observed during this period. Less disturbed periods named in the order of their severity were June 28 and the early morning hours of June 25 and 1.

The critical frequencies of both the F_1 and F_2 layers decreased during the disturbance and the virtual heights increased.

Out of 232 hours of night measurements of $f_F{}^x$ only one value was more than 20 per cent above and four values more than 15 per cent above the undisturbed average. Two values were over 35 per cent below, 28 values over 20 per cent below and 37 values over 15 per cent

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below the undisturbed average. Of the 28 hours when the $f_F{}^x$ was more than 20 per cent below the undisturbed average all occurred on the magnetically disturbed nights of June 1, 6, 7, 25, and 28. Of the 37 values more than 15 per cent below the undisturbed average all but seven occurred on these same nights.

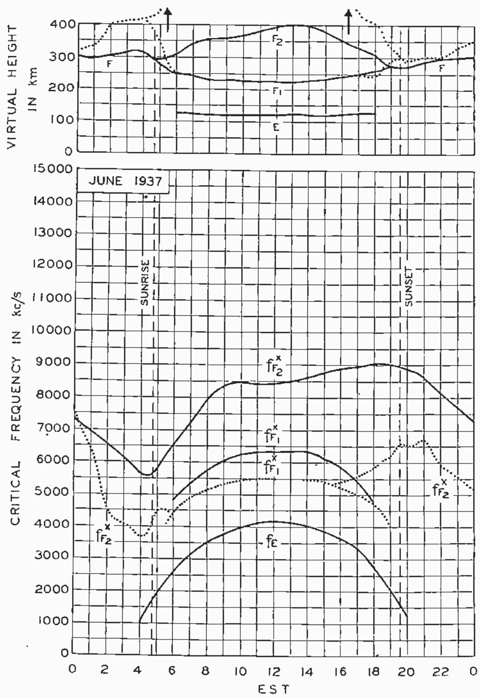

Fig. 1—Virtual heights and critical frequencies of the E, F_1, F_2 and F, layers of the ionosphere for June, 1937. The solid-line graphs represent averages for the magnetically quiet days and the dotted-line graphs are for the magnetically disturbed day of June 6. On this day the F_2 layer reflections were obscured by the F_1 layer from about 0600 to 1500 E.S.T.

Manual daytime measurements were made from about 0400 to 2300 E.S.T. on June 2, 8, 16, 23, and 30. Of the daytime measurements of $f_{F_2}{}^x$ 5 hourly values were more than 10 per cent above or below the undisturbed average values. These values were less than 15 per cent from the average values.

Emissions from station W8XAL, 6060 kilocycles, 650 kilometers, were propagated regularly by the F layer at night. E layer transmission began on the average at 0554 E.S.T. and continued until 1849 E.S.T.

The fade-outs observed during June are shown in Table I.

TABLE I

Date	Beginning of fadeout	Beginning of recovery	Recovery complete	Location of transmitter	Intensity
June 2	1938	1954	2038	Ohio, Mass.	0.00
June 4	2126	—	2224	Ohio	0.4
June 6	1738	—	1804	Ohio, Mass.	0.05
June 9	1420	1428	1504	Ohio, Mass.	0.05
June 10	1852	—	1930	Ohio, Mass., Cuba, D. C.	0.2
June 14	1532	1554	1608	Ohio, Mass., Germany	0.05
	1632	1704	1730	Ohio, Mass., Cuba	0.00
June 17	2028	2035	2048	Ohio	0.05
	2058	2104	2114	Ohio	0.05
June 20	1517	1528	1540	Ohio, Mass.	0.00
	1806	1815	1830	Ohio, Mass.	0.01
	2006	—	2026	Ohio	0.01
	2109	2117	2130	Ohio	0.00
June 24	1326	1330	1354	Ohio, Mass.	0.0
	1751	—	1810	Ohio, Mass.	0.05
	1836	1848	1858	Ohio, Mass.	0.0
June 25	1323	—	1330	Ohio	0.1

Emissions from station W1XK, 9570 kilocycles, 600 kilometers, were propagated regularly by normal E layer for several hours during the middle of the day, by the F layer frequently for a few hours around sunset, and by sporadic E frequently at irregular intervals.

Emissions from station DJB, 15200 kilocycles, 6700 kilometers, were received very poorly on the evenings of June 5, 24, 25, 27, 29; fair on the evenings of May 31, June 15, 20, 22, 28, and very well on the other evenings.

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CHARACTERISTICS OF THE IONOSPHERE AT WASHINGTON, D. C., JULY 1937*

By

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(National Bureau of Standards, Washington, D. C.)

THE critical frequency and virtual height data for July, 1937, are shown in Fig. 1. The characteristics of the ionosphere were similar to those which have already been described for June, 1937. The characteristics were also similar to those of July, 1936, with the exception that all the critical frequencies were higher in July, 1937. The critical frequencies for July, 1937, exceeded those of July, 1936, by approximately the following amounts: f_E −250 kilocycles, f_{F_1} −300 kilocycles, daytime f_{F_2}−1300 kilocycles, nighttime f_F−800 kilocycles. Strong sporadic E reflections were present up to 4400 kilocycles during 17 per cent of the hours of observation. When this condition existed sporadic E layer rather than the F layer controlled long-distance transmissions. Strong sporadic E reflections were present up to 6200 kilocycles 7 per cent of the hours of observation and were occasionally observed up to 11,000 kilocycles. Strong sporadic E reflections were observed from midnight to noon up to 4400 kilocycles during thirty-four hourly measurements and up to 6200 kilocycles or above during seven hourly measurements. From noon to midnight they were observed up to 4400 kilocycles during sixty-two hourly measurements and up to 6200 kilocycles or above during thirty hourly measurements.

Ionospheric disturbances associated with magnetic storms were rather numerous during July but usually not very severe. Several well marked ionospheric disturbances were observed during fairly minor magnetic disturbances. These results are typical for summer conditions in Washington. The nature of these ionospheric disturbances has been discussed in previous reports of this series and in other papers from the National Bureau of Standards. The ionospherically disturbed days listed approximately in the order of the severity of the ionospheric disturbances were as follows: July 22, 24, 23, 14, 25, 7, 20, and 10.

Out of 126 hours of night measurements of $f_F{}^x$ between 2300 and 0500 E.S.T. 16 values were more than 15 per cent below the undisturbed average. All of these occurred during the disturbed period of

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the early morning hours of July 22 and the two following nights. Values of $f_F{}^x$ for 38 hours of night measurements were more than 10 percent below the undisturbed average. All but three of these occurred on the early mornings and preceding evenings of the disturbed days listed in the preceding paragraph. For three hours of observations $f_F{}^x$ was over 15 per cent and for eight hours was over 10 per cent above the undisturbed average.

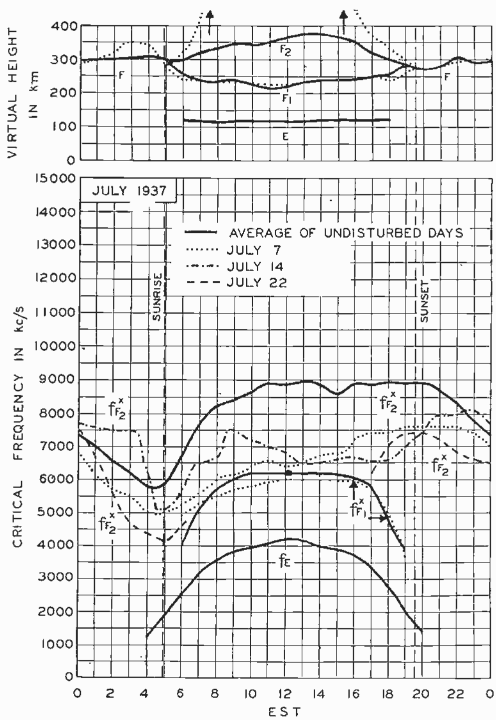

Fig. 1—Virtual heights and critical frequencies of the E, F_1, F_2, and F layers of the ionosphere for July, 1937. The solid-line graphs represent averages for the magnetically quiet days. The graphs for July 7, 14, and 22 represent conditions for ionospherically disturbed days.

Out of fifty-five hours of observations of $f_{F_2}{}^x$ on the disturbed days of July 7, 14, 22, 23, and 24 between the hours of 0600 and 2200 E.S.T. 21 values were over 25 percent below the undisturbed average. Out of thirty-four hours of observations between the same hours on July 21 and 28 no values were more than 10 per cent above or below the undisturbed average. Because the $f_{F_2}{}^x$ was usually above the range of the automatic recorder available the daytime measurements of $f_{F_2}{}^x$ were made principally by manual runs each Wednesday.

Emissions from station W8XAL 6060 kilocycles, 650 kilometers, were propagated regularly by the F layer at night except for short periods between midnight and 0600 E.S.T. on the magnetically disturbed days of July 7, 10, 22, 23, and 24. E layer transmission began on the average at 0603 E.S.T. and continued to 1852 E.S.T.

Emissions from station W1XK 9570 kilocycles, 600 kilometers, were propagated regularly by normal E layer from about 0800 E.S.T. to 1500 E.S.T. and by the F layer on magnetically quiet days from about 1730 to 2130 E.S.T., and frequently by sporadic E during the afternoon and evening.

Sudden disturbances of the ionosphere were frequent. They were marked by the following radio fade-outs, observed at Washington during July:[1]

TABLE I

Date	Beginning of fade-out	Beginning of recovery	Recovery complete	Location of transmitter	Minimum intensity
July 7	1847	1900	1920	Ohio, Mass., D.C.	0.0
July 8	1404	1409	1430	Ohio, Mass.	0.0
July 8	1727	1742	1840	Ohio, Mass.	0.0
July 9	1334	1344	1400	Ohio, Mass.	0.01
July 9	1419	—	1450	Ohio, Mass.	0.5
July 9	1610	—	1730	Ohio, Mass., D.C.	0.1
July 9	1928	1938	2003	Ohio, Mass.	0.05
July 10	1635	1649	1730	Ohio, Mass.	0.0
July 11	1536	1545	1558	Ohio, Mass.	0.05
July 11	1913	1958	2030	Ohio, Mass., D.C.	0.0
July 12	1806	1843	1910	Ohio, Mass.	0.1
July 13	1756	1803	1812	Ohio, Mass., Germany	0.0
July 16	1510	—	1730	Ohio	0.1
July 16	2250	—	2320	Ohio	0.5
July 17	1243	1250	1312	Ohio, Mass., D.C.	0.0
July 18	1841	1858	1918	Ohio, Mass., D.C.	0.0
July 19	1548	1604	1652	Ohio, Mass., D.C.	0.0
July 19	2020	—	2154	Ohio, D.C.	0.01
July 20	1945	2003	2021	Ohio, Mass., D.C.	0.0
July 21	1438	1450	1459	Ohio, Mass., D.C.	0.01
July 21	1515	—	1538	Ohio, Mass., D.C.	0.2
July 21	1743	—	1816	Ohio, D.C.	0.2
July 23	1948	—	2008	Ohio	0.1
July 24	1406	1414	1530	Ohio, D.C.	0.0
July 24	1740	—	1810	Ohio	0.1
July 24	1946	1951	1955	Ohio, D.C.	0.0
July 25	1630	—	1800	Ohio, D.C.	0.2
July 25	1959	2012	2038	Ohio, Mass., D.C.	0.0
July 25	2108	—	2130	Ohio	0.1
July 25	2138	2142	2152	Ohio, Mass.	0.0
July 25	2201	—	2228	Ohio, D.C.	0.05
July 26	1615	—	1630	Ohio	0.05
July 26	1635	—	1800	Ohio, Mass.	0.05
July 27	1506	1522	1550	Ohio, Mass.	0.0
July 27	2123	2128	2148	Ohio	0.0
July 28	2049	2102	2200	Ohio	0.1
July 29	1425	—	1448	Ohio	0.05
July 29	1538	1546	1605	Ohio, Mass.	0.0
July 31	1610	1800	2100	Ohio, Mass., D.C.	0.0

[1] All times G.M.T., minimum intensities given in terms of transmissions from station W8XAL, 6060 kc/s 650 km.

Emissions from station DJB, Berlin, 15,200 kilocycles, 7200 kilometers, were propagated regularly by F layer on magnetically quiet nights. These transmissions which were received during the evening hours at Washington were made during the early morning hours at Berlin and proved to be an excellent indicator of ionospheric disturbances associated with magnetic storms for the day following their reception at Washington. These transmissions failed partially or completely on the evening preceding the ionospherically disturbed days of July 7, 10, 22, 24, and 25.

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THE PHYSICAL REALITY OF SPACE AND SURFACE WAVES IN THE RADIATION FIELD OF RADIO ANTENNAS*

By
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Summary—*Evidence is presented which indicates that, notwithstanding the change in sign made by Sommerfeld in his 1926 paper on radio wave propagation, the radiation field of a vertical electric dipole may be separated into space and surface wave components. Sommerfeld's original concepts as to the characteristics of two such waves in radio transmission are largely substantiated. It is shown that a space and a surface wave are generated by a simple vertical dipole antenna at the surface of the earth and that this surface wave has the same wave tilt as the Sommerfeld surface wave. Evidence is given which would indicate that this surface wave travels around the curve of the earth in much the same manner as a guided wire wave travels around a bend on a wire. In the appendix formulas are given for the space and surface waves in the radiation fields of a horizontal electric dipole and of horizontal and vertical magnetic dipoles.*

IN 1909 Professor A. Sommerfeld[1] solved the general problem of the effect of the finite conductivity of the ground on the radiation from a short vertical antenna at the surface of a plane earth. Sommerfeld[2] in 1911 and Bruno Rolf[3] in 1930 published graphs of the ground wave "attenuation factor" which were based on the above solution; these graphs predicted such anomalies as "negative attenuation" and dips to zero of the ground wave field intensity at finite distances from the transmitting dipole. H. Weyl[4] in 1919, Professor Sommerfeld[5] in 1926, Balth. van der Pol and K. F. Niessen[6] in 1930, and W. H. Wise[7] in 1931 each obtained independent solutions of the Sommerfeld problem which agreed with his 1909 solution *except for a difference in one sign.* Apparently none of these authors noticed this discrepancy until the author, in a letter to the Editor of *Nature*[8] pointed it out and showed that it was responsible for the anomalies in propagation predicted by the Sommerfeld-Rolf graphs; an empirical formula (based on the van der Pol-Niessen solution) was also given for the ground wave "attenuation factor" which, for the first time, correctly took into account the effect of the dielectric constant of the ground. These results were later given in considerably greater detail in a recent issue of the Proceedings.[9]

In his original discussion[1] Sommerfeld found that it was possible to divide the ground wave field intensity into two parts, a space wave

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[1] Numbers refer to bibliography.

and a surface wave. Some of his discussion concerning these two waves is quoted here. "Two contrasting concepts arise which—at least in their general outlines—may be designated by the terms 'space waves' and 'surface waves.' In acoustics and in the majority of optical phenomena we are concerned with space waves. The Hertzian electrodynamic waves are also in the same category. The classical example of surface waves is found in hydrodynamics. In optics they appear in the less dense medium in the phenomena of total reflection as Voigt has shown both theoretically and experimentally. Further, electrodynamic waves on wires are typical surface waves. Finally in the domain of elasticity we have both types of waves clearly distinguishable in modern seismological observations.

"With which type are the waves utilized in wireless telegraphy to be identified? Are they like Hertzian waves in air or electrodynamic waves on wires?

"The first point of view would seem to predominate. It has been developed quantitatively by M. Abraham who made a substantial contribution in this field when he succeeded in developing laws for the propagation of the electric and magnetic force and their dependence on distance and azimuth, from the simple potential function previously developed by Hertz. . . .

"Now the diagrams plotted by Hertz in connection with his solution show that the electrical lines of force are perpendicular to the equatorial plane of the dipole. This is the basis of Abraham's application of Hertz' solution to wireless telegraphy *assuming a perfectly conductive ground,* on which the electrical lines of force also must terminate perpendicularly. The surface of the ground replaces the equatorial plane of the dipole. On this assumption the ground would have no other effect than preventing the space waves from the sender from penetrating the earth. . . .

"The opposite point of view, namely, that in wireless telegraphy waves similar to those guided by wires are involved and that the ground considerably influences the propagation of waves has been set forth on various occasions. It has received a quantitative development in a dissertation by Uller and by the recent work of Zenneck. Uller investigated a definite type of plane wave (Voigt calls them 'inhomogeneously plane') which are concentrated more or less on the border of the two media, earth and air. On the assumption of this form of wave Zenneck draws a number of remarkable conclusions with respect to the behavior of electric waves under various ground conditions and to the use of receivers and senders in directive telegraphy. . . .

"The term surface wave should not be interpreted to mean that the seat of the energy in the waves is really in the second medium as in the case of water waves or those due to seismic disturbances. On the contrary, the greater part of the energy is in the first medium, air, in the same manner as with waves on wires. The amplitude decreases slowly from the surface of the ground upwards, and rapidly downwards (skin effect). Heretofore, however, this interesting type of wave has been entirely hypothetical. There was no definite proof that such waves could be developed from the waves coming from the sender at a great distance. *The main task of the present investigation is to give this proof and to settle the question: space waves or surface waves?* Attention is called to the fact that the answer will not be unconditional and the same for all cases as our simplified assumptions and relations are correct only in given limiting cases without in general doing justice to the complexity of the phenomena. Similarly the terms incident and reflected light in general optical problems lose their exact meanings and merge into the conception of the optical field of diffraction phenomena. In the same manner there are various stages of transition between space and surface waves in our case, and a sharp division between them becomes practically impossible. Still the Uller-Zenneck surface waves appear as an important and occasionally predominant component of the electromagnetic field accompanied by space waves which on their part predominate under certain other conditions."

The purpose of this paper is to show that this separation of the radiated wave into a space wave and a surface wave is still possible and useful and is not invalidated by the change in sign which Sommerfeld made in his 1926 paper. The results obtained by C. R. Burrows,[10] and by W. H. Wise[11] confirm the results obtained by Sommerfeld[5] in 1926 and shown graphically by the author,[9] and might lead one to assume the nonexistence of a surface wave in radio propagation. Burrows[10] obtained experimental data over a fresh water lake which confirmed the conclusion reached by C. B. Feldman[12] who obtained experimental data over level land and showed that Rolf's graphs were invalid in those cases when the dielectric constant of the ground had to be taken into account; their data were in good agreement with the author's graphs. Wise showed[11] that his solution[7] of the Sommerfeld problem indicates that the asymptotic expansion of the wave function for a vertical electric dipole does not contain a term which may be identified with the Zenneck surface wave and thus confirmed the results obtained by the author[9] by expanding the van der Pol-Niessen[6] solution and using the well-known asymptotic expansion for the error function. This absence of the Zenneck surface wave wave function in the asymptotic expansion is the principal difference between the formulas obtained by Rolf[3] and the author.[9] It will now be shown that, although the Zenneck surface wave does not appear in the asymptotic field of a vertical electric dipole, nevertheless the expression for the vector electric field of such a dipole can be divided into two terms which may be readily identified with a space wave, which predominates at large distances above the earth, and a surface wave predominating near the surface of the earth and having a wave tilt and polarization the same as that of the Sommerfeld surface wave.

In a recent paper[13] the author derived a formula (based upon an expression for the wave potential given by van der Pol[14]) for the vector electric field of a vertical electric dipole, the formula being applicable at any point in space not too near the antenna. If we confine the following results to distances greater than several wave lengths from the antenna and let the vertical dipole be at the surface of a plane earth, the resulting simplified equations obtained from (55) and (70) in the above paper[13] may be combined and expressed as two components $\mathbf{E}_{sp}{}^v$ and $\mathbf{E}_{su}{}^v$ which will be identified as space and surface waves:

$$\mathbf{E}_{sp}{}^v = ik\cos\psi(1+R_v)\frac{e^{i(kR-\omega t)}}{R}\boldsymbol{\psi}. \tag{1}$$

$$\mathbf{E}_{su}{}^v = ik(1-R_v)F\frac{e^{i(kR-\omega t)}}{R}$$
$$\left[\mathbf{k}+\mathbf{r}\cos\psi\left(1+\frac{\sin^2\psi}{2}\right)u\sqrt{1-u^2\cos^2\psi}\right] \tag{2}$$

where $R>>\lambda$.

$$R_v = \frac{\sin\psi - u\sqrt{1-u^2\cos^2\psi}}{\sin\psi + u\sqrt{1-u^2\cos^2\psi}}. \tag{3}$$

$$u^2 = 1/(\epsilon + ix). \tag{4}$$

$$x = 1\cdot 8\cdot 10^{18}\sigma_{\mathrm{e.m.u.}}/f_{kc}. \tag{5}$$

$$F = \left[1 + i\sqrt{\pi w}\,e^{-w}\,\mathrm{erfc}\,(-i\sqrt{w})\right]. \tag{6}$$

$$w = \frac{ikRu^2(1-u^2\cos^2\psi)}{2}\left[1+\sin\psi/u\sqrt{1-u^2\cos^2\psi}\right]^2. \tag{7}$$

$$\mathrm{erfc}\,(x) = \frac{2}{\sqrt{\pi}}\int_x^\infty e^{-u^2}du. \tag{8}$$

$k=2\pi/\lambda$, $\omega=2\pi f$, $\mathbf{k}$ and $\mathbf{r}$ are unit vectors, respectively parallel and perpendicular to the vertical dipole. $\boldsymbol{\psi}=(\mathbf{k}\cos\psi-\mathbf{r}\sin\psi)$ is also a unit vector. The symmetry of the problem permits the received field to be specified in terms of the polar coordinates R and ψ of the receiving point where ψ is the angle measured from the surface of the earth (see Fig. 1). λ is the wave length and is to be expressed in the same units as the other quantities having the dimension length with which it is associated. ϵ is the dielectric constant of the ground referred to air as unity, σ is the conductivity of the ground expressed in electromagnetic

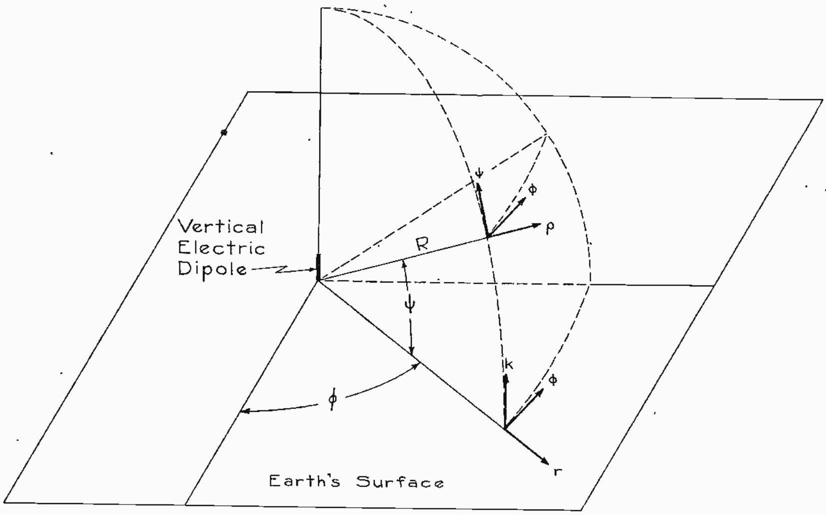

Fig. 1—Geometry for radiation formulas.

units, and f is the frequency in kilocycles per second. R_v is the Fresnel reflection coefficient of a plane wave with its electric vector parallel to the plane of incidence and with angle of incidence $\left(\frac{\pi}{2}-\psi\right)$. F is the function discussed by the author in an earlier paper,[9] its absolute value being equal to the "ground wave attenuation factor" when $\psi=0$. When $\psi=0$, $|w|$ is the "numerical distance" which played such a prominent part in Sommerfeld's original solution.[1] The sum of (1) and (2) represents the total radiation field of a vertical electric dipole; these equations apply at any point in space above the surface of a plane earth such that $R>>\lambda$ and may be used for any frequency or set of ground constants found in practise. (1) is seen to be a plane polarized space wave and reduces in the case of a perfectly conducting earth (for which $R_v=1$) to the well-known Abraham solution. The effect of the finite conductivity of the earth is to reduce the intensity of this space wave by the factor $(1+R_v)/2$ and to introduce the surface wave given by (2) since in this case $R_v \neq 1$. The properties which identify (2) with a surface wave are its forward tilt and polarization (as determined by the quantity in the square brackets in (2) which indicates a similar form to that of the Sommerfeld surface wave), and its predominance near the surface of the earth; in fact at the surface it represents the total field since here $R_v = -1$. Conversely at large heights above the ground the surface wave becomes negligible and the total field is effectively the space wave.

It will be noted that, although our surface wave has nearly the same wave tilt as the Zenneck surface wave, it is not attenuated exponentially at large distances as was the Zenneck wave; this is due to the fact that the Zenneck wave was a plane wave guided along a plane imperfectly conducting surface while our surface wave originates in an antenna and its "attenuation factor" varies with distance first exponentially and finally, at large distances, inversely with the distance.

Another property of our surface wave is that it supplies all of the energy to the ground currents. This may be shown by determining the Poynting vector, which shows the direction of energy flow in the wave, for the space and surface waves separately. Equation (117) in a recent paper by the author[13] may be used to determine the magnetic vector for each of the two waves:

$$\mathbf{H}_{sp}{}^{v} = -ik \cos\psi (1 + R_v) \frac{e^{i(kR-\omega t)}}{R} \boldsymbol{\phi}. \tag{9}$$

$$\mathbf{H}_{su}{}^{v} = -ik \cos\psi \left(1 + \frac{\sin^2\psi}{2}\right)(1 - R_v) F \frac{e^{i(kR-\omega t)}}{R} \boldsymbol{\phi} \tag{10}$$

where $R >> \lambda$ and $\boldsymbol{\phi} = \mathbf{k} \times \mathbf{r}$.

The Poynting vector $\mathbf{S} = \dfrac{c}{4\pi} \mathbf{E} \times \mathbf{H}$ may now be determined separately for the space and surface waves:

$$\mathbf{S}_{sp}{}^{v} = \left[ik \cos\psi (1 + R_v) \frac{e^{i(kR-\omega t)}}{R}\right]^2 \boldsymbol{\varrho} \tag{11}$$

where $\boldsymbol{\varrho} = \mathbf{r} \cos\psi + \mathbf{k} \sin\psi$.

$$\mathbf{S}_{su}{}^{v} = \cos\psi \left(1 + \frac{\sin^2\psi}{2}\right)\left[ik(1 - R_v) F \frac{e^{i(kR-\omega t)}}{R}\right]^2 \left[\mathbf{r} - \mathbf{k} \cos\psi \left(1 + \frac{\sin^2\psi}{2}\right) u\sqrt{1 - u^2 \cos^2\psi}\right]. \tag{12}$$

Thus we see that the energy in the space wave flows in a direction normal to the surface of a hemisphere centered on the base of the antenna while a part of the energy in the surface wave flows downwards toward the ground and is thus the seat of the energy of the ground currents.

In addition to showing the physical reality of a space wave and a surface wave in the radiation field of a vertical antenna, the above separation of the total field into two components is useful in the solution of problems requiring a knowledge of sky waves which arrive at the receiver after reflection at the ionosphere. In such problems we require the intensity of the field at the ionosphere and at this large height above the earth the space wave alone constitutes the total field. The intensity of this space wave is easily computed, even in the case of rather elaborâte antenna systems.

All of the above formulas are strictly applicable only to a plane earth. It is of interest to quote an additional paragraph at this point from Sommerfeld's 1909 paper.[1] "Our theory presupposes a plane interface. It would not be difficult to extend the solution to the curved surface of the earth; only the discussion of the series thus produced would lead to expansions in spherical and cylindrical harmonics. But also without knowledge of this solution we may say that *the conditions are changed in favor of the surface wave by the curvative of the earth* since the space waves are screened off by the curvature of the earth, unless they do overcome it by a diffraction process, but the surface waves are not noticeably obstructed. It is quite possible that the space wave is eliminated by the curvature of the earth at large numerical distances and that it predominates over the surface wave only for very short numerical distances. In popular treatises on wireless telegraphy (Poincaré, for example) the effect of diffraction in overcoming the curvature of the earth seems to be overestimated. When in these treatises reference is made to the magnitude of the wave length, with which the bending increases, it should be remembered that only the ratio between the wave length and the radius of curvature of the obstacle which is to be overcome is of importance and that this ratio is no more favorable for the waves of wireless telegraphy and the curvature of the earth than for visible light and a moderàtely rounded edge." Thus we see that Sommerfeld expected that the surface wave would be only slightly affected by the curvature of the earth since it is guided around the curve of the earth in much the same manner as an electrodynamic wave on a wire passes around a bend with a comparatively small loss of energy. Since the Sommerfeld attenuation formula for a wave at the surface of the earth is identical to the "attenuation factor" in our surface wave, we now have an explanation of the success of the Sommerfeld plane earth formula at distances far beyond the line of sight.[15] In fact it has been shown theoretically[9] that the ground wave "attenuation factor" at ordinary radio communication frequencies is very little affected by diffraction at distances less than about 100 miles (i.e., it was shown that diffraction modified the ground wave field intensity at distances greater than a "numerical distance" p_x which depended on the frequency and conductivity, the resulting real distance being of the order of 100 miles).

If our surface wave is truly a guided wave, then we would expect the curvature of the earth to affect it differently than the space wave at points below the line of sight. This is strikingly illustrated by a comparison of some results obtained by B. Wwedensky[16] for a curved earth with those obtained by the author[13] for a plane earth showing the theoretical variation of received field intensity with height above the earth. Fig. 2 in the latter paper shows the relative importance of the surface and space waves at various heights above the ground for several different distances at a frequency of 1000 kilocycles over an earth of conductivity $\sigma = 10^{-13}$ e.m.u. The dotted curve corresponds to the surface wave while the long dashed curve (and at high angles the solid curve) represents the space wave. The predominance of the surface wave at low heights and of the space wave at large heights is clearly evident from this figure. At this frequency and conductivity (at which the dielectric constant of the ground has little effect, i.e., $x >> \epsilon$), it is evident that the space and surface waves are out of phase near the surface of the earth which causes an initial decrease in the intensity of the total received wave as the receiving antenna is raised above the surface. In Figs. 1 and 2 of the paper by Wwedensky, this initial reduction in field is shown to be smaller in the case of the curved earth solution than for the plane earth solution as would be expected if the curvature of the earth affected the space wave differently than the surface wave. Although the above evidence is illuminating the final establishment of Sommerfeld's view that the surface wave is similar to a guided wave on a wire must await further theoretical and experimental studies.

Appendix I

The Space and Surface Waves in the Radiation Field of a Horizontal Electric Dipole.

We may determine the space and surface waves in the radiation field (i.e., $R >> \lambda$) of a horizontal electric dipole at the surface of the earth and lying in the plane $\phi = 0$ by using (72), (73), and (74) in an earlier paper.[13]

$$\mathbf{F}_{sp}{}^{h} = ik\,\frac{e^{i(kR-\omega t)}}{R}\,\{\cos\phi\sin\psi(1-R_v)\boldsymbol{\psi} + \sin\phi(1-R_h)\boldsymbol{\phi}\}. \tag{13}$$

$$\begin{aligned}\mathbf{F}_{su}{}^{h} = ik\,\frac{e^{i(kR-\omega t)}}{R}\Bigg\{&\cos\phi u\sqrt{1-u^2\cos^2\psi}(1-R_v)F\\ &\cdot\left[\cos\psi\left(1+\frac{\sin^2\psi}{2}\right)\mathbf{k} + u\sqrt{1-u^2\cos^2\psi}\right.\\ &\left.\cdot\left(\frac{1-\sin^4\psi-\dfrac{(1+R_h)G}{(1-R_v)u^2F}}{1-u^2\cos^2\psi}\right)\mathbf{r}\right] + \sin\phi(1+R_h)G\boldsymbol{\phi}\Bigg\}\end{aligned} \tag{14}$$

where

$$R_h = \frac{\sqrt{1-u^2\cos^2\psi}-u\sin\psi}{\sqrt{1-u^2\cos^2\psi}+u\sin\psi}, \tag{15}$$

$$G = [1+i\sqrt{\pi v}e^{-v}\operatorname{erfc}(-i\sqrt{v})], \tag{16}$$

$$v = \frac{ikR(1-u^2\cos^2\psi)}{2u^2}\,[1+u\sin\psi/\sqrt{1-u^2\cos^2\psi}]^2. \tag{17}$$

R_h is the Fresnel reflection coefficient of a plane wave with its electric vector normal to the plane of incidence. (13) shows that the electric vector of the space wave for a horizontal electric dipole lies in a plane normal to the radius vector ρ. In the direction $\phi=0$ the electric vector of the space wave is plane polarized and lies in the vertical plane $\phi=0$. In the direction $\phi=\pi/2$ the electric vector of the space wave is plane polarized in a horizontal plane. For intermediate directions the electric vector of the space wave is elliptically polarized in the plane normal to the radius vector ρ.

At large distances such that $|w|>20$, we may use the asymptotic expansions for F and G and find that $G=u^4F$ asymptotically. Substituting this in (14) we see that the electric vector in the surface wave of a horizontal electric dipole is mostly in the plane normal to the unit vector ϕ and has a wave tilt and polarization near the earth nearly the same as the surface wave in the radiation field of a vertical electric dipole but with an intensity which is smaller by the factor $\cos\phi u\sqrt{1-u^2\cos^2\psi}$. This is the factor discussed by H. von Hoerschelmann[17] in his paper on the effect of the earth on the directional characteristics of a flat top antenna. There is also a small horizontal component in the surface wave; it is usually negligible since its intensity is u^3 times the intensity of the vertical component.

Appendix II

The Space and Surface Waves in the Radiation Field of Loop Antennas

A. The Horizontal Magnetic Dipole

A loop antenna with its largest dimension small with respect to a wave length and its axis normal to the plane $\phi=0$ is equivalent to the horizontal magnetic dipole discussed in an earlier paper.[13] From (75), (76), and (77) in that paper we obtain for the space and surface waves in the radiation field:

$$\mathbf{E}_{sp}{}^{mh} = ik\frac{e^{i(kR-\omega t)}}{R}\left\{\cos\phi(1+R_v)\boldsymbol{\psi} + \sin\phi\sin\psi(1+R_h)\boldsymbol{\phi}\right\}. \qquad 18)$$

$$\mathbf{E}_{su}{}^{mh} = ik\frac{e^{i(kR-\omega t)}}{R}\left\{\cos\phi(1-R_v)F\left[\cos\psi\left(1+\frac{\sin^2\psi}{2}\right)\mathbf{k}\right.\right.$$

$$\left.\left. + u\sqrt{1-u^2\cos^2\psi}\,\mathbf{r}\right] + \sin\phi\frac{\sqrt{1-u^2\cos^2\psi}}{u}(1+R_h)G\boldsymbol{\phi}\right\}. \qquad (19)$$

We see that the components of the space and surface waves of the horizontal magnetic dipole which lie in the vertical plane normal to the unit vector $\boldsymbol{\phi}$ are nearly the same as those for a vertical electric dipole except for the factor $\cos\phi$. In addition, there is a horizontally polarized component in both the space and surface waves.

B. The Vertical Magnetic Dipole

The electric vectors in the space and surface waves of a vertical magnetic dipole at the surface of the earth are horizontally polarized.[13]

$$\mathbf{E}_{sp}{}^{mv} = ik\cos\psi(1-R_h)\frac{e^{i(kR-\omega t)}}{R}\boldsymbol{\phi}. \qquad (20)$$

$$\mathbf{E}_{su}{}^{mv} = ik(1+R_h)G\left(\cos\psi+\frac{\sin^2\psi}{2}\right)\frac{e^{i(kR-\omega t)}}{R}\boldsymbol{\phi}. \qquad (21)$$

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THE PROPAGATION OF RADIO WAVES OVER THE SURFACE OF THE EARTH AND IN THE UPPER ATMOSPHERE*

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PART II

The Propagation from Vertical, Horizontal, and Loop Antennas over a Plane Earth of Finite Conductivity

Summary.—*Completely general formulas are given for computing at any point above a plane earth of finite conductivity the vector electric field for a source which may be a combination of vertical and horizontal electric dipoles or a loop antenna with its axis parallel or perpendicular to the earth. As illustrations of the above general methods, formulas are derived for the ground-wave radiation from (1) a grounded vertical antenna carrying a sinusoidal current distribution and (2) elevated vertical and horizontal half-wave antennas. The "effective height" of the grounded vertical antenna is determined as a function of the ground constants, and this formula is then used to determine the effect of the ground constants on the ground-wave field intensity in the neighborhood of a quarter-wave antenna. The formulas are also used to show the influence of antenna height on the attenuation of high and ultra-high frequencies. The forward tilt, i.e.,* E_r/E_z, *which occurs for the electric vector lying in the vertical plane passing through the antenna, is also easily computed from the formulas given and is shown graphically. An expression for the Poynting vector is derived, and it is shown that a part of the energy in the wave near the ground flows downward into the ground.*

1. Introduction

IN PART I of this paper[1] a formula was given for the vertical component of the ground-wave field intensity at the surface of a plane earth of finite conductivity and radiated from a short vertical antenna at the surface of the earth. In this part, completely general formulas will be derived for the vector electric field at any point above the surface of a plane earth of finite conductivity for a radiating system which may consist of any configuration of vertical and horizontal electric dipoles. Formulas also will be given for loop antennas with their axes parallel and perpendicular to the earth. These formulas, in addition to making possible the determination of the various components of the electric field at different heights above the earth, may also be

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[1] Proc. I.R.E., vol. 24, pp. 1367–1387; October, (1936).

used to determine the effects of transmitting antenna height on the attenuation of the ground waves as well as the effects of the ground constants on the effective heights of antennas. It will be found that the attenuation formula of Part I may be used without appreciable error providing the transmitting and receiving antennas are less than a half wave length above the earth and the distance along the ground is greater than a wave length. In all other cases, the formulas to be given below should be used.

2. The Vector Electric Field Intensity from Vertical and Horizontal Dipoles over a Plane Earth of Finite Conductivity

Recent results obtained by Balth. van der Pol[2] and W. H. Wise[3] make possible comparatively simple expressions for the vector electric field intensity at great distances from and in the neighborhood of an antenna carrying an arbitrary distribution of current and located near a plane earth of finite conductivity. Let the origin of our co-ordinate system be at the surface of the earth under the antenna and choose a right-handed set of unit vectors **i**, **j**, **k** in the direction of x, y, and z. For the specification of the components of the vector electric field, we shall use cylindrical co-ordinates r, ϕ, and z, and a corresponding set of right-handed unit vectors **r**, $\boldsymbol{\phi}$, and **k**. As was shown by H. von Hoerschelmann,[4] the wave potential of a unit[5] vertical dipole placed over an imperfectly conducting earth has only a single component $\Pi_z{}^v$ while that for a unit horizontal dipole parallel to the x axis has two components $\Pi_x{}^h$ and $\Pi_z{}^h$. In either case the vector electric field is given by

$$\mathbf{E} = ik\left[\Pi + \frac{1}{k^2}\nabla\nabla\cdot\Pi\right]e^{-i\omega t} \tag{1}$$

$$\Pi_z{}^v = \frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2} + V \tag{2}$$

where,

$$V = \int_0^\infty \frac{2}{l + u^2 m} J_0(\lambda r) e^{-(a+z)l} \lambda d\lambda \tag{3}$$

$$\Pi_x{}^h = -\frac{e^{ikR_1}}{R_1} + \frac{e^{ikR_2}}{R_2} - H \tag{4}$$

[2] *Physica*, vol. 2, pp. 843–853; August, (1935).
[3] *Bell Sys. Tech. Jour.*, vol. 8, pp. 662–671; October, (1929).
[4] *Jahr. der Drahtl. Tel. und Tel.*, vol. 5, pp. 14–34 and 188–211; September, (1911).
[5] A unit dipole is here considered to be one with an infinitesimal length dl and a unit moment $\mathbf{I}dl$ where $\mathbf{I}$ denotes the current.

where,

$$H = \int_0^\infty \frac{2}{l+m} J_0(\lambda r) e^{-(a+z)l} \lambda d\lambda \tag{5}$$

$$II_z^h = -\cos\phi \int_0^\infty \frac{2(1-u^2)}{(l+m)(l+u^2 m)} J_0'(\lambda r) e^{-(a+z)l} \lambda^2 d\lambda. \tag{6}$$

The unit vectors **i**, **j**, **k**, **r**, and **ϕ**, and R_1, R_2, a, r, ϕ, x, y, z are adequately defined in Fig. 1. Sin $\psi = z/R$, and $R^2 = x^2+y^2+z^2$, $k = 2\pi/\lambda$,

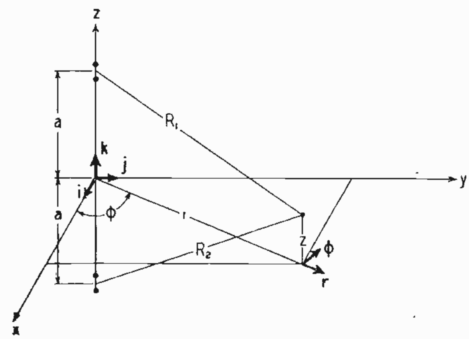

Fig. 1.—Geometry for dipole radiation formulas.

$k_2^2 = k^2(\epsilon + ix)$, $x = 1.8 \cdot 10^{13}\sigma_{\text{emu}}/f_{kc}$ where ϵ is the dielectric constant of the ground referred to air as unity while σ is the conductivity of the ground in electromagnetic units and f is the frequency in kilocycles per second. $l^2 = \lambda^2 - k^2$, $m^2 = \lambda^2 - k_2^2$, and $u = k/k_2$. (In Part I of this paper $k/k_2 = y$.) x is given graphically in Part I. Van der Pol[2] has given simplified expressions for V and H which will be further simplified in this paper. Thus we have only to express the electric field components in terms of these two integrals. Following Wise,[3] we can write

$$\nabla \cdot \mathbf{II}^h = \frac{\partial}{\partial x} II_x^h + \frac{\partial}{\partial z} II_z^h$$

$$\frac{\partial}{\partial z} II_z^h = \frac{\partial}{\partial x} \int_0^\infty \frac{2(1-u^2)l}{(l+m)(l+u^2 m)} J_0(\lambda r) e^{-(a+z)l} \lambda d\lambda$$

and when this is substituted in the above and combined with $\partial/\partial x II_x^h$ we obtain

$$\nabla \cdot \mathbf{II}^h = -\frac{\partial}{\partial x}\left[\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2} + u^2 V\right]. \tag{7}$$

The integral in II_z^h must also be eliminated from the expression for E_z^h. Using (7), we have

$$E_z{}^h = ik\left\{II_z{}^h - \frac{1}{k^2}\frac{\partial^2}{\partial z\partial x}\left(\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2} + u^2V\right)\right\}e^{-i\omega t}$$

$$= -\frac{i}{k}\cos\phi\left\{\frac{\partial^2}{\partial r\partial z}\left(\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2}\right)\right.$$

$$+ 2\int_0^\infty\left[\frac{(1-u^2)k^2}{(l+m)(l+u^2m)}\right.$$

$$\left.\left. - \frac{u^2l}{l+u^2m}\right]J_0{}'(\lambda r)e^{-(a+z)l}\lambda^2 d\lambda\right\}e^{-i\omega t}$$

$$= \frac{i}{k}\cos\phi\left\{\frac{\partial^2}{\partial r\partial z}\left(\frac{e^{ikR_2}}{R_2} - \frac{e^{ikR_1}}{R_1}\right)\right.$$

$$\left. + 2\int_0^\infty \frac{u^2m}{l+u^2m}J_0{}'(\lambda r)e^{-(a+z)l}\lambda^2 d\lambda\right\}e^{-i\omega t}$$

$$= \frac{i}{k}\cos\phi\left\{\frac{\partial^2}{\partial r\partial z}\left(\frac{e^{ikR_2}}{R_2} - \frac{e^{ikR_1}}{R_1}\right)\right.$$

$$\left. + 2\frac{\partial}{\partial r}\int_0^\infty\left[1 - \frac{l}{l+u^2m}\right]J_0(\lambda r)e^{-(a+z)l}\lambda d\lambda\right\}e^{-i\omega t}.$$

Finally, noting that

$$\frac{\partial}{\partial z}\frac{e^{ikR_2}}{R_2} = -\int_0^\infty J_0(\lambda r)e^{-(a+z)l}\lambda d\lambda,$$

we obtain

$$E_z{}^h \equiv E_z{}'^h\cos\phi = \frac{i}{k}\cos\phi\,\frac{\partial^2}{\partial r\partial z}\left(V - \frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2}\right)e^{-i\omega t}. \tag{8}$$

Thus the integral in $II_z{}^h$ has been eliminated. We may now write the appropriate expressions for the remaining components of $\mathbf{E}^v$ and $\mathbf{E}^h$

$$E_z{}^v = ik\left(II_z{}^v + \frac{1}{k^2}\frac{\partial^2}{\partial z^2}II_z{}^v\right)e^{-i\omega t} \tag{9}$$

$$E_r{}^v = \frac{i}{k}\frac{\partial^2}{\partial r\partial z}II_z{}^v e^{-i\omega t} \tag{10}$$

$$E_\phi{}^v = 0 \tag{11}$$

$$E_r{}^h \equiv E_r{}'^h\cos\phi$$

$$= ik\cos\phi\left(II_x{}^h - \frac{1}{k^2}\frac{\partial^2}{\partial r^2}\left[\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2} + u^2V\right]\right)e^{-i\omega t} \tag{12}$$

$$E_\phi{}^h \equiv E_\phi{}'^h \sin\phi$$
$$= -ik \sin\phi\left(II_x{}^h - \frac{1}{k^2 r}\frac{\partial}{\partial r}\left[\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2} + u^2 V\right]\right)e^{-i\omega t}. \quad (13)$$

For a horizontal dipole parallel to the y axis, the appropriate expressions for the components of $\mathbf{E}^h$ may be obtained from (8), (12), and (13) by replacing $\cos\phi$ and $\sin\phi$ by $\sin\phi$ and $\cos\phi$, respectively.

The above expressions are exact, but are not very useful in their present form. They will be simplified in several later sections for practical numerical computation.

3. The Vector Electric Field from an Arbitrary Current Distribution over a Plane Earth of Finite Conductivity

In order to determine the field from an antenna with an arbitrary current distribution, it is necessary to resolve the various components of the current along the **i**, **j**, and **k** directions, integrate along the antenna, and thus determine the resulting fields from each component. This may be expressed most compactly by using the following linear vector function:

$$\mathbf{E} = \int \mathbf{I}\cdot \mathrm{A}\,dl \quad (14)$$

where,

$$\begin{aligned} \mathrm{A} \equiv\ & \mathbf{ir}E_r{}'^h \cos\phi + \mathbf{i}\boldsymbol{\phi}E_\phi{}'^h \sin\phi + \mathbf{ik}E_z{}'^h \cos\phi \\ &+ \mathbf{jr}E_r{}'^h \sin\phi + \mathbf{j}\boldsymbol{\phi}E_\phi{}'^h \cos\phi + \mathbf{jk}E_z{}'^h \sin\phi \\ &+ \mathbf{kr}E_r{}^v \qquad + 0 \qquad\qquad + \mathbf{kk}E_z{}^v. \end{aligned} \quad (15)$$

Examples of the use of this representation of the vector electric field will be given later.

4. The Loop Antenna

The electric field from a loop antenna parallel to the x–z plane can be determined from paragraph 3 above by adding together the fields from four current elements, two vertical and two horizontal, but it is simpler to consider such a loop antenna as a magnetic dipole parallel to the y axis. This correspondence holds only when the largest dimensions of the loop are small in comparison to the wave length. In this case the vector electric field may be expressed

$$\mathbf{E}^{mh} = -\nabla \times \mathbf{\Pi}^{mh} e^{-i\omega t} \quad (16)$$

[6] "Differentialgleichungen der Physik, Frank —v Mises," vol. 2, pp. 949–953

and A. Sommerfeld[6] has shown that $\mathbf{\Pi}^{mh}$ has two components,

$$\mathbf{\Pi}^{mh} = \mathbf{j}\Pi_y^{mh} + \mathbf{k}\Pi_z^{mh} = \mathbf{r}\Pi_y^{mh}\sin\phi + \boldsymbol{\phi}\Pi_y^{mh}\cos\phi + \mathbf{k}\Pi_z^{mh} \tag{17}$$

$$\Pi_y^{mh} = -\Pi_z^v \tag{18}$$

$$\Pi_z^{mh} = \tan\phi\,\Pi_z^h \tag{19}$$

$$E_r^{mh} = -\left\{\frac{1}{r}\frac{\partial}{\partial\phi}(\tan\phi\,\Pi_z^h) + \cos\phi\frac{\partial\Pi_z^v}{\partial z}\right\}e^{-i\omega t} \tag{20}$$

$$E_\phi^{mh} = \sin\phi\left[\frac{\partial\Pi_z^v}{\partial z} + \frac{1}{\cos\phi}\frac{\partial\Pi_z^h}{\partial r}\right]e^{-i\omega t} \tag{21}$$

$$E_z^{mh} = \cos\phi\frac{\partial\Pi_z^v}{\partial r}e^{-i\omega t} \tag{22}$$

$$\begin{aligned}\frac{\partial\Pi_z^v}{\partial z} + \frac{1}{\cos\phi}\frac{\partial\Pi_z^h}{\partial r} &= \frac{\partial}{\partial z}\left(\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2}\right)\\ &\quad - \int_0^\infty \frac{2l}{l+u^2m}J_0(\lambda r)e^{-(a+z)l}\lambda d\lambda\\ &\quad - 2\frac{\partial}{\partial r}\int_0^\infty \frac{1-u^2}{(l+m)(l+u^2m)}J_0'(\lambda r)e^{-(a+z)l}\lambda^2 d\lambda \qquad (23)\\ &= \frac{\partial}{\partial z}\left(\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2}\right)\\ &\quad - 2\int_0^\infty \frac{[l(l+m)J_0(\lambda r) + (1-u^2)J_0''(\lambda r)\lambda^2]e^{-(a+z)l}}{(l+m)(l+u^2m)}\lambda d\lambda. \qquad (24)\end{aligned}$$

In (24) $J_0''(\lambda r)$ may be replaced by $-J_0(\lambda r)$ and the second order term $J_0'(\lambda r)/\lambda r$ appearing in Bessel's equation may be neglected since it is of the second order in $1/r$. Thus (24) becomes

$$= \frac{\partial}{\partial z}\left(\frac{e^{ikR_1}}{R_1} - \frac{e^{ikR_2}}{R_2}\right) - 2\int_0^\infty\left[1 - \frac{l}{l+m}\right]J_0(\lambda r)e^{-(a+z)l}\lambda d\lambda \tag{25}$$

and finally

$$\frac{\partial\Pi_z^v}{\partial z} + \frac{1}{\cos\phi}\frac{\partial\Pi_z^h}{\partial r} = \frac{\partial}{\partial z}\left(\frac{e^{ikR_1}}{R_1} + \frac{e^{ikR_2}}{R_2} - H\right). \tag{26}$$

Neglecting the second order term in $1/r$ appearing in (20), we have

$$E_z^{mh} = \cos\phi\frac{\partial}{\partial r}\Pi_z^v e^{-i\omega t} \tag{27}$$

$$E_r^{mh} = -\cos\phi \frac{\partial}{\partial z} \Pi_z^v e^{-i\omega t} \tag{28}$$

$$E_\phi^{mh} = +\sin\phi \frac{\partial}{\partial z}\left(\frac{e^{ikR_1}}{R_1} + \frac{e^{ikR_2}}{R_2} - H\right)e^{-i\omega t}. \tag{29}$$

The vertical magnetic dipole will also be considered, although it would seem to be a case of little practical importance since it corresponds to a loop antenna with its axis perpendicular to the earth. In this case

$$\mathbf{\Pi}^{mv} = -\mathbf{k}\Pi_x^h \tag{30}$$

$$\mathbf{E}^{mv} = -\nabla \times \mathbf{\Pi}^{mv} e^{-i\omega t} = -\boldsymbol{\phi} \frac{\partial}{\partial r} \Pi_x^h e^{-i\omega t}. \tag{31}$$

5. Practical Formulas for $\mathbf{E}^v$, $\mathbf{E}^h$, $\mathbf{E}^{mh}$, and $\mathbf{E}^{mv}$

In the preceding paragraphs, the problem of determining the vector electric field intensity from an antenna over an imperfectly conducting earth has been reduced to the problem of computing V and H and their partial derivatives with respect to r and z. Formulas for V and H will be given below which may readily be interpreted numerically but which are valid only at a distance of several wave lengths from the antenna.

Van der Pol[2] has given the following formula for V which is valid when the distance R_2 is greater than a few wave lengths from the antenna and is accurate to the first order in u^2:

$$V = 2\left[\frac{e^{ikR_2}}{R_2} + ik_2 \int_0^\infty \frac{e^{ik(R'+s/u)}}{R'} ds\right] \tag{32}$$

$$R'^2 = r^2 + (a + z + s/u^2)^2. \tag{33}$$

Since the coefficient of s in the exponential in the integral in (32) has a large negative real part, most of the value of the integral is obtained for small values of s and we may write

$$R' = R_2 + \frac{(a+z)s}{R_2 u^2} + \frac{s^2}{2R_2 u^4} \tag{34}$$

$$V = 2\,\frac{e^{ikR_2}}{R_2}\left[1 + ik_2 \int_0^\infty e^{ik[(a+z)s/R_2u^2 + s^2/2R_2u^4 + s/u]} ds\right]. \tag{35}$$

Using the following identity which is valid when a has a finite imaginary part,

$$\int_0^\infty e^{[a^2s^2+bs]} ds \equiv i\,\frac{\sqrt{\pi}}{2a}\, e^{-b^2/4a^2} erfc(-i\sqrt{b^2/4a^2}) \tag{36}$$

where,

$$erfc(x) \equiv \frac{2}{\sqrt{\pi}}\int_x^\infty e^{-u^2}du \equiv i\frac{2}{\sqrt{\pi}}\int_{i\infty}^{ix} e^{u^2}du \tag{37}$$

we may write (35)

$$V = 2\frac{e^{ikR_2}}{R_2}[1 + i\sqrt{\pi p'}e^{-w'}erfc(-i\sqrt{w'})] \tag{38}$$

$$p' = ikR_2u^2/2 \tag{39}$$

$$w' = p'[1 + (a+z)/uR_2]^2. \tag{40}$$

Using the asymptotic expansion

$$i\sqrt{\pi p'}e^{-w'}erfc(-i\sqrt{w'})$$
$$= -[1 + (a+z)/uR_2]^{-1}\left(1 + \frac{1}{2w'} + \frac{1\cdot 3}{(2w')^2} + \cdots\right) \tag{41}$$

we obtain

$$V = 2\frac{e^{ikR_2}}{R_2}\left[1 - (1 + (a+z)/uR_2)^{-1}\left(1 + \frac{1}{2w'}\right)\right] \tag{42}$$

where,

$$|w'| > 20.$$

Wise[3] has given an exact asymptotic expansion for V which may be written

$$V = 2\frac{e^{ikR_2}}{R_2}[1 - (1 + \sin\psi'/u\sqrt{1 - u^2\cos^2\psi'})^{-1}] \tag{43}$$

where,

$$\sin\psi' \equiv (z+a)/R_2. \tag{44}$$

It is evident that for sufficiently large values of w' the approximate formula (42) is equal to (43) except that u must be multiplied by the factor $\sqrt{1-u^2\cos^2\psi'}$. Since it has been found in practice that ϵ is always greater than 5, $|u^2|$ is less than 0.2 for any frequency or conductivity so that this factor is usually near unity. The absence of this factor in the asymptotic expansion of V is not surprising since it may be shown that the integral expression in (32) is valid only to the first order in u/R_2 due to approximations made by van der Pol. However, it is a simple matter to introduce this factor in (32) so that it will agree asymptotically with the exact asymptotic expansion for V. We have then

$$V = 2\left[\frac{e^{ikR_2}}{R_2} + ik_2\sqrt{1 - u^2\cos^2\psi'}\int_0^\infty \frac{e^{ik(R'+s\sqrt{1-u^2\cos^2\psi'}/u)}}{R'}\,ds\right]. \tag{45}$$

In order to differentiate V with respect to z, we note that

$$\frac{\partial R'}{\partial z} = u^2\frac{\partial R'}{\partial s}$$

and integrate by parts.

$$\frac{\partial V}{\partial z} = ik\left[\sin\psi' 2\frac{e^{ikR_2}}{R_2}(1 - 1/ikR_2) - u\sqrt{1 - u^2\cos^2\psi'}V\right] \tag{46}$$

$$\begin{aligned}\frac{\partial^2 V}{\partial z^2} = -k^2\Bigg[\sin^2\psi' 2\frac{e^{ikR_2}}{R_2} &- u\sqrt{1-u^2\cos^2\psi'}\sin\psi' 2\frac{e^{ikR_2}}{R_2}\left(1 - \frac{1}{ikR_2}\right)\\ &+ \left(\frac{1}{ikR_2} - \frac{1}{(ikR_2)^2}\right)(1 - 3\sin^2\psi')2\frac{e^{ikR_2}}{R_2}\\ &+ u^2(1 - u^2\cos^2\psi')V\Bigg].\end{aligned} \tag{47}$$

Writing R' in the approximate form (34), substituting in (45), and retaining only first order terms in $1/R_2$, we obtain

$$V = 2\frac{e^{ikR_2}}{R_2}\left[1 + ik_2\sqrt{1 - u^2\cos^2\psi'}\int_0^\infty e^{ik[(a+z)s/R_2u^2 + s^2/2R_2u^4 + s\sqrt{1-u^2\cos^2\psi'}/u]}ds\right]. \tag{48}$$

Using the identity (36), we obtain for V

$$V = [(1 - R_v)F + 1 + R_v]\frac{e^{ikR_2}}{R_2} \tag{49}$$

where,

$$R_v \equiv \frac{\sin\psi' - u\sqrt{1 - u^2\cos^2\psi'}}{\sin\psi' + u\sqrt{1 - u^2\cos^2\psi'}}$$

$$F \equiv [1 + i\sqrt{\pi w}\,e^{-w}\,erfc(-i\sqrt{w})] \tag{50}$$

$$w = p_1\left[1 + (z + a)/R_2 u\sqrt{1 - u^2 \cos^2\psi'}\right]^2 = 4p_1/(1 - R_v)^2 \quad (51)$$

$$p_1 = ikR_2u^2(1 - u^2 \cos^2\psi')/2 \equiv pe^{ib} \quad (52)$$

$$p \cong \frac{\pi}{x}\frac{R_2}{\lambda}\cos b \cong \frac{\pi}{\epsilon + \cos^2\psi'}\frac{R_2}{\lambda}\sin b \quad (53)^7$$

$$\tan b \cong (\epsilon + \cos^2\psi')/x. \quad (54)$$

The above equations for p, the "numerical distance," and b are accurate to the second order in u^2. The function F was discussed in Part I of this paper, ascending and descending series expansions were given, and its absolute value was shown graphically and in a table.

We may now perform the indicated differentiations in (9) and obtain for a vertical electric dipole

$$\begin{aligned} E_z^v = ik\Bigg\{&\cos^2\psi''\frac{e^{ikR_1}}{R_1} + R_v\cos^2\psi'\frac{e^{ikR_2}}{R_2} \\ &+ (1 - R_v)(1 - u^2 + u^4\cos^2\psi')F\frac{e^{ikR_2}}{R_2} \\ &- u\sqrt{1 - u^2\cos^2\psi'}\sin\psi' 2\frac{e^{ikR_2}}{ikR_2^2} \\ &- \frac{e^{ikR_1}}{R_1}\left(\frac{1}{ikR_1} - \frac{1}{(ikR_1)^2}\right)(1 - 3\sin^2\psi'') \\ &- \frac{e^{ikR_2}}{R_2}\left(\frac{1}{ikR_2} - \frac{1}{(ikR_2)^2}\right)(1 - 3\sin^2\psi')\Bigg\}e^{-i\omega t} \end{aligned} \quad (55)$$

$$\sin\psi'' \equiv (z - a)/R_1. \quad (56)$$

R_v is the coefficient of reflection for a plane wave with its electric vector in the plane of incidence. Thus the first two terms of (55) are just what one would get by applying the reciprocal theorem to two dipoles, one near the earth and the other far away. These terms are of the first order in $1/R$; higher order terms are contained in F. The last five terms in (55) correspond to the induction and electrostatic fields of the dipole and its image. Since $R_v = -1$ when $a + z = 0$, we obtain from (55) for the ground wave from a dipole near the earth

$$E_z^v = 2ik\left[F - u^2(1 - u^2)F - \frac{1}{ikr} + \frac{1}{(ikr)^2}\right]\frac{e^{i(kr-\omega t)}}{r} \quad (57)$$

and along the ground w becomes very nearly the same as the param-

[7] $x = 1.8\cdot 10^{18}\sigma_{emu}/f_{kc}$ and is the same as in Part I where it is shown graphically as a function of σ and f.

eter p_1 used in Part I; e.g., along the ground, (53) and (54) of this paper are identically the same as (5) and (6) of Part I. The first term in (57) is the same ground-wave attenuation function as given in (3), Part I, and has been derived in an entirely independent manner. Using the first term in the asymptotic expansion for F, we see that the second and third terms in (57) just cancel, showing that the use of F as the attenuation function involves neglecting terms of the order $1/R_2^3$. In fact, for distances greater than a wave length the difference between

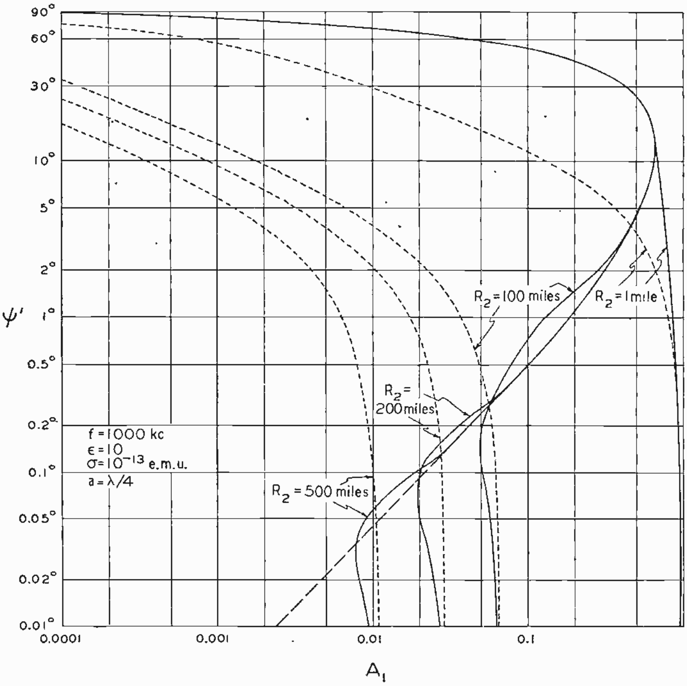

Fig. 2—Variation of attenuation factor A_1 with angle ψ' for a vertical electric dipole ($a=\lambda/4$, 1000 kilocycles, high conductivity).
——————— corresponds to vertical electric field.
– – – – – – first order sky-wave terms only.
-------------- higher order ground-wave term only.

$|F|$ and the absolute value of the square-bracketed quantity in (57) is negligible, thus justifying the use of the graphs of Part I for the attenuation factor. The more general ground-wave formula (57) need be used only for distances less than a wave length.

In order to demonstrate the relative importance of the various terms in (55) as a function of the angle ψ' and for various distances and frequencies, two graphs were prepared showing the variation of attenuation factor A_1 with angle ψ' where $A_1 \equiv |E_z{}^v R e^{-ikR}/2ik|$. Fig. 2

is for 1000 kilocycles, $\sigma=10^{-13}$ electromagnetic units, and $\epsilon=10$. The dipole is at a height $\lambda/4$ above the ground. The long dashed curve is for the first order terms in (55) and is the same for any distance, the dotted curves are for the third term only; i.e., the ground wave when $z=0$, and the solid curves represent the absolute value of the sum of all three terms. It is evident that only the first order terms in (55) are required for computing the field at large angles and at any distance.

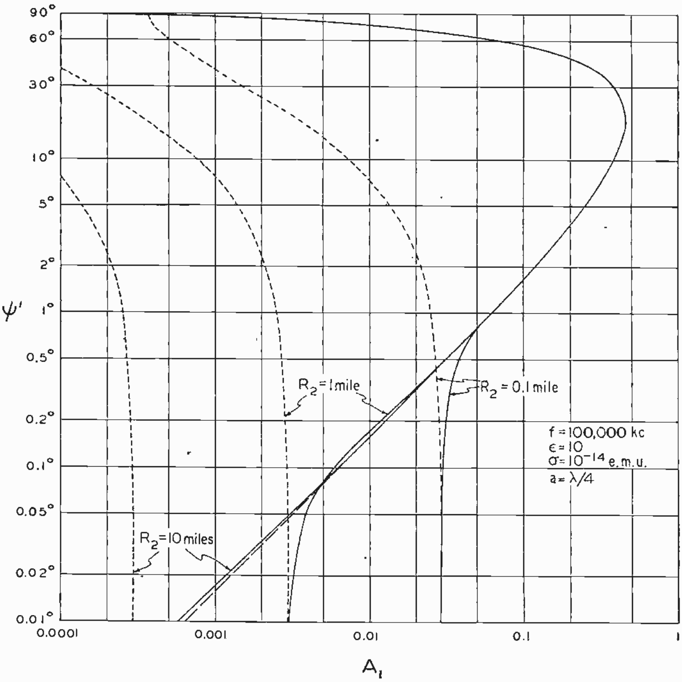

Fig. 3—Variation of attenuation factor A_1 with angle ψ' for a vertical electric dipole ($a=\lambda/4$, 100,000 kilocycles, low conductivity).
——————— corresponds to vertical electric field.
– – – – – – first order sky-wave terms only.
-------------- higher order ground-wave term only.

For small angles and short distances, the third term must be used together with the first order terms. As the distance is increased, the first order terms are sufficient for the specification of the field down to a very small angle which is inversely proportional to the distance. This latter property of the first order terms is very useful in computing the sky-wave radiation from antennas. In this case we are interested in the total field at the ionosphere where the waves are reflected or refracted back to earth. At high angles, the distance to the ionosphere is of the order of 100 miles, but at high angles the first order terms may be used at very short distances; at lower angles, the distance to the ionosphere becomes inversely proportional to the angle (neglecting the curvature of the earth) in just the manner required to make possible the use of the first order terms for computing the sky-wave radiation.

Fig. 3 is similar to Fig. 2 except that $f = 100{,}000$ kilocycles, $\sigma = 10^{-14}$ electromagnetic units, and $\epsilon = 10$. It is evident that the third term is much smaller, at all angles, than the corresponding term in Fig. 2.

In order to complete our formulation, equations are required for H. These may be derived in a manner similar to that used for obtaining the above formulas for V. The details are not interesting; the results are as follows:

$$H = [(1 + R_h)G + 1 - R_h]\frac{e^{ikR_2}}{R_2} \tag{58}$$

where,

$$R_h \equiv \frac{\sqrt{1 - u^2\cos^2\psi'} - u\sin\psi'}{\sqrt{1 - u^2\cos^2\psi'} + u\sin\psi'} \tag{59}$$

$$G \equiv [1 + i\sqrt{\pi v}e^{-v}erfc(-i\sqrt{v})] \tag{60}$$

$$v \equiv q_1[1 + (a + z)u/R_2\sqrt{1 - u^2\cos^2\psi'}]^2 = 4q_1/(1 + R_h)^2 \tag{61}$$

$$q_1 = ikR_2(1 - u^2\cos^2\psi')/2u^2 \equiv -qe^{-ib'} \tag{62}$$

$$q \cong \frac{\pi x}{\cos b'}\,\frac{R_2}{\lambda} \cong \frac{\pi(\epsilon - \cos^2\psi')}{\sin b'}\,\frac{R_2}{\lambda} \tag{63}$$

$$\tan b' \cong (\epsilon - \cos^2\psi')/x. \tag{64}$$

R_h is the coefficient of reflection for a plane wave with its electric vector perpendicular to the plane of incidence.

We also require the partial derivatives of V and H with respect to r and z

$$\frac{\partial V}{\partial r} = ik\cos\psi'\left\{\left[1 - \frac{1}{ikR_2}\right]V - (1 - R_v)\frac{e^{ikR_2}}{2ikR_2{}^2}\right.$$
$$-\frac{(1 - R_v)F}{2}\left[u^2(1 - u^2\cos^2\psi')\right.$$
$$\left.\left. - \sin^2\psi' - \frac{1}{ikR_2}\right]\frac{e^{ikR_2}}{R_2}\right\}. \tag{65}$$

$$\frac{\partial^2 V}{\partial r^2} = -k^2\left\{\left[\cos^2\psi' + \left(\frac{1}{ikR_2} - \frac{1}{(ikR_2)^2}\right)(1 - 3\cos^2\psi')\right]V\right.$$
$$-\cos^2\psi'(1 - R_v)\left(1 - \frac{1}{ikR_2}\right)\left(F\left[u^2(1 - u^2\cos^2\psi')\right.\right.$$

$$- \sin^2 \psi' - \frac{1}{ikR_2}\bigg] + \frac{1}{ikR_2}\bigg)\frac{e^{ikR_2}}{R_2} + \text{higher order terms}\bigg\}. \quad (66)$$

$$\frac{\partial^2 V}{\partial r \partial z} = - k^2 \sin \psi' \cos \psi' 2 \frac{e^{ikR_2}}{R_2}\left\{1 - \frac{3}{ikR_2} + \frac{3}{(ikR_2)^2}\right\}$$

$$- iku\sqrt{1 - u^2 \cos^2 \psi'}\, \frac{\partial V}{\partial r}. \quad (67)$$

$$\frac{\partial H}{\partial z} = ik\left[\sin \psi' 2 \frac{e^{ikR_2}}{R_2}\left(1 - \frac{1}{ikR_2}\right) - \frac{\sqrt{1 - u^2 \cos^2 \psi'}}{u} H\right]. \quad (68)$$

$$\frac{\partial H}{\partial r} = ik \cos \psi' \left\{\left[1 - \frac{1}{ikR_2}\right] H - (1 + R_h)\frac{e^{ikR_2}}{2ikR_2}\right.$$

$$- \frac{(1 + R_h)}{2} G\left[\frac{(1 - u^2 \cos^2 \psi')}{u^2}\right.$$

$$\left. - \sin^2 \psi' - \frac{1}{ikR_2}\right]\frac{e^{ikR_2}}{R_2}\bigg\}. \quad (69)$$

$$E_r^v = - ik\left\{\sin \psi'' \cos \psi'' \frac{e^{ikR_1}}{R_1} + R_v \sin \psi' \cos \psi' \frac{e^{ikR_2}}{R_2}\right.$$

$$- \cos \psi'(1 - R_v) u \sqrt{1 - u^2 \cos^2 \psi'}\, F \frac{e^{ikR_2}}{R_2}$$

$$\cdot\left(1 - \frac{u^2(1 - u^2 \cos^2 \psi')}{2} + \frac{\sin^2 \psi'}{2} - \frac{1}{2ikR_2}\right)$$

$$+ \sin \psi' \cos \psi'(1 - R_v)\frac{e^{ikR_2}}{ikR_2^2}$$

$$- 3 \sin \psi'' \cos \psi''\left(\frac{1}{ikR_1} - \frac{1}{(ikR_1)^2}\right)\frac{e^{ikR_1}}{R_1}$$

$$+ \cos \psi' u \sqrt{1 - u^2 \cos^2 \psi'}(1 - R_v)\frac{e^{ikR_2}}{2ikR_2^2}$$

$$\left. - 3 \sin \psi' \cos \psi'\left(\frac{1}{ikR_2} - \frac{1}{(ikR_2)^2}\right)\frac{e^{ikR_2}}{R_2}\right\} e^{-i\omega t}. \quad (70)$$

$$E_\phi^v = 0. \quad (71)$$

$$E_z^h = ik \cos \phi \left\{\sin \psi'' \cos \psi'' \frac{e^{ikR_1}}{R_1} - R_v \sin \psi' \cos \psi' \frac{e^{ikR_2}}{R_2}\right.$$

$$+ \cos \psi'(1 - R_v) u \sqrt{1 - u^2 \cos^2 \psi'}\, F \frac{e^{ikR_2}}{R_2}$$

$$\cdot\left(1 - \frac{u^2(1 - u^2\cos^2\psi')}{2} + \frac{\sin^2\psi'}{2} - \frac{1}{2ikR_2}\right)$$

$$- \sin\psi'\cos\psi'(1 - R_v)\frac{e^{ikR_2}}{ikR_2{}^2}$$

$$- 3\sin\psi''\cos\psi''\left(\frac{1}{ikR_1} - \frac{1}{(ikR_1)^2}\right)\frac{e^{ikR_1}}{R_1}$$

$$- \cos\psi'(1 - R_v)u\sqrt{1 - u^2\cos^2\psi'}\,\frac{e^{ikR_2}}{2ikR_2{}^2}$$

$$+ 3\sin\psi'\cos\psi'\left(\frac{1}{ikR_2} - \frac{1}{(ikR_2)^2}\right)\frac{e^{ikR_2}}{R_2}\Bigg\} e^{-i\omega t}. \tag{72}$$

$$E_r{}^h = -ik\cos\phi\Bigg\{\sin^2\psi''\frac{e^{ikR_1}}{R_1} - R_v\sin^2\psi'\frac{e^{ikR_2}}{R_2} + (1 + R_h)G\frac{e^{ikR_2}}{R_2}$$

$$- \cos^2\psi' u^2(1 - R_v)F\frac{e^{ikR_2}}{R_2}$$

$$- \left(\frac{1}{ikR_1} - \frac{1}{(ikR_1)^2}\right)(1 - 3\cos^2\psi'')\frac{e^{ikR_1}}{R_1}$$

$$+ \left(\frac{1}{ikR_2} - \frac{1}{(ikR_2)^2}\right)(1 - 3\cos^2\psi')[1 - u^2(1 + R_v)$$

$$- u^2(1 - R_v)F]\frac{e^{ikR_2}}{R_2}$$

$$+ u^2\cos^2\psi'(1 - R_v)\left(1 - \frac{1}{ikR_2}\right)\left(F\left[u^2(1 - u^2\cos^2\psi')\right.\right.$$

$$\left.\left.- \sin^2\psi' - \frac{1}{ikR_2}\right] + \frac{1}{ikR_2}\right)\frac{e^{ikR_2}}{R_2}\Bigg\} e^{-i\omega t}. \tag{73}$$

$$E_\phi{}^h = ik\sin\phi\Bigg\{\frac{e^{ikR_1}}{R_1} - R_h\frac{e^{ikR_2}}{R_2}$$

$$+ (R_h + 1)G\frac{e^{ikR_2}}{R_2} - \left(1 - \frac{1}{ikR_1}\right)\frac{e^{ikR_1}}{ikR_1{}^2}$$

$$+ \left(1 - \frac{1}{ikR_2}\right)[1 - u^2(1 + R_v) - u^2(1 - R_v)F]\frac{e^{ikR_2}}{ikR_2{}^2}$$

$$+ \frac{u^2(1 - R_v)}{2}\left(F\left[u^2(1 - u^2\cos^2\psi')\right.\right.$$

$$- \sin^2 \psi' - \frac{1}{ikR_2}\Bigg] + \frac{1}{ikR_2}\Bigg) \frac{e^{ikR_2}}{ikR_2^2}\Bigg\} e^{-i\omega t}. \tag{74}$$

$$E_z^{mh} = ik \cos \phi \Bigg\{ \cos \psi'' \frac{e^{ikR_1}}{R_1}\left(1 - \frac{1}{ikR_1}\right)$$

$$+ R_v \cos \psi' \frac{e^{ikR_2}}{R_2}\left(1 - \frac{1}{ikR_2}\right)$$

$$+ \cos \psi'(1 - R_v)F \frac{e^{ikR_2}}{R_2}\Bigg(1 - \frac{u^2(1 - u^2 \cos^2 \psi')}{2}$$

$$+ \frac{\sin^2 \psi'}{2} - \frac{1}{2ikR_2}\Bigg) - \cos \psi'(1 - R_v) \frac{e^{ikR_2}}{2ikR_2^2}\Bigg\} e^{-i\omega t}. \tag{75}$$

$$E_r^{mh} = - ik \cos \phi \Bigg\{ \sin \psi'' \frac{e^{ikR_1}}{R_1} + R_v \sin \psi' \frac{e^{ikR_2}}{R_2}$$

$$- u\sqrt{1 - u^2 \cos^2 \psi'}(1 - R_v)F \frac{e^{ikR_2}}{R_2}$$

$$- \sin \psi'' \frac{e^{ikR_1}}{ikR_1^2} - \sin \psi' \frac{e^{ikR_2}}{ikR_2^2}\Bigg\} e^{-i\omega t}. \tag{76}$$

$$E_\phi^{mh} = ik \sin \phi \Bigg\{ \sin \psi'' \frac{e^{ikR_1}}{R_1} + R_h \sin \psi' \frac{e^{ikR_2}}{R_2}$$

$$+ \frac{\sqrt{1 - u^2 \cos^2 \psi'}}{u}(1 + R_h)G \frac{e^{ikR_2}}{R_2}$$

$$- \sin \psi'' \frac{e^{ikR_1}}{ikR_1^2} + \sin \psi' \frac{e^{ikR_2}}{ikR_2^2}\Bigg\} e^{-i\omega t}. \tag{77}$$

$$E_\phi^{mv} = ik \Bigg\{ \cos \psi'' \frac{e^{ikR_1}}{R_1}\left[1 - \frac{1}{ikR_1}\right] - R_h \cos \psi' \frac{e^{ikR_2}}{R_2}\left[1 - \frac{1}{ikR_2}\right]$$

$$+ \cos \psi'(1 + R_h)\left[1 - \frac{1}{ikR_2}\right] G \frac{e^{ikR_2}}{R_2}$$

$$- \frac{(1 + R_h)}{2} G \Bigg[\frac{(1 - u^2 \cos^2 \psi')}{u^2} - \sin^2 \psi'$$

$$- \frac{1}{ikR_2}\Bigg] \frac{e^{ikR_2}}{R_2} - (1 + R_h) \frac{e^{ikR_2}}{2ikR_2^2}\Bigg\} e^{-i\omega t} \tag{78}$$

and

$$G \equiv [1 + i\sqrt{\pi v}\, e^{-v}\, erfc(-i\sqrt{v})]. \tag{79}$$

Terms arising from the differentiation of $\sqrt{1-u^2\cos^2\psi'}$ were neglected. The higher order terms which were dropped in $\partial^2V/\partial r^2$ in order to save space are just

$$2\frac{e^{ikR_2}}{R_2}\frac{\partial^2}{\partial r^2}[1+i\sqrt{\pi}p_1e^{-w}\,erfc(-i\sqrt{w})];$$

these terms are also dropped in E_r^h where they are of the order u^2/R_2^3 and u^4/R_2^2 and thus affect the value of E_r^h slightly at distances less than a wave length. In deriving equations for E_r^{mh} and E_ϕ^{mh}, terms in $1/ikrR_2$ were dropped so that these equations may be used only when $r>\lambda$. Except for the above limitations and very minor approximations[8] in F and G the above formulas apply at any point in space. In each equation the first two terms represent what would result by applying the reciprocal theorem to two properly oriented dipoles, one near the earth and the other far away. E is expressed in electrostatic units, I in electromagnetic units, and R in centimeters. In order to obtain E in millivolts per meter, I in amperes, and d in miles, it is only necessary to replace I/R in the above formulas by 18.64 I (amperes)/d (miles).

6. The Ground-Wave Radiation from a Vertical Antenna

As an example of the method of adding together the fields from several dipoles to determine the field from an antenna, we shall consider a vertical antenna of height h with a sinusoidal current distribution

$$\mathrm{I} = \mathrm{k}I_L \sin\,(A + B - ka) \tag{80}$$

where $A=2\pi h/\lambda$, $\sin B\equiv I_h/I_L$, and I_h denotes the current at the top of the antenna flowing into a nonradiating top load while I_L denotes the loop current. Such antennas are principally used at medium and low frequencies for the production of strong fields near the surface of the ground. If the antenna has a flat top, it will tend to increase its radiation resistance and efficiency but will add little to the ground-wave radiation as may be seen in (72), (73), and (74), which indicate that the E_z, E_r, and E_ϕ components of the ground-wave radiation from the flat top will be of the order u, u^2, and u^4, respectively, times the E_z component of the radiation from a corresponding vertical portion of the antenna. Thus the equations to be derived may be used for approximately computing the ground-wave radiation from antennas with horizontal portions providing the current distribution is approximately equivalent to (80). By (14) the vector electric field from the antenna is given by

[8] These approximations are briefly discussed in the conclusion; they may introduce small errors at the ultra-high frequencies for distances less than one wave length.

$$\mathbf{E} = I_L \int_0^h [\mathbf{k}E_z{}^v + \mathbf{r}E_r{}^v] \sin (A + B - ka)da. \tag{81}$$

Considering only the vertical component and restricting the results to ground waves, i.e., to distances such that $r >> h+z$, we may write $\cos \psi'' = \cos \psi' = 1$, $R_1 = R_2 = r$, $R_1 - r = -az/r$, $R_2 - r = az/r$ and, neglecting the induction and electrostatic terms, we obtain

$$E_z = ikI_L \frac{e^{i(kr-\omega t)}}{r} \int_0^h \Big[e^{-ikaz/r}$$
$$+ e^{ikaz/r}\left\{1 + 2i\sqrt{\pi p_1}\, e^{-w}\, erfc(-i\sqrt{w})\right\}\Big] \sin (A + B - ka)da \tag{82}$$

where,

$$r \gg h + z.$$

Using the identity (36), we may write $erfc(-i\sqrt{w})$ in a form which may easily be integrated. Then with $b = z/r + s/ru^2$, the general term to be integrated in (82) may be expressed by

$$\int_0^h e^{ikba} \sin (A + B - ka)da \tag{83}$$

and this may be integrated by parts and is equal to

$$\frac{1}{k(1 - b^2)} \Big\{ [\cos B + ib \sin B]e^{iAb}$$
$$- [\cos (A + B) + ib \sin (A + B)]\Big\}. \tag{84}$$

The following identity is also used in effecting the integration of (82):

$$\int_0^\infty s e^{[a^2s^2+bs]}ds \equiv -\frac{1}{2a^2}[1 + i\sqrt{\pi b^2/4a^2}\, e^{-b^2/4a^2} erfc(-ib/2a)]. \tag{85}$$

Neglecting b^2 with respect to unity and dropping higher order terms in $1/r$, we obtain

$$E_z = 2iI_L \frac{e^{i(kr-\omega t)}}{r} \Big[\cos B\left\{\cos (Az/r) + i\sqrt{\pi p_1}\, e^{iAz/r - w_1} erfc(-i\sqrt{w_1})\right\}$$
$$- \cos (A + B)\left\{1 + i\sqrt{\pi p_1}\, e^{-w_2} erfc(-i\sqrt{w_2})\right\}$$
$$- i \sin B u\sqrt{1 - u^2}\, e^{iAz/r}\left\{1 + i\sqrt{\pi w_1}\, e^{-w_1} erfc(-i\sqrt{w_1})\right\}$$
$$+ i \sin (A + B)u\sqrt{1 - u^2}\left\{1 + i\sqrt{\pi w_2}\, e^{-w_2} erfc(-i\sqrt{w_2})\right\}\Big] \tag{86}$$

where,

$$r \gg h + z$$

$$w_1 = p_1[1 + (h + z)/ru\sqrt{1 - u^2}]^2 \tag{87}$$

$$w_2 = p_1[1 + z/ru\sqrt{1 - u^2}]^2. \tag{88}$$

It is convenient, whenever possible, to divide the expression for the ground-wave field intensity into two factors, an "inverse distance" factor, 37.28 kh_eI_L/d (see equation (1), Part I), and an "attenuation factor," A_1. We see by (86) that h_e and A_1 are inextricably tied together, kh_eA_1 being equal to the absolute value of the quantity between the square brackets in (86). However, at the surface of the earth ($z=0$) it is possible to separate h_e and A_1 in two cases.

Case I $\qquad (r \ll h/u\sqrt{1-u^2},\ z=0,\ r>\lambda).$

Since the limit as $r \to 0$ of $i\sqrt{\pi p_1}\ e^{-w_1}\ erfc(-i\sqrt{w_1}) = -iAu\sqrt{1-u^2}$, and $A_1 \to 1$,

$$kh_e = \left| \{\cos B - \cos(A+B) - iu\sqrt{1-u^2}[A\cos B + \sin B - \sin(A+B)]\} \right| \tag{89}$$

Case II $\qquad (p > 20,\ z=0).$

Using the asymptotic expansion given in (41), we obtain from (86)

$$kh_eA_1 = \left| \{\cos B - \cos(A+B) - iu\sqrt{1-u^2}[A\cos B + \sin B - \sin(A+B)\} \right| /2p. \tag{90}$$

When we note that for large values of p

$$A_1 = 1/2p \tag{91}$$

it is evident that the "effective height" in the case of large "numerical distances" is the same as in the case of short distances. At intermediate distances h_e and A_1 cannot be easily separated. However, for most practical purposes, (89) may be used to determine the effect of the ground constants on the "effective height" of a vertical antenna and equation (3), Part I, used for the "attenuation factor."

It is evident that the finite height of the antenna has little effect on the ground-wave attenuation factor. However, the "effective height" of the antenna is a function of the ground constants. In a recent paper, W. W. Hansen[9] states that the "effective height" of an antenna may be determined independently of the ground constants. However, his conclusion was drawn from an approximate expression for the field and his more exact expressions indicate no such independence.

7. The Ground-Wave Field Intensity at a Short Distance from a Quarter-Wave Antenna

Substituting $\sqrt{1000\ P/R}$ for I_L in the equation for the field intensity and using practical units, we obtain

[9] *Physics*, vol. 7, pp. 460–465; December, (1936).

$$E = \frac{37.28kh_e\sqrt{1000P/R}}{d} A_1 \tag{92}$$

where P is the power in the antenna expressed in kilowatts, E is expressed in millivolts per meter, and d in miles. The total antenna resistance consists of a radiation and a loss component $R = R_r + R_L$. Using (89) to determine the effective height of the quarter-wave antenna, (92) becomes

$$E = \frac{1179}{d}\sqrt{P/R}\,\left|\,(1 - iu\sqrt{1 - u^2}[\pi/2 - 1])\,\right| A_1. \tag{93}$$

The unabsorbed field intensity at one mile (which is often used as a measure of antenna efficiency) is obtained by setting $A_1 = 1$ and d = one mile. In a recent paper, W. W. Hansen and J. G. Beckerley[10] gave values of R_r for a quarter-wave antenna for $\epsilon = 7$ and various values of x. Since kh_e is a function of x and ϵ, it is possible to determine the expected unabsorbed field intensity at one mile as a function of x and ϵ. Further, since the attenuation at a given distance in wave lengths from the antenna is also a function of x and ϵ, it is possible to determine the actual value of $E \cdot d$ at a distance of, say, two wave lengths. These values are given in Table I for various values of x and R_L, together with a representative set of values of f and σ. Other values of f and σ corresponding to the given values of x may be obtained by consulting Fig. 2 in Part I of this paper.

It is evident from Table I that $E \cdot d$ varies throughout wide limits for various values of x and R_L. It would seem from these data that the measured value of $E \cdot d$ (at, say 2λ) would be a better measure of a sta-

TABLE I
MILLIVOLT MILES PER METER PER KILOWATT FOR A QUARTER-WAVE ANTENNA AS A FUNCTION OF x AND R_L
($\epsilon = 7$)

x		0	1	10	100	∞
f_{kc} (for $\sigma = 10^{-13}$ emu):		Greater than 1,800,000	180,000 kc	18,000 kc	1800 kc	0
σ emu (for $f = 1000$ kc):		Less than $5.55 \cdot 10^{-17}$	$5.55 \cdot 10^{-16}$	$5.55 \cdot 10^{-15}$	$5.55 \cdot 10^{-14}$	∞
kh_e		1.019	1.006	0.937	0.962	1
	R_r[10] =	16.8	16.3	16.8	23.8	36.6
$E \cdot d$ (for $A_1 = 1$)	$R_L = 0$	293	294	270	232	195
	$R_L = 2$	277	277	255	223	190
	$R_L = 5$	258	257	237	211	183
	$R_L = 10$	232	232	213	195	173
A_1 (at 2λ)		0.34	0.37	0.59	0.96	1
$E \cdot d$ (at 2λ):	$R_L = 0$	99	109	159	223	195
	$R_L = 2$	95	102	150	214	190
	$R_L = 5$	88	95	140	203	183
	$R_L = 10$	79	85	126	187	173

[10] PROC. I.R.E., vol. 24, pp. 1594–1621; December, (1936).

tion's antenna efficiency than the unabsorbed field intensity at one mile. It is evident also that a fairly low conductivity under the antenna has a tendency to increase the ground-wave radiation near the antenna. At large distances this beneficial effect is canceled by the greater attenuation. Since, in the broadcast band, x is of the order of 100, it is evident from the table that $E \cdot d$ near the antenna is theoretically larger than the values for a perfectly conducting earth.

8. The Ground-Wave Radiation from Elevated Half-Wave Antennas

In view of the practical importance of elevated antennas at the ultra-high frequencies, formulas will be derived for the ground-wave radiation from vertical and horizontal half-wave antennas with their mid-points at a height h_1 above the ground. In the case of the vertical antenna

$$\mathrm{I} = \mathrm{k} I_L \cos (H_1 - ka) \tag{94}$$

where,

$$H_1 = kh_1.$$

Making the same approximations as in section 6 and integrating in a similar manner, we obtain

$$E_z{}^v = 2iI_L \frac{e^{i(kr-\omega t)}}{r} \cos (\pi z/2r) \left[e^{-iH_1 z/r} + e^{iH_1 z/r}\left\{1 + 2i\sqrt{\pi p_1}\, e^{-w_1}\, erfc(-\, i\sqrt{w_1})\right\}\right] \tag{95}$$

where,

$$r \gg h_1 + z$$

$$w_1 = p_1[1 + (h_1 + z)/ru\sqrt{1 - u^2}]^2. \tag{96}$$

Using the asymptotic expansion for $erfc(-i\sqrt{w_1})$, e.g., see (41), we obtain for large values of w_1

$$E_z{}^v = 2iI_L \frac{e^{i(kr-\omega t)}}{r} \left[e^{-iH_1 z/r} + e^{iH_1 z/r}\left\{R_v - (1 - R_v)/2w_1\right\}\right] \tag{97}$$

where,

$$|w_1| > 20 \qquad r \gg (h_1 + z)$$

and (97) may be written in terms of the reflection coefficients

$$E_z{}^v = 2iI_L \frac{e^{i(kr-\omega t)}}{r} \left\{e^{-iH_1 z/r} + e^{iH_1 z/r}\left[R_v - \frac{(1 + R_v)^2(1 - R_v)r}{4ik(h_1 + z)^2}\right]\right\}. \tag{98}$$

Finally, when $\sin(H_1z/r) \gg (h_1+z)/r|u\sqrt{1-u^2}|$ and $|w_1| \gg 20$, we obtain

$$E_z^v = 4I_L \frac{e^{i(kr-\omega t)}}{r} \sin(H_1z/r). \tag{99}$$

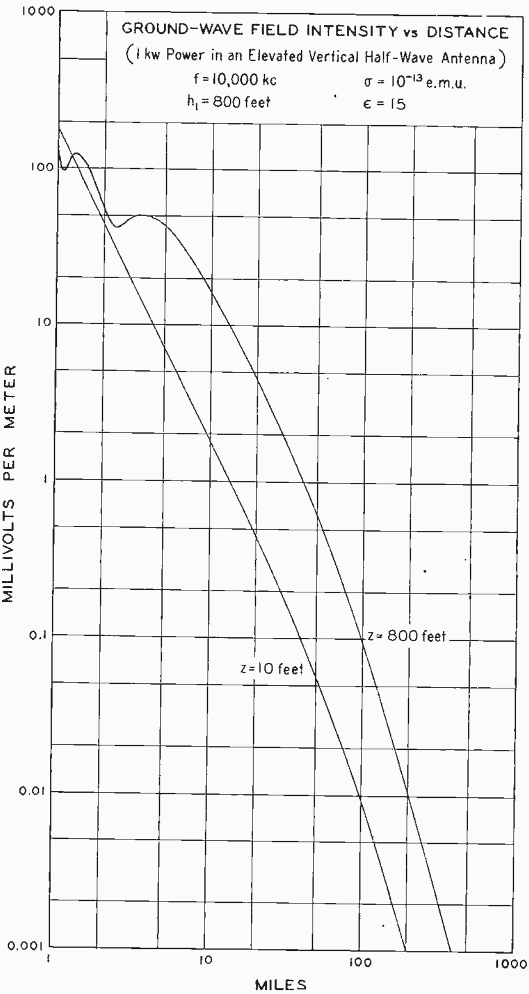

Fig. 4—Ground-wave field intensity vs. distance for an elevated vertical half-wave antenna (f = 10,000 kilocycles).

The above asymptotic ground-wave formula has been used by several investigators[11] to explain ultra-high-frequency propagation between elevated transmitting and receiving antennas.

Using (95), several transmission curves were prepared showing transmission from an elevated vertical half-wave antenna at a height 800 feet above the earth and with reception at 800 feet and ten feet above the earth. The antenna was assumed to have loss resistance of

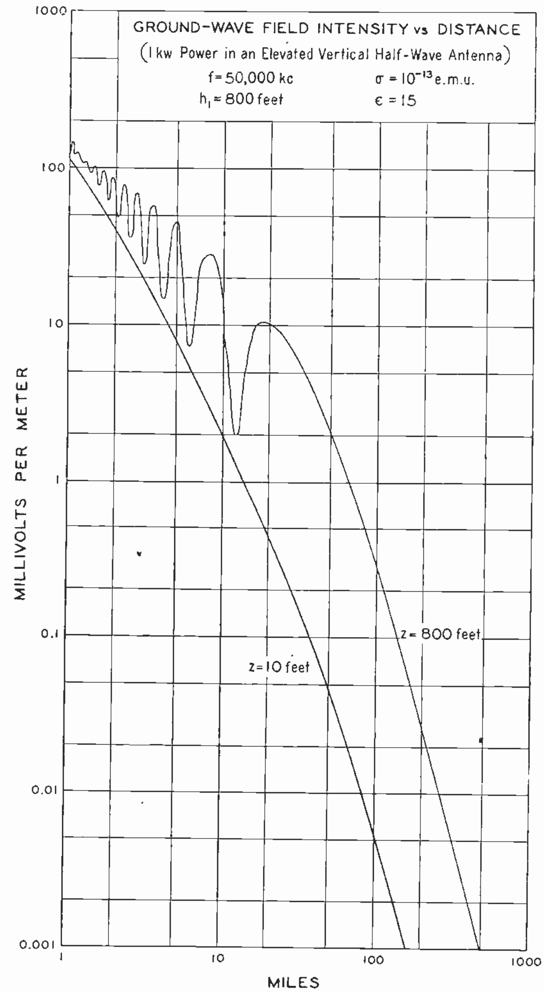

Fig. 5—Ground-wave field intensity vs. distance for an elevated vertical half-wave antenna ($f=50{,}000$ kilocycles).

two ohms and is fed with one kilowatt of power at 10,000, 50,000, and 100,000 kilocycles. The ground constants used are $\sigma=10^{-13}$ electromagnetic units and $\epsilon=15$. Beyond a distance of about ten miles, the curves were corrected for diffraction by means of the formula given in

Fig. 10 of the Burrows, Decino, and Hunt paper.[11] These curves are given in Figs. 4, 5, and 6. For the range of distances covered, i.e., from 1 to 500 miles, it was found that the asymptotic expansion (97) could be used for most of the calculations.

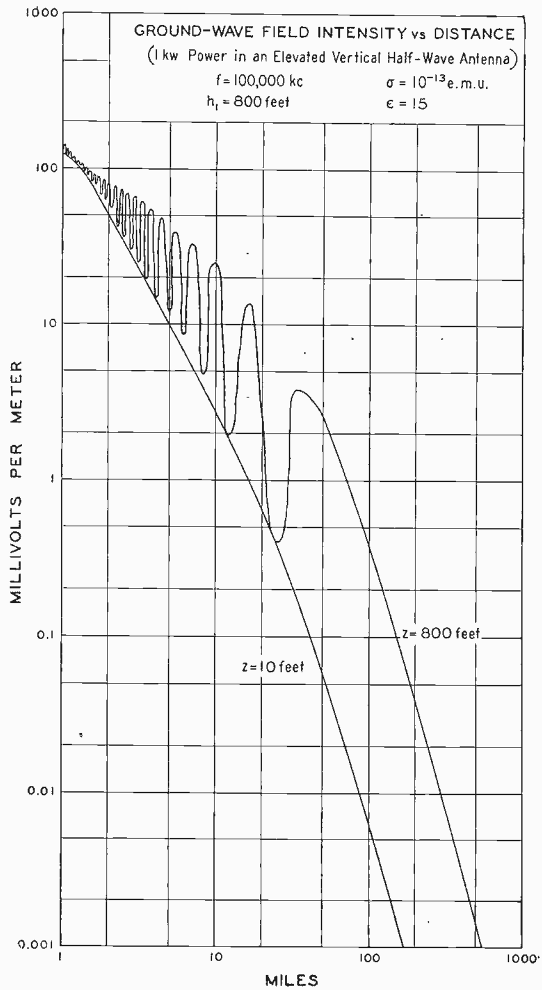

Fig. 6—Ground-wave field intensity vs. distance for an elevated vertical half-wave antenna ($f = 100{,}000$ kilocycles).

[11] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, Proc. I.R.E., vol. 21, pp. 427–463; March, (1933); C. R. Englund, A. B. Crawford, and W. W. Mumford, Proc., I.R.E., vol. 21, pp. 464–492; March, (1933); B. Trevor and P. S. Carter, Proc. I.R.E., vol. 21, pp. 387–426; March, (1933), and C. R. Burrows, A. Decino, and L. E. Hunt, Proc. I.R.E., vol. 23, pp. 1507–1535; December, (1935).

As another example of the propagation of the ultra-high frequencies from vertical antennas, Fig. 7 shows the attenuation factor versus distance (determined by equation (95)) for 150,000-kilocycle transmission over fresh water for which $\sigma = 5 \cdot 10^{-14}$ electromagnetic units and $\epsilon = 80$. Curves are given for $h_1 = z = 0$, 0.25λ, and 2.5λ above the earth. These results for varying antenna heights offer a plausible explanation of the discrepancy between experiment and theory as illustrated by Fig. 4 in Part I. It is evident from these graphs that the short antenna formula (equation (3), Part I) may be used without appreciable error when the transmitting and receiving antennas are less than a half wave length above the earth.

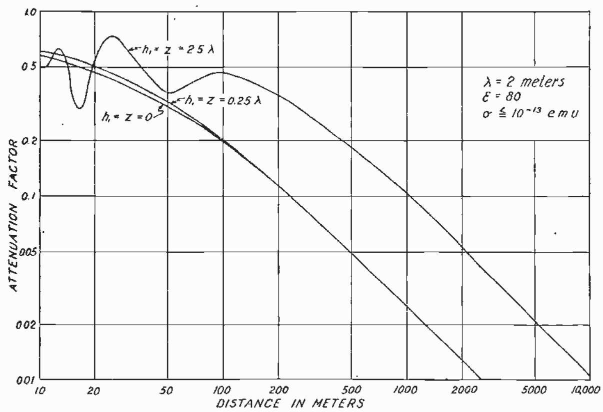

Fig. 7—Ultra-short-wave propagation over fresh water on 150 megacycles.

In the case of the half-wave horizontal antenna parallel to the x axis and at a height h_1, we have

$$\mathbf{I} = \mathbf{i} I_L \cos (kx). \tag{100}$$

Since the integration is from $x = -\lambda/4$ to $+\lambda/4$ and does not involve a, we may replace a by h_1 in (72), (73), and (74) and obtain, after making the same approximations as in section 6, the following expressions for the components:

$$E_z^h = 2iI_L \cos \phi \frac{e^{i(kr-\omega t)}}{r} \left[e^{-iH_1 z/r}(z - h_1)/r \right.$$
$$\left. + e^{iH_1 z/r}\left\{(z + h_1)/r + u\sqrt{1 - u^2}\, 2F\right\} \right] \tag{101}$$

$$E_r^h = 4iI_L \cos \phi \frac{e^{i(kr-\omega t)}}{r} \left[u^2 F - G \right] \tag{102}$$

$$E_\phi{}^h = 2iI_L \sin\phi \frac{e^{i(kr-\omega t)}}{r} [e^{-iH_1 z/r} + e^{iH_1 z/r}\{1 + 2i\sqrt{\pi q_1}\, e^{-v}\, erfc(-i\sqrt{v})\}] \quad (103)$$

where,

$$r \gg h_1 + z.$$

In evaluating F and G in the above formulas, a is to be replaced by h_1. For $E_\phi{}^h$ we have the following asymptotic formulas:

$$E_\phi{}^h = 2iI_L \sin\phi \frac{e^{i(kr-\omega t)}}{r} [e^{-iH_1 z/r} - e^{iH_1 z/r}\{R_h + (1 + R_h)/2v\}] \quad (104)$$

where,

$$|v| > 20 \quad \text{and} \quad r \gg h_1 + z$$

$$E_\phi{}^h = 2iI_L \sin\phi \frac{e^{i(kr-\omega t)}}{r} \left\{ e^{-iH_1 z/r} - e^{iH_1 z/r}\left[R_h + \frac{(1 - R_h)^2(1 + R_h)r}{4ik(h_1 + z)^2}\right]\right\}. \quad (105)$$

Finally, when $\sin(H_1 z/r) \gg (h_1+z)|u|/r|\sqrt{1-u^2}|$ and $|v| > 20$, we obtain

$$E_\phi{}^h = 4I_L \sin\phi \frac{e^{i(kr-\omega t)}}{r} \sin(H_1 z/r). \quad (106)$$

It is evident that in the limit for large enough distances the attenuation of the horizontal field from a horizontal antenna is no greater than the attenuation of the vertical field from a vertical antenna. The other components of $\mathbf{E}^h$ are usually of little practical importance.

9. The Forward Tilt and Polarization of The Electric Vector

The electric vector of the radiation from a vertical antenna lies in the vertical plane which passes through the antenna. Near the ground the electric vector is tilted forward and polarized in a manner which may be determined by taking the ratio of (55) and (70); e.g., when $(a+z) \ll R_2$ and $R_2 > \lambda$ we obtain

$$E_r{}^v/E_z{}^v = u\sqrt{1 - u^2}/\left[1 - \frac{u^2(1 - u^2)}{2} - \frac{1}{2ikR_2F} + \frac{z}{R_2 u\sqrt{1 - u^2}\,F}\right]. \quad (107)$$

It is evident that the tilt is independent of a, the height of the transmitting dipole, and will thus be the same for an antenna as for a dipole. At still greater distances, when $p>20$, (107) becomes

$$E_r^v/E_z^v = u\sqrt{1-u^2}/[1 - ikzu\sqrt{1-u^2}]. \qquad (108)$$

Comparing (107) and (108) we see that, along the ground, the wave assumes its tilt and polarization at a distance of the order of a wave length from the antenna and retains this form with little change for all greater distances. Thus the electric vector along the ground may be simply expressed by

$$\mathbf{E}^v = E_z^v[\mathbf{k} + u\sqrt{1-u^2}\,\mathbf{r}]. \qquad (109)$$

From (75) and (76) the same results are obtained for a wave originating in a loop antenna. In the case of a wave originating in a horizontal antenna, that part which lies in the vertical plane parallel to the direction of propagation also has this same property. This is most easily seen in (101) and (102). Thus the wave tilt is seen to be a general property of the electric vector in the vertical plane parallel to the direction of propagation. If we write $u\sqrt{1-u^2}=\alpha\, e^{-i\beta}$ and multiply the square bracket in (109) by the time factor $e^{-i\omega t}$, the real part of the resulting expression gives the equation of the ellipse

$$\mathbf{E}^v/E_z^v = \mathbf{k}\cos\omega t + \mathbf{r}\alpha\cos(\omega t + \beta). \qquad (110)$$

This vector reaches its maximum extension when $\omega t=-\delta$ and its minimum extension when $\omega t=\pi/2-\delta$ where

$$\tan\delta = \frac{1}{2\tau}\sqrt{1+4\tau^2} - \frac{1}{2\tau} = \tau - \tau^3 + 2\tau^5 - \cdots \qquad (111)$$

$$\tau = \alpha^2\cos\beta\sin\beta/[1 + \alpha^2\cos^2\beta - \alpha^2\sin^2\beta]. \qquad (112)$$

The measurable properties of the ellipse are θ, the forward tilt of the major axis and K, the ratio of the short to the long axis

$$\tan\theta = \alpha[\cos\beta + \sin\beta\tan\delta] \qquad (113)$$

$$K = \tan\delta\cot\theta. \qquad (114)$$

Fig. 8 shows θ and K as a function of x for $\epsilon=5$, 10, 20, and 80. It is evident that the maximum tilt which can be encountered in practice is less than about 22 degrees and, at broadcast frequencies, less than 15 degrees. Various investigators[12] have used the properties of the ellipse as a function of frequency for determining the ground constants. It is evident that measurements of both θ and K are required for a deter-

[12] See, e.g., C. B. Feldman, Proc. I.R.E., vol. 21, pp. 764–801; June, (1933).

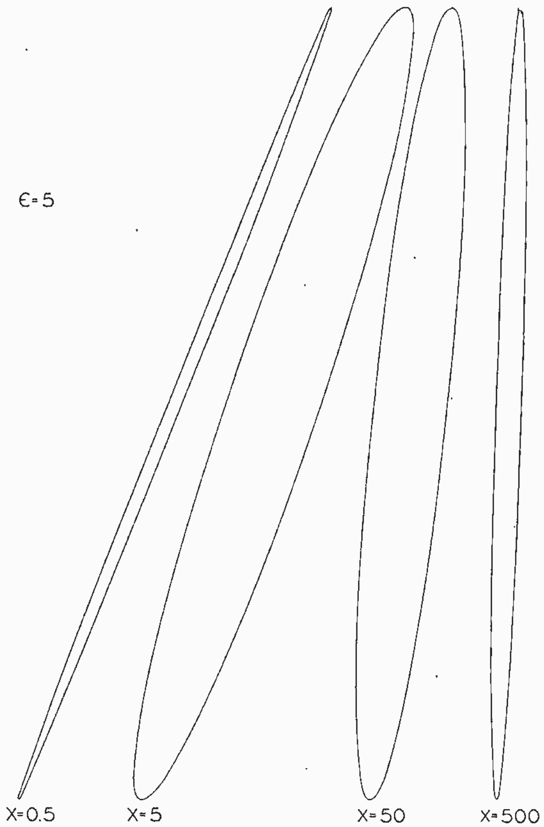

Fig. 9—The polarization of the electric vector at the surface of the earth.

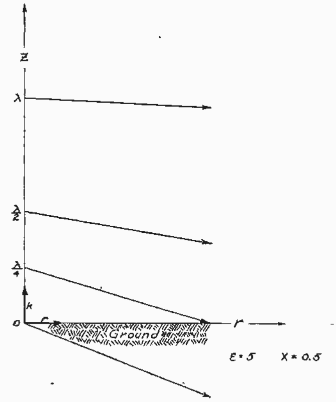

Fig. 10—The directions of the Poynting vector near the surface of the earth.

mination of both ϵ and σ. Fig. 9 was drawn to show the nature of the path traced by the electric vector during each cycle of the radio wave. The ellipses were drawn for $\epsilon=5$ and $x=0.5$, 5, 50, and 500. The largest tilt is obtained when $x=0.5$ corresponding to an ultra-high frequency. An average case in the broadcast band is shown for $x=50$. The case of $x=500$ corresponds to a low frequency for which the tilt has become very small.

10. The Poynting Vector

In order to obtain a clearer physical insight into the attenuation of ground waves, the Poynting vector will be determined along the ground in the case of propagation from a vertical dipole.

$$\mathbf{S} = \frac{c}{4\pi}\mathbf{E}^v \times \mathbf{H}^v \tag{115}$$

$$\mathbf{H} = \nabla \times \Pi^v e^{-i\omega t} = -\boldsymbol{\phi}\frac{\partial \Pi_z^v}{\partial r}e^{-i\omega t} \tag{116}$$

$$\begin{aligned} \mathbf{H} = -ik\boldsymbol{\phi}\Bigg\{&\cos\psi''\frac{e^{ikR_1}}{R_1}\left(1-\frac{1}{ikR_1}\right) + R_v\cos\psi'\frac{e^{ikR_2}}{R_2}\left(1-\frac{1}{ikR_2}\right) \\ &+ (1-R_v)F\cos\psi'\frac{e^{ikR_2}}{R_2}\left(1-\frac{u^2(1-u^2\cos^2\psi')}{2}\right. \\ &\left.+\frac{\sin^2\psi'}{2}-\frac{1}{2ikR_2}\right) - \cos\psi'(1-R_v)\frac{e^{ikR_2}}{2ikR_2^2}\Bigg\}e^{-i\omega t}. \end{aligned} \tag{117}$$

Using (109) for $\mathbf{E}^v$ and (117) for $\mathbf{H}^v$, we obtain

$$\begin{aligned} \mathbf{S} &= -\frac{c}{4\pi}E_z^v H_\phi^v[\mathbf{k} + \mathbf{r}u\sqrt{1-u^2}] \times \boldsymbol{\phi} \\ &= \frac{c}{4\pi}E_z^v H_\phi^v[\mathbf{r} - \mathbf{k}u\sqrt{1-u^2}] \end{aligned} \tag{118}$$

where,

$$a + z = 0.$$

If we let $u\sqrt{1-u^2}=\alpha e^{-i\beta}$ and take the real parts of E_z^v and H_ϕ^v, (118) becomes

$$\mathbf{S} = \frac{c}{4\pi}|E_z^v||H_\phi^v|\,[\mathbf{r}\cos^2\omega t - \mathbf{k}\alpha\cos(\omega t+\beta)\cos(\omega t)]. \tag{119}$$

Equation (119) defines a vector which oscillates in an ellipse similar to that traced by the electric vector but rotated forward in space a little less than ninety degrees. It shows that the instantaneous flow of energy takes place in two spurts each period of the radio wave; the radial component of S is always positive while the vertical component changes its sign.

In order to determine the direction of the average flow of energy and the fractional part flowing into the ground, S may be integrated throughout a single period of the wave

$$\mathrm{S}_{\text{average}} = \frac{c}{8\pi} \left| E_z{}^v \right| \left| H_\phi{}^v \right| [\mathrm{r} - \mathrm{k}\alpha \cos \beta]. \tag{120}$$

Equation (120) shows that a fractional part (very nearly equal to sin θ) of the energy in the wave at the ground flows downward into the ground. Fig. 10 shows the vector direction of the average flow of energy for the ultra-high-frequency case $\epsilon=5$ and $x=0.5$ for several heights above ground. For sufficiently large heights above ground the energy all flows radially.

11. Conclusion

Formulas have been given for computing the vector electric field from vertical, horizontal, and loop antennas. Since they yield asymptotically the correct expressions for the first order sky-wave terms and second order ground-wave term, they will no doubt give a satisfactory solution to any practical problem requiring a knowledge of the intensity of radio waves as radiated from an antenna over a plane earth. Due to the approximations made in V, which were necessary to obtain F in closed form, the formulas will be subject to errors of a few per cent at the ultra-high frequencies when $R_2<\lambda$ and $|\epsilon+ix|<10$; in particular, the formulas do not apply at any distance in the limiting case when the earth is replaced by air.

The principal novelty in these formulas is that the groundwork is laid for computing each component of the vector electric field from any kind of antenna in terms of the reflection coefficients R_v and R_h, and the attenuation function F. No further simplification, except in special cases, seems to be possible. Thus, in order to simplify the numerical computations, a table of the real and imaginary components of F as a function of the complex argument w is necessary. It is hoped that such a table will be published in the near future. In the meantime, the asymptotic expansion for F may be used for the solution of many important practical problems.

A brief discussion, with illustrations, was given of the ground-wave fields near the surface of the earth. The sky-wave radiation will be discussed in Part III and applied to the problem of E layer sky-wave propagation, giving results in good agreement with recent experimental data in the frequency range 550 to 1500 kilocycles.

Definitions of the Symbols

$\mathbf{E}^v$, $\mathbf{E}^h$, $\mathbf{E}^{mh}$, and $\mathbf{E}^{mv}$ denote the vector electric field produced respectively by vertical and horizontal electric dipoles and by horizontal and vertical magnetic dipoles.

$\mathbf{\Pi}^v$ and $\mathbf{\Pi}^h$ denote the wave potentials of vertical and horizontal electric dipoles.

V denotes an integral term in the wave potential of the vertical electric dipole. Equation (3) is an exact definition; a useful approximation to V is (49) and its asymptotic expansion is (43).

H denotes an integral term in the wave potential of the horizontal electric dipole. Equation (5) is an exact definition, while a useful approximation is given in (58).

$$F \equiv [1 + i\sqrt{\pi w}\, e^{-w} erfc(- i\sqrt{w})]$$
$$G \equiv [1 + i\sqrt{\pi v}\, e^{-v}\, erfc(- i\sqrt{v})]$$

F and G are the ground-wave attenuation functions for vertical and horizontal electric dipoles.

$$erfc(- i\sqrt{w}) \equiv \frac{2}{\sqrt{\pi}} \int_{-i\sqrt{w}}^{\infty} e^{-x^2} dx \equiv i \frac{2}{\sqrt{\pi}} \int_{i\infty}^{\sqrt{w}} e^{x^2} dx$$
$$w \equiv p_1[1 + (z + a)/R_2 u\sqrt{1 - u^2 \cos^2 \psi'}]^2 = 4p_1/(1 - R_v)^2$$
$$p_1 \equiv ikR_2 u^2(1 - u^2 \cos^2 \psi')/2 \equiv pe^{ib}$$

p is the "numerical distance." Useful approximations to p and b are given in (53) and (54).

$$v \equiv q_1[1 + (a + z)u/R_2\sqrt{1 - u^2 \cos^2 \psi'}]^2 = 4q_1/(1 + R_h)^2$$
$$q_1 \equiv ikR_2(1 - u^2 \cos^2 \psi')/2u^2 \equiv - qe^{-ib'}.$$

Useful approximations to q and b' are given in (63) and (64).

$\omega \equiv 2\pi f$

$k \equiv 2\pi/\lambda$

c is the velocity of light

$u^2 \equiv 1/(\epsilon + ix)$

$x \equiv 1.8 \cdot 10^{18} \sigma_{\text{emu}}/f_{kc}$ (given graphically in Part I). λ is the wave length and is to be expressed in the same units as the other quantities with which it is associated and with the dimension length. ϵ is the dielectric constant of the ground referred to air as unity, σ is the conductivity of the ground in electromagnetic units, and f is the frequency in kilocycles.

$\sin \psi' \equiv (z+a)/R_2$

$\sin \psi'' \equiv (z-a)/R_1$

$$R_v \equiv \frac{\sin \psi' - u\sqrt{1 - u^2 \cos^2 \psi'}}{\sin \psi' + u\sqrt{1 - u^2 \cos^2 \psi'}}$$

$$R_h \equiv \frac{\sqrt{1 - u^2 \cos^2 \psi'} - u \sin \psi'}{\sqrt{1 - u^2 \cos^2 \psi'} + u \sin \psi'}$$

R_v and R_h are the coefficients of reflection of a plane wave with its electric vector respectively parallel and perpendicular to the plane of incidence and with angle of incidence $(\pi/2 - \psi')$.

$A = kh$

h = height of a grounded vertical antenna.

$\sin B \equiv I_h/I_L$

I_h and I_L are the current at the top and at the current loop of a top-loaded vertical antenna with a sinusoidal current distribution on the vertical portion.

$H_1 = kh_1$

h_1 = height of the midpoint of an elevated half-wave antenna.

$A_1 \equiv |E_z{}^v \; Re^{-ikR}/2k|$ and is the "attenuation factor" where $R^2 = r^2 + z^2$.

P is the input power to the antenna in kilowatts.

R_r and R_L are the radiation and loss resistances of an antenna at the point where P is measured.

d is the distance along the ground expressed in miles.

$\alpha e^{-i\beta} \equiv u\sqrt{1-u^2}$

δ and τ are defined in (111) and (112).

θ and K are defined in (113) and (114) and in Fig. 8.

$\mathbf{S}$ is the Poynting vector.

The unit vectors $\mathbf{i}$, $\mathbf{j}$, $\mathbf{k}$, $\mathbf{r}$, and $\boldsymbol{\phi}$, and R_1, R_2, a, r, ϕ, x, y, and z are adequately defined in Fig. 1.

Acknowledgment

The writer wishes to thank Dr. L. P. Wheeler and Mr. Raymond Asserson for their many helpful suggestions and their encouragement during the progress of this work.

NOTE: Since this part of the paper was written, four papers have appeared which deal with this general subject. In order to assist the reader in comparing these results, the following remarks are added: There will be certain differences in signs between this paper and some of these others since the author has used $e^{-i\omega t}$ for the time factor, following Sommerfeld, van der Pol, Niessen, and others. The paper by W. H. Wise, "The physical reality of Zenneck's surface wave," which appeared in the *Bell System Technical Journal,* vol. 16, pp. 35–44; January, (1937), and the paper by C. R. Burrows, "The surface wave in radio propagation over plane earth," which appeared in the PROC. I.R.E., vol. 25, pp. 219–229; February, (1937), deal with the error in sign which was made by Professor Sommerfeld in his original paper and was pointed out by the author several years ago in *Nature.* The above two papers show theoretically and experimentally that the correct expression for the wave potential of a vertical electric dipole (which was first obtained by H. Weyl and later by Professor A. Sommerfeld and Balth. van der Pol, and which was shown graphically by the author in Part I of this paper) does not contain a term in the asymptotic expansion which may be identified with the Zenneck surface wave. Burrows' experimental results, which were obtained under nearly ideal conditions, are striking evidence of the accuracy of the ground wave attenuation formula in Part I of this paper. Wise then shows, that vertically polarized plane waves at grazing incidence and ground waves generated by vertical electric dipoles are characterized by a forward tilt almost identical to that of the Zenneck surface wave. He obtains a neat expression for the tilt of a wave generated by a vertical electric dipole near the earth. This expression, which is exact, is applicable for large "numerical distances" and short distances above ground. In this paper equations (55) and (70) for a vertical electric dipole, (72) and (73) for a horizontal electric dipole, and (75) and (76) for a horizontal magnetic dipole have been used to determine the wave tilt of the electric vector lying in the vertical plane parallel to the direction of propagation and apply at any point in space. The equation obtained for the vertical electric dipole is in exact agreement with the expression given by Wise in the limiting case which he considered.

In the paper "Radio propagation over plane earth—field strength curves," *Bell System Technical Journal,* vol. 16, pp. 45–75; January, (1937), C. R. Burrows presents equations and curves for determining the vertical field at the surface of the earth for a vertical electric dipole near the earth. These curves are based upon the exact series expansions obtained by W. H. Wise ("The grounded condenser antenna radiation formula," PROC. I.R.E., vol. 19, pp. 1684–1689; September, (1931)). The curves agree with those given in Fig. 1, Part I, of this paper as closely as they can be read. He also gave approximate curves for the field at distances less than a wave length in the limiting ultra-high-frequency case where the earth is a nonconducting dielectric. Burrows has informed the author that the approximation in these latter curves has been eliminated in a paper which he expects to publish in the October, (1937), issue of the *Bell System Technical Journal.* His new results consist of a revision of equations (17) and (19) and Fig. 3 of the above paper and show that equation (57) of this paper gives correct average results for any frequency and for any value of the ground constants found in practice. Equation (57) is only deficient in that it does not reproduce the small ripples in the field intensity which occur in Burrows' revised curves at distances less than a few wave lengths in the limiting ultra-high-frequency case. These ripples are so small and occur at such short distances from the transmitting dipole that it does not seem likely that they will be observed in practice. Thus we see that the approximate methods used in the derivations in this paper are fully justified. The use of these methods was necessary in order to obtain in a simplified closed form results which would be applicable at any point in space, for all frequencies, and for any set of ground constants found in practice. Burrows also presents equations and curves for determining the field a short distance above the earth in the case of large "numerical distances." Here again his equations are based on series expansions due to Wise and thus provide an independent check on the corresponding equations (98) and (105) in this paper. The factor $(1-R_v)/2$ appearing in equation (98) does not appear in Burrows' equation (27); since it is nearly unity near the ground, it was probably dropped as being of little importance although it is an essential factor in the limiting case of a perfectly conducting earth.

In the paper, "Series for the wave function of a radiating dipole at the earth's surface," *Bell System Technical Journal*, vol. 16, pp. 101–109; January, (1937), S. O. Rice obtains series expansions for the wave function of a dipole and verifies some series expansions due to Wise.

In the paper, "Uber die wirkung eines vertikalen dipolsenders auf ebener erde in einem entfernungsbereich von der ordnung einer wellenlänge," *Ann. der Phys.*, vol. 28, pp. 209–224; January, (1937), K. F. Niessen discusses the ground wave field at distances of the order of a wave length. Graphical results are to be given in a later paper, and this will afford a comparison with the results of this paper.

Proceedings of the Institute of Radio Engineers
Volume 25, Number 9 *September, 1937*

BOOK REVIEWS

Television Technical Terms and Definitions, by E. J. G. Lewis. Sir Isaac Pitman & Sons, Ltd., London, England. Available from Pitman Publishing Corporation, 2 West 45th St., New York, N. Y. Price, $1.75. 95 pages and 13 illustrations.

This little book will be helpful to newcomers to the art of television who meet strange words in newspaper accounts and technical periodicals. However, very few of this list of about 1400 words will be new to the television engineer. The terms used in the latest cathode-ray television systems, as well as the older scanning disk systems, are found listed in this book.

The definitions are not of the type that engineers serving on I.R.E. or R.M.A. standardizing committees spend so many hours in formulating. Such "committee definitions" often attempt to be too explicit, too legal in form, too all-embracive. The author, Mr. Lewis, words his definitions more in the way he would chat with you. For example:

> "SYNCHRONIZING. One of the most, if not the most, important factors in television systems. The scanning at the receiver must operate at exactly the same time and speed as that at the transmitter. When the exploring spot starts on its hundredth line across the scene, the receiver light spot must commence to trace its hundredth line across the screen; when the last line of a frame is scanned and a fresh frame commenced, the receiver light spot must act accordingly. Synchronizing is generally established by the transmitter sending out periodic synchronizing impulses which keep the local scanning arrangements at the receiver in step with the transmitter scanning system." (Taken from "Television Technical Terms & Definitions.")
>
> "SYNCHRONIZING. Synchronizing of images is the maintaining of the time and space relations between the transmitted and reproduced pictures." (R.M.A. 1933 Definition.)

Although published in London, where commercial television is before the eyes of the public (and the television jargon in their ears) there are few instances where typically British terms are given in preference to the American; in fact, in several instances special American words are listed. Of course there is more similarity between television terms in the two countries than between radio terms. One common term in which, however, there is a difference is the process of suppressing the return trace in a reproduced cathode-ray television picture. Our author gives only the English version:

> "BLACK-OUT. . . . The suppression of the return lines of a cathode ray tube during the frame synchronizing period by the application of a large negative bias to the grid of the cathode ray tube."

In the U.S.A. we prefer the corresponding term "BLANK OUT."

Mr. Lewis is to be complimented on the high degree of accuracy he uses in explaining technical terms in a simple manner suitable for the nontechnical reader.

*A. F. Murray

* Philco Radio and Television Corporation, Philadelphia, Pa.

Television Optics, by L. M. Myers. Sir Isaac Pitman & Sons, Ltd., London, England. Available from Pitman Publishing Company, 2 West 45th St., New York, N. Y. Price $8.50. 338 pages, 214 illustrations, and 1 chart.

When an author writes a treatise on television optics his writings must deal principally with mechanico-optical television systems. The first two chapters deal with geometrical optics and photometry. The material is accurate and clearly explained; especially good are the portions dealing with lenses and their aberrations. However, regrettable looseness is shown in Figs. 45 and 46, which deal with solid angles. Here the impression is given that a plane area of one square foot subtends one steradian at any point one foot distant from that plane. To be accurate this area should be shown as a surface of a sphere having a radius of one foot.

The attempt to untangle photometric units is only partially successful. It is believed that the discussion of many units not in general use confuses rather than clarifies. In the table of photometric units, page 59, one of the most important units in the English system, the foot candle (or more correctly the lumen-per-square foot) unfortunately is not mentioned. Photometry of lens systems for both point and extended sources is briefly and clearly treated.

In the third chapter is given a very good treatise on transmission of light through crystal media. This forms a very complete introduction to the material that follows on the Kerr effect. The construction, testing, and operation of Kerr cells is fully described.

Mechanico-optical scanning systems, single and multispiral disks, mirror drums, and lens disks are discussed and their relative photometric efficiencies are derived. This portion of the book contains material which appears in collected form for the first time.

The final chapter "Electron Optics" deals with the photometry of cathode-ray tubes and electron camera tubes. Electromagnetic and electrostatic deflection is mentioned briefly. The theory of the Farnsworth type and the Iconoscope type of camera tube, with a comparison of their relative efficiencies, is discussed. There is a description of intermediate methods employed with electron receiving devices. About twenty pages are devoted to electrostatic and electromagnetic lens systems and their optical analogies. The book concludes with a good bibliography. Most of the references are to English and German publications.

Summarizing, more than one-third of the book is devoted to the Kerr cell and mechanical scanning. The modern television engineer dealing with cathode-ray systems is naturally more interested in the subject of Electron Optics. The studio engineer will be interested in the chapter on Photometry, while the receiving engineer, anticipating large screen television images, should also find interesting this chapter and the one on Geometrical Optics.

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* Philco Radio and Television Corporation, Philadelphia, Pa.

Electrical Characteristics of Power and Telephone Transmission Lines, by Ferris W. Norris, and Lloyd A. Bingham. The International Textbook Company, Scranton, Pa. 272 pages, 44 figures.

This is an introductory textbook designed to be used early in the student's engineering training. In the first third of the book formulas are developed for the primary constants of transmission lines. In the middle part the steady state equations are solved, the derived constants are introduced, and equivalent lumped constant networks are set up. In the remainder of the text are a chapter on power relations, four chapters on transmission and interference on telephone lines, and a final chapter on traveling waves and transient phenomena. There are also an appendix giving the constants of some representative lines, a second appendix giving some mathematical tables, and a very brief index.

This book is best studied under the guidance of a teacher. The authors seek to lead the student through as much of rigorous development as possible but confine themselves to qualitative explanation in the more difficult problems, such as that of skin effect. To the mature reader the dividing line may seem rather arbitrary. But in the hands of a competent instructor the rather lengthy discussion of the expansion of hyperbolic functions will be used to drive home a familiarity that few students acquire in their early mathematical work, while the use of the language of line integrals will be a means to interest the student in broadening his mathematical education.

Many readers will notice that shunt conductance, including the effects of corona, is dismissed as negligible, on page 44, while on page 165 corona loss is given as the controlling factor in the selection of wire size for the Boulder Dam-Los Angeles line which heads the list, in the appendix, of representative power lines. Others will object to the statement, on page 101, that a certain telephone line will be infinitely long if its length exceeds 33.8 miles—a statement which is accompanied by no warning, save a single phrase in the appendix, that even the intended meaning is true only at 1000 cycles.

Each chapter except the last is followed by a group of well-chosen problems, whose solution will do much to fix in the student's mind the information presented in the text.

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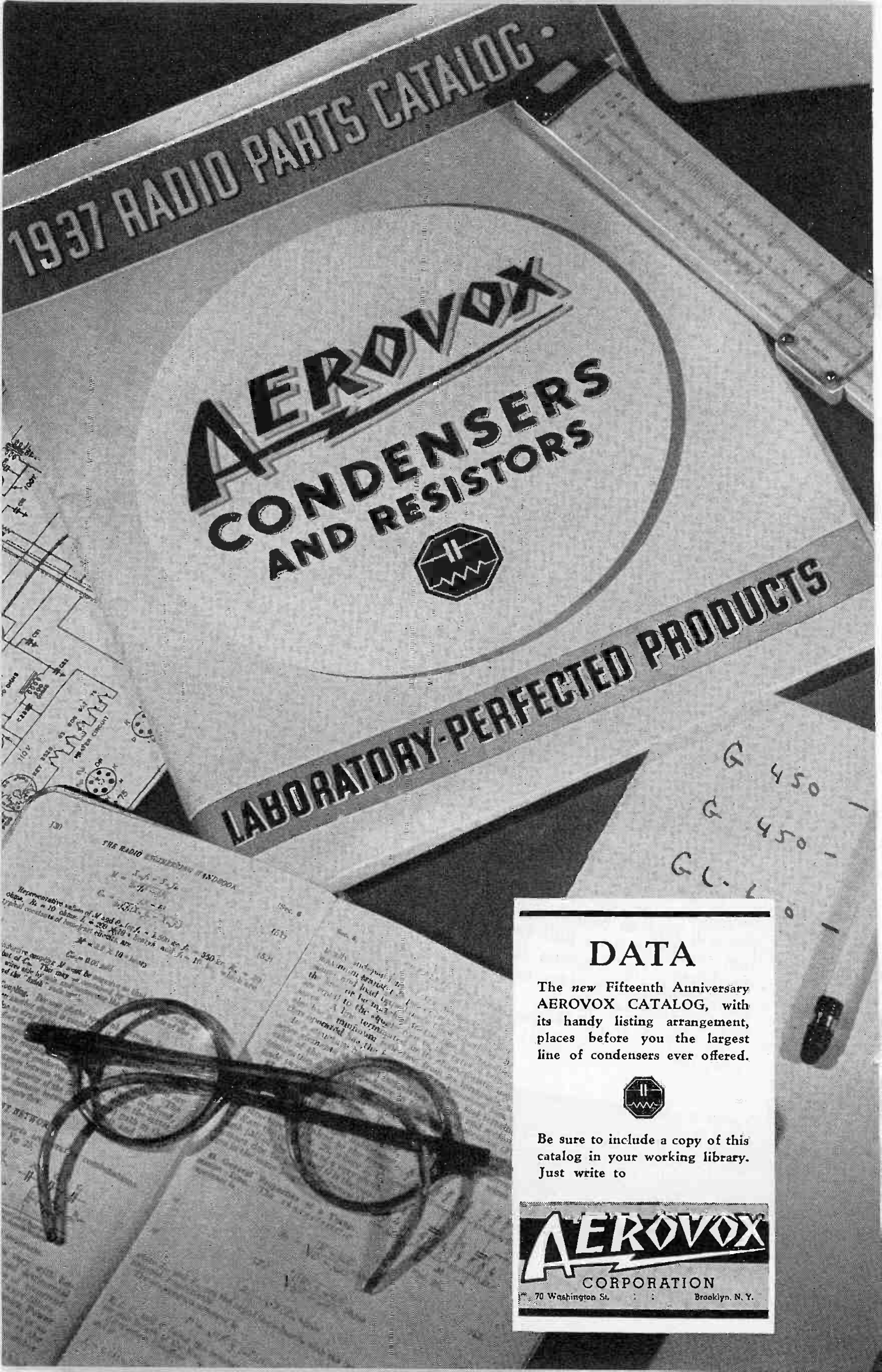
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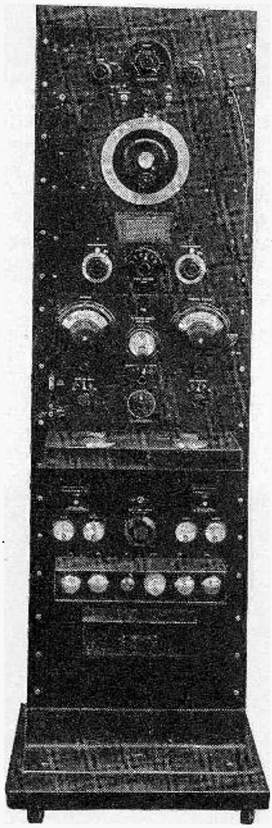

A COMPLETE rack-mounted a-c operated assembly comprising all apparatus required for measurement of the electroacoustical efficiency and frequency response characteristic of telephone receivers by means of the artificial ear. Comprises beat oscillator, (1 watt, 0-21,000 cycles), Type 502 Artificial Ear (1-100,000 bars), A-C voltmeter (.001-100 volts) and stabilized power supply. Adapted for both research and production testing. As supplied to the U. S. Army and the U. S. Navy.

SEND FOR BULLETIN 1B

Ballantine Laboratories, Inc.

Boonton, New Jersey

ELECTROACOUSTICAL INSTRUMENTS

When writing to advertisers mention of the PROCEEDINGS *will be mutually helpful.*

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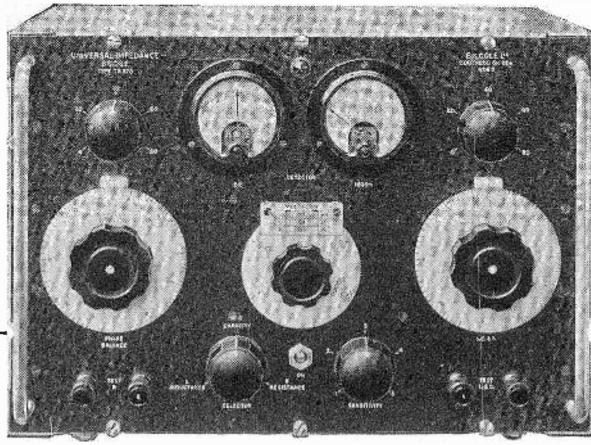

Universal Impedance Bridge. Type TF373.

An exceedingly useful general purpose instrument incorporating 1,000 cycle valve oscillator with a visual detector giving high discrimination for the measurement of:

CAPACITANCE

5 micro micro-farads to 100 micro-farads.
Condenser loss 0-1 ($R\omega C$).

INDUCTANCE

5 micro-henries to 100 henries.
Coil magnification 0-100 $\frac{\omega L}{R}$.

RESISTANCE

to 1 megohm using D.C.

Mains operated—no batteries or headphones required.

May we send prices and further details?

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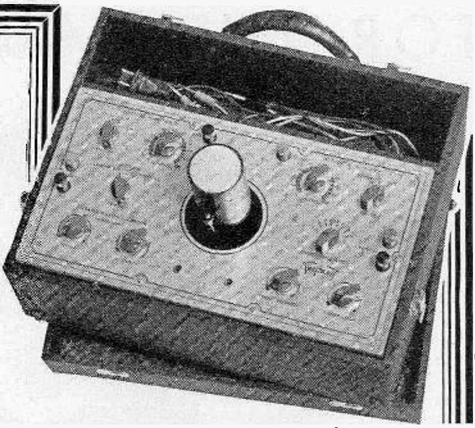

. . . A PRACTICAL OSCILLOSCOPE

- Exclusive Turret Type Tube Mounting for adjusting Screen to any angle for Direct View . . .
- Extremely Brilliant Pattern
- Linear Sweep 15 to 20,000 Cycles

Model 1690
In Metal Case
DEALER PRICE . . $47.00

- 2″ Cathode Ray Tube

MEETS EVERY REQUIREMENT OF SERVICING ENGINEERS FOR VISUAL STUDY AND ADJUSTMENT OF RADIO RECEIVER CIRCUIT PROBLEMS

ADVANTAGES

This new Triplett Oscilloscope is complete. In addition to radio servicing, it has invaluable applications in radio transmitter, sound equipment, industrial and educational fields.

The cathode ray tube with exclusive Triplett turret type mounting can be moved up or down or to either side for adjustment of the screen to an angle always in direct alignment with the user's line of vision. Adjustable shield can be pulled out to project beyond the screen for easy reading in brightest daylight.

The Oscilloscope incorporates separately controlled resistance coupled vertical and horizontal amplifiers. Vertical and horizontal plates can be either direct coupled with amplitude control, or through amplifiers with amplitude control. Can be used with any type frequency modulated Signal Generator. 60 Cycle A.C. operated.

Linear sweep from 15 to 20,000 cycles. Sweep (either internal, external or 60 cycle) may be used through amplifier or direct with amplitude control. Positive lock. All controls on panel. No interactional controls.

Model 1690—Contained in black leatherette case (as shown above) with removable cover. Built-in compartment for accessories.

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Model 1690—also furnished in black wrinkle finish metal case. (Same as above less black leather carrying case and cover)

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TRIPLETT Precision ELECTRICAL INSTRUMENTS

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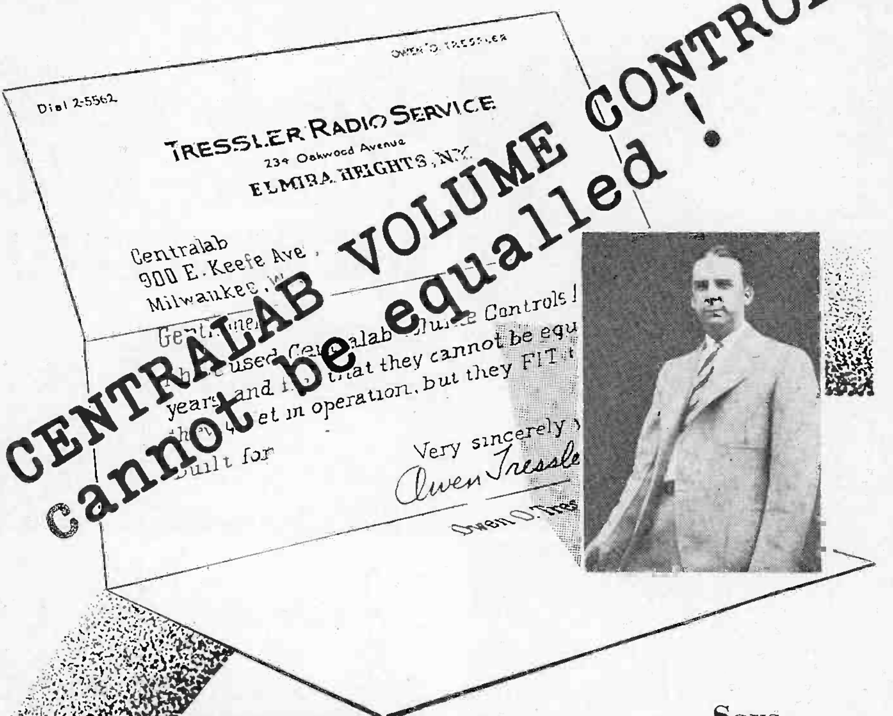
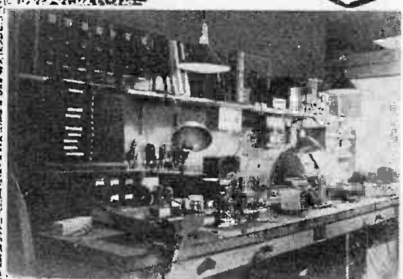
ALPHABETICAL INDEX TO ADVERTISEMENTS

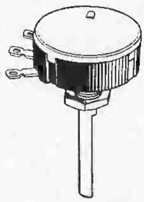
CENTRALAB VOLUME CONTROLS cannot be equalled !





A corner of Mr. Tresslers' service shop, located in Elmira Heights, a suburb of Elmira, N.Y. The largest service shop in Elmira—employs three service men. Mr. Tressler has a service background of thirteen years.

Says Mr. Owen O. Tressler



"I have used Centralab volume controls for the past three years, and FIND THEY CANNOT BE EQUALLED. Not only are they quiet in operation, but they FIT the sets they are built for."

Mr. Tressler is but one of the many service men throughout the land who prefer CENTRALAB controls for satisfactory replacements. Manufacturers, experimenters and set builders have in growing numbers changed to CENTRALAB for better performance. Specify "CENTRALAB."

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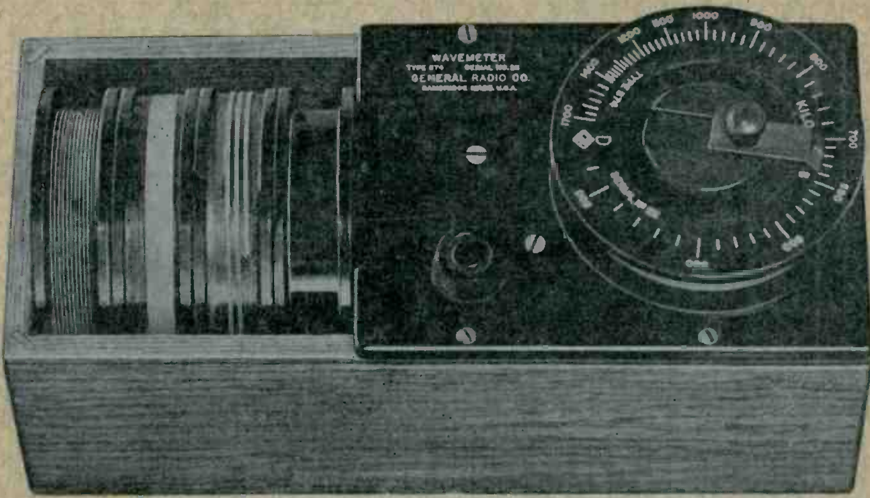

FINDING OUT WHERE YOU ARE

THE determination of the approximate frequency of the fundamental or harmonics of oscillators, transmitters, receivers and laboratory frequency-generating or detecting devices is often very difficult. Even rough guesses at the frequency from the physical dimensions of coils and condensers often lead to considerable confusion and doubt.

An extremely useful device for measurements of this type is a simple absorption wavemeter. The G-R Type 574 Wavemeter is being used widely in a large number of laboratories for just this kind of approximation.

This wavemeter features:

DIRECT READING—Scales individually engraved on each of the five plug-in coils.

WIDE RANGE—166 kc to 70,000 kc (18,000 m to 4.3 m)

PORTABLE—Can be held in one hand . . . weighs only 4½ lbs.

CONVENIENT TO USE—Spare coils held firmly in place in compartment . . . will not fall out.

GOOD ACCURACY—Within 3%.

INEXPENSIVE—Complete with all coils only $50.00.

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