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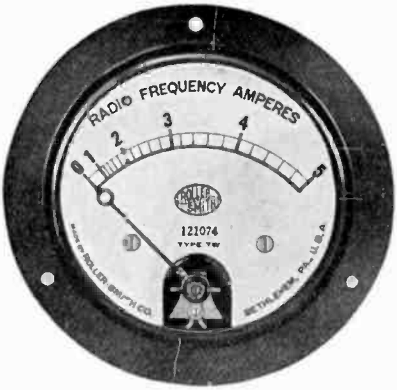
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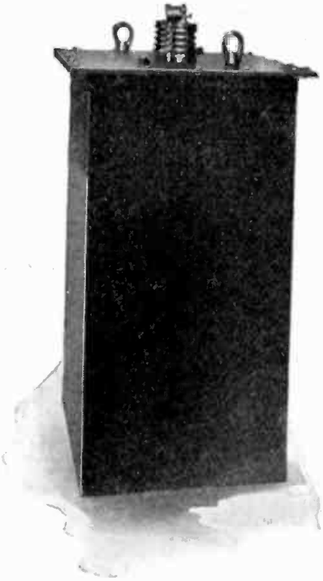

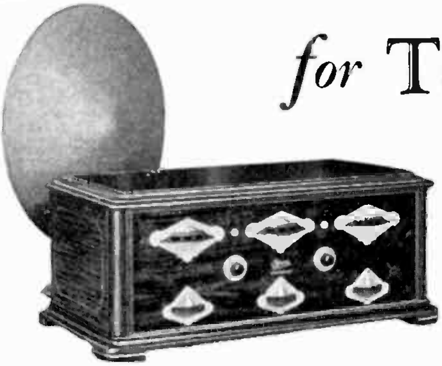


Doctor Wu

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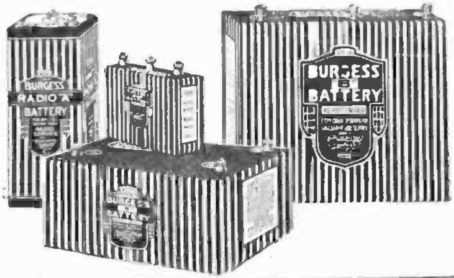

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PROCEEDINGS OF
The Institute of Radio Engineers
Volume 15 FEBRUARY, 1927 Number 2

CONTENTS

Page

Officers and Board of Direction 64
I. R. E. Sections 65
Committees, 1926 66
Institute Activities 67
Annual Convention 67
News of the Sections 69
Appointment of Assistant Secretary 69
Donald McNicol, "Opening Address at Annual Convention" . . 71
Ralph Bown, "Response of Incoming President" 79
G. W. Picard, "Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism" 83
W. A. MacDonald, "Importance of Laboratory Measurements in the Design of Radio Receivers" 99
E. L. Chaffee and G. H. Browning, "A Theoretical and Experimental Investigation of Detection for Small Signals" . . 113
Discussion on Austin Paper 155
John B. Brady, Digest of U. S. Radio Patents 159
Geographical List of New Members 163

GENERAL INFORMATION

The Proceedings of the Institute are published monthly and contain the papers and the discussions thereon as presented at meetings.

Payment of the annual dues by a member entitles him to one copy of each number of the Proceedings issued during the period of his membership.

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INSTITUTE ACTIVITIES

Annual Convention

For three days beginning January 10th, the Annual Convention of the Institute was held. The attendance at all of the sessions was very good. At the opening meeting on the 10th the new officers for 1927 were installed and the Liebmann Memoriay Prize for the year 1926 was presented to Dr. Ralph Bown. There followed a paper on "The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism" presented by Dr. G. W. Pickard. Much discussion followed. In the afternoon of this day many of the convention attendants visited the American Telephone and Telegraph Company's station on Walker Street. At 8 P. M. Dr. E. F. W. Alexanderson gave a talk, illustrated with apparatus set-ups, lantern slides, and moving-pictures on the General Electric Company's progress in transmission of photographs at a high rate of speed. The auditorium was packed to capacity at this meeting. On January 11th a paper by Professors Hugh A. Brown and Charles T. Knipp on "The Behavior of Alkali Vapor in Detector Tubes" was read. A series of informal discussions followed. In the afternoon trips were made to the Electrical Testing Laboratories, Dubilier Radio & Condenser Corporation, and the Federal Brandes Products Corporation. At 8 P. M. on this evening a paper by D. K. Martin on "Simultaneous Transmission of Broadcasts by Chain Stations on the Same Wave Lengths" was read by Mr. Martin. The last day of the convention, January 12, started off with four inspection trips; the Radio Corporation of America's transoceanic terminal apparatus station, the Driver-Harris Company, studio and control room of WEAF and the studio and control room of WJZ. In the afternoon a paper by W. A. McDonald was read. The subject of this paper was "The Importance of Laboratory Measurements in The Designs of Radio Receivers." The convention closed with the banquet at the Waldorf-Astoria where about three hundred members were present. Mr. D. Rigney acted as toast-master introducing one form of light entertainment after another until the session was officially brought to a close at midnight.

LUNCHEON FOR RETIRING PRESIDENT

On the first day of the Convention, January 10, a luncheon was given at the Fraternity Club to retiring president Donald McNicol. The affair was attended by six past-presidents and by officers who were in attendance at the first day's sessions. Those present were: Donald McNicol, Dr. M. I. Pupin, Dr. J. H. Dellinger, Dr. Ralph Bown, Dr. G. W. Pickard, J. V. L. Hogan, Lloyd Espenschied, Laurens E. Whittemore, Raymond A. Heising and R. H. Marriott.

JANUARY MEETING OF BOARD OF DIRECTION, JANUARY 5, 1927

At the meeting of the Board of Direction, held at Institute Headquarters on January 5th, the following were present: Donald McNicol, President; Dr. Ralph Bown, Vice-President; R. H. Marriott, J. V. L. Hogan, W. F. Hubley, and J. M. Clayton, Assistant Secretary. The members of the Board counted the ballots in the annual election of officers, the result being as follows: for President, Dr. Ralph Bown; for Vice-President, Frank Conrad; for Managers, R. A. Heising, and J. F. Dillon. Three additional managers are to be appointed at the February meeting of the Board to fill the terms of the three members leaving the Board.

The following were transferred to Member grade: Wilson Aull, and Clair D. Mitchell. The following were elected directly to the grade of Member: A. G. Lee, J. R. Land, Valerian Bashenoff, J. L. Bain, Jr. R. Martin. L. G. Pacent was transferred to the grade of Fellow.

One hundred and fifty-eight Associate members and 19 Junior members were elected. Treasurer Hubley submitted a statement of the disbursements for the year 1926.

SECTIONS COMMITTEE

The Sections Committee, D. H. Gage, Chairman, held a meeting at the Institute Headquarters on January 11th. A number of out-of-town Section Chairmen and representatives of different Sections were present. Many matters of importance in the organizing and managing of Sections were discussed. Much work was done toward standardizing the procedure to be followed in establishing a Section.

News of the Sections

LOS ANGELES SECTION

The Los Angeles Section held a meeting in the Commercial Club of Los Angeles on December 20th. Mr. Edward H. Gilford presented a talk on "Locating Unknown Metal Deposits by Radio, or Re-Radiation Applied to the Locating of Unknown Conductors." A general discussion followed. Forty-seven members attended the meeting which was preceded by a dinner.

CANADIAN SECTION

The Canadian Section held a meeting on the evening of January 5th at the Electrical Building, University of Toronto. Mr. Hepburn, Chairman, presided. Mr. V. G. Smith of the University of Toronto, presented a paper on "A Mathematical Study of Radio Frequency Amplification." This paper was discussed by C. I. Soucy, A. M. Patience and others. Due to the resignation of G. F. Eaton, Mr. Patience was elected Vice-Chairman. There were forty-six members present.

The next meeting of the Section will be on February 2nd in the Electrical Building of the University of Toronto at which time Mr. R. H. Langley will present a paper "Investigation Into Broadcasting Conditions."

Appointment of Assistant Secretary

The Board of Direction of the Institute has employed Mr. John M. Clayton as Assistant Secretary of the Institute. Mr. Clayton will be permanently located at headquarters, 37 West 39th Street, New York, and will have direct charge of office management, records, correspondence, and all matters pertaining to Institute activities.

Mr. Clayton has for some years been assistant technical editor of QST Magazine, of the American Radio Relay League, Hartford, Conn. He is well versed in radio engineering literature, is familiar with the business of radio organizations and has a wide acquaintance among radio engineers throughout the United States and Canada. He took up his duties with the Institute beginning January 1, 1927.

OPENING ADDRESS OF DONALD McNICOL RETIRING PRESIDENT

Annual Meeting and Convention, Institute of Radio Engineers, January 10, 1927.

This meeting is the opening session of the fifteenth annual meeting of the Institute of Radio Engineers. Beginning in 1926 the annual meeting was conducted in connection with a Convention continuing over several days. The 1926 Convention was the first gathering of the kind carried out by the Institute. The meeting was an outstanding success and it is likely that the Conventions will become permanent annual events.

From one standpoint it can be imagined that the public may view with little patience the holding of a three-days convention by radio engineers. There is no other art, no other science, in which the public has taken or is taking so keen an interest, such a proprietary interest, as in radio. Even the Doctors of Medicine, and surgeons, who deal with the affairs of physical well-being—with life itself—are permitted to carry on their researches in their own constructive, if leisurely way, without being assailed as delinquents. And, when the best these savants can do to outwit parasitic influences, and to destroy malignant interferences, fails, they are forthwith absolved on the ground that the purposes of destiny or fate must be served.

The public demands of radio engineers solution of electrical and meteorological problems involved in establishing and maintaining year 'round, satisfactory radiophone transmission and reception.

Had the popular use of radio for entertainment purposes not been precipitated at a time when but a small number of engineers had become proficient in the art it is probable that advance for a time would have been conducted along lines other than that of broadcasting.

It is true, of course, that the great pressure brought upon the engineer to unravel the difficulties of electrical transmission over an uncontrollable medium have served to hasten desirable results. And, measured by the achievements of the past five years much has been accomplished. To realize the extent of the advance it is necessary only to view the best of the equipment in use five years ago and to recall the cacophonous emenations vibrated by the horns of 1921 into the virgin air.

However, we shall hope that the public will be indulgent and await the close of this convention before demanding More and Better Radio.

Some day one of us may have the leisure to write a story about the intellectual rise in radio. If the job is well done it is likely that a very interesting picture will be developed recording the purely electrical advance up to the point where a metal conductor between the source of power and the device to be operated no longer remained a necessity. Thenceforward the account will be less difficult to prepare historically because of the large amount of written matter available, produced by physicists, engineers and students engaged in carrying on research and experiments.

The elementary but spectacular 'spark' operation for telegraph purposes, for a decade prior to the advent of the Audion, enabled communication engineers to get used to the idea of wireless working.

The chronicler will perhaps need to exercise a sort of journalistic charity while dealing with radio advance from 1906 until 1912, but from the latter year forward when the possibilities of the wonder-working three-eelctrode tube were fully realized, the science has progressed to the stage where every man's house has a pair of ears—or, at least, an ear. It is a magic ear—one not only with which the listener may hear the Sunday sermon a few blocks distant from his morris chair, but with which he purposes to overhear neighborhood gossip taking place in a hamlet across the continent, or across the seas.

Indeed, even now, the secrets of the laboratories leaking out, every man will soon require that he be enabled at will visually to observe what is taking place at any point in the country where his inquiring eye may desire to roam.

A few years ago these developments were of the stuff of which dreams are made. The task of the radio engineer is to make these dreams come true, and the part which the Institute of Radio Engineers has played and shall continue to play in this realization is of the first importance.

In the beginning, fifteen years ago, the Institute was nursed into being by an earnest, enthusiastic, small group of amateurs, who were in fact the engineers of their day.

From 1912 until 1924, the Institute existed mainly for the purpose of holding periodical technical meetings in New York and for the purpose of financing the publication of radio technical papers; until 1915 groups of papers were published four times each year, and from 1916 until 1926 inclusive the Proceedings have been published six times each year, and with the January, 1927, issue the Proceedings are to be published monthly.

In 1924, President Morecroft introduced and fostered several organization and management betterments which paved the way for the aggressive, constructive administration of Institute affairs carried on by President Dellinger throughout the year 1925. These years witnessed marked gains in Institute development, activities and usefulness.

At the time the Institute was organized in May, 1912, there were less than fifty members. On June 1, 1914, there were 79 Members and 311 Associates. On January 1, 1916, there were 39 Fellows, 108 Members, 764 Associates and 73 Juniors; a total of 984. On January 1, 1922, the total was 1600, all grades, and on January 1, 1926, 2300, all grades. On January 1, 1927, the total membership was approximately 3,800.

The gratifying increase in membership during the past two years has been due largely to further development of the Section idea.

The Washington Section was organized in January, 1914; the Boston Section in November, 1914; the Seattle Section in February 1915, and the San Francisco Section in 1916.

The Chicago and Toronto Sections were organized in the summer of 1925, and in 1926 Sections have been organized in Rochester, N. Y.; Los Angeles, Calif.; Hartford, Conn., and the Philadelphia Section reorganized. During the year 1926, in addition to the eleven New York meetings, forty-nine meetings were conducted by the various Sections, at which technical papers were presented.

PUBLICATIONS

In 1926 a Year Book was published containing a complete list of the membership, together with a considerable amount of other matter of interest to members.

In the Proceedings forty-two technical papers were published, and supplies of pamphlet copies of all papers presented at New York were distributed at the meetings.

Work was begun on an Index of all papers presented before the Institute 1913 to 1926, inclusive.

Information has been gathered for the publication of a Year Book to be printed early in 1927.

Diplomas have been issued to Fellows and Members, and membership cards to Associates.

If among new features of management introduced in 1926, any may be noted as being an innovation, we might refer to the effort put forth to bridge over the gulf which in learned societies often separates the younger men and less experienced members from the advanced engineers and leaders in the science.

There is need and opportunity for further development along this line, but in the past year we have succeeded in having present at New York meetings many of the outstanding leaders in radio science, so that the younger men, largely in the majority, might become acquainted with and listen to oral discussion presented by engineers recognized as authorities.

Both expert and novice have been given opportunity to speak from the floor on the subjects presented at meetings. Already there is evidence that from this good has followed.

After all, the Institute's main purpose is to disseminate engineering information bearing on the present problems of radio, and it is plain that methods followed to set forth and to interpret the laboratory achievements of the physicists are successful and widely useful in proportion to the number of students who leave each meeting with knowledge gained of such advances and with clearer understanding of fundamentals.

MEETINGS AND PAPERS COMMITTEE

The Meetings and Papers Committee during the past year, under the direction of R. H. Marriott, provided excellent papers for all meetings and for the pages of the Proceedings. The committee held regular monthly meetings during the year and carried on a continuous correspondence for the purpose of procuring suitable papers for presentation.

MEMBERSHIP COMMITTEE

The Membership Committee, H. F. Dart, chairman, was thoroughly active throughout the year. Much of the substantial gain in membership is directly due to the work of this committee. Regular monthly meetings were held, usually in the evening, and there were many meetings of subcommittees.

ADMISSIONS COMMITTEE

This committee, Ralph Bown, chairman, held monthly meetings throughout the year. During the year the committee approved five applications for transfer to Fellow grade; sixty-seven applications for transfer from Associate to Member grade, and fifty-four direct elections to Member grade.

Twenty-four applicants for Member grade were recommended for Associate grade; seventeen applicants for transfer from Associate to Member grade were not approved, and four applications for transfer to Fellow grade were not approved.

STANDARDIZATION COMMITTEE

The Standardization Committee, L. E. Whittemore, chairman, has had a number of meetings during the year, the work done being preliminary to a revision of the 1926 Standards Report, to be published perhaps in 1928.

SECTIONS COMMITTEE

The Sections Committee, David H. Gage, chairman, is to be credited with much of the work done toward rehabilitating the activities of the older Sections and organization of the new Sections. The Committee has in hand constructive plans looking to the establishment of one or two additional Sections.

COMMITTEE ON PUBLICITY

The Committee on Publicity, W. G. H. Finch, chairman, has procured for the Institute a very gratifying amount of space in daily newspapers and radio periodicals. All of the New York meetings of the Institute have had satisfactory prior notice in news sections of dailies, and a considerable amount of matter relating to Institute activities has been published in weekly and monthly periodicals. The committee also has had entire charge of the extensive publicity given the conventions, and has attended to details of circularization and printing.

ADVERTISING AND NEWS NOTES IN PROCEEDINGS

The agency having charge of the procurement of paid advertisements for the Proceedings has done good work during the year as is evidenced by the number of full-page advertisements appearing regularly in those pages. With the Proceedings on a monthly basis the advertising space occupies should materially increase.

The new feature of incorporating in the first six or seven pages of each issue of the Proceedings notices, reports and news items of interest to members has had a particularly beneficial effect in increasing the interest of Institute members in all parts of the country, and Canada, in the activities of the officers and of the committees.

PRESENTATION OF THE LIEBMANN MEMORIAL PRIZE

The Institute each year makes two awards to radio investigators in recognition of noteworthy inventions or other meritorious developments in radio technique.

One award, the Institute Gold Medal of Honor has been awarded annually since the year 1918. The following engineers have received this prize: E. H. Armstrong, E. F. W. Alexanderson, G. Marconi, R. A. Fessenden, Lee De Forest, John Stone, M. I. Pupin and G. W. Pickard.

The Liebmann Memorial Prize consists of five hundred dollars in cash, being the annual interest on a sum of money presented to the Institute for the purpose of maintaining the award. This prize is awarded by the Board of Direction of the Institute each year to the engineer who has made what is clearly the most outstanding valuable contribution to the art and science of radio during the previous year.

Engineers who have received this prize since it was established in the year 1919, are: Leonard F. Fuller, Roy A. Weagant, Raymond A. Heising, C. S. Franklin, H. H. Beverage, John R. Carson and Frank Conrad.

The elements of mystery bound up in the term "Radio Transmission Phenomena" have a different meaning for the average non-technical broadcast listener, than for the radio engineer. The radio engineer is at all times several giant steps ahead of the merely curious devotee of the art. Even in the profession of radio engineering these are grades through which the studious and the ambitious may progress. An engineer's first-hand, exact knowledge of radio transmission phenomena may well be regarded as a test of his fitness to occupy a place in an advanced grade.

This year the Institute awards the Liebmann prize to an engineer whose researches and investigations into the more difficult element of transmission phenomena have resulted in giving us extensive and useful additions to existing knowledge.

It is my privilege to hand to Dr. Ralph Bown, this award, which goes to him with the respect and admiration of the members of the Board of Direction, and I am sure, with the congratulations of every member of the Institute, both here and abroad.

INTRODUCTION OF NEW OFFICERS

The next and final duty which rests upon me is to introduce the president-elect and vice-president elect.

Due largely to the great growth and extension of radio engineering and to the interest taken by the members in the affairs of the Institute, the organization now occupies a place in the family of engineering associations approaching in prestige and importance that enjoyed by the older and larger engineering groups.

The office of president of the Institute of Radio Engineers has become one in which there is great opportunity for constructive doing. The prestige of the Institute as an engineering source is widely recognized. The membership of the Institute is made up mainly of comparatively young men who look to the organization for instruction, for technical information and for opportunity to meet and mingle with their fellows.

The president's task is to initiate and support Institute activities which will insure the availability of these facilities to all members no matter where they may be situated. The gentleman this year elected to occupy the chair is well qualified to carry forward the interests of our beloved Institute to higher levels and to greater usefulness. I take pleasure in presenting president-elect Dr. Ralph Bown.

RESPONSE OF DR. RALPH BOWN

Incoming President of the Institute of Radio Engineers

I wish to say to you frankly that I am glad to be President of this Institute, that I will do all in my power to be worthy of the responsibility and will carry out the duties of the office to the best of my ability.

I feel sure that in many respects I will be unable to reach the standard of performance exhibited by our retiring President, Mr. McNicol. The devotedness with which he has served the Institute in the past year, while well known to the Board of Directors, is not sufficiently known to the membership at large. To him we owe great advances in the business operation of the Institute. He has devoted probably more than half his time to Institute affairs and has been particularly active in aiding the growth of our various sections. He has just finished presenting to you a report of various activities of the Institute during the past year and some of the plans which have been made for the present year. That report is so complete that I hesitate to burden you with anything further of the same character, so I think that it would perhaps be more fitting if I spent a few minutes in drawing your attention to some other aspect of Institute affairs. I have in mind some reflections which have been engendered by the opportunity I have had during the past year to meet a number of our foreign members. Or perhaps I might better say of our members in foreign countries, since the Institute is so organized that the nationality of a member has no relation to his membership status. In this respect at least, the Institute may be said to be international or world wide in scope.

The total number of foreign members on January 7th was 412 or slightly more than 10 per cent. of our entire membership. Many of these are distinguished engineers of world-wide fame such, for instance, as Senator Marconi, the inventor of wireless telegraphy, Valdemar Poulsen, inventor of the Poulsen Arc, J. Zenneck, the well-known German engineer, E. H. Shaughnessy, Assistant Engineer-in-Chief of the British General Post Office, C. P. Edwards, Director of Radio Service of the Canadian Government, to mention only a few.

If the foreign membership were represented in this meeting today in the same ratio as the domestic membership, they would make a very impressive showing and occupy a goodly allotment of seats.

The greatest concentration of foreign members is in England, there being 235 or more than half the total number. One reason for this is that the Proceedings of the Institute are printed in the English language and they carry technical papers which all radio engineers find to be of interest and value. Perhaps this explains also why Japan is prominently represented by 27 members since the English language is frequently used by Japanese engineers and scientists.

It is not uncommon to hear foreign members characterize the Proceedings as the best thing of the kind available in the world. Many foreign members have become members because they wish to obtain the Proceedings and membership is the most satisfactory way to obtain them. Perhaps I should have used the past tense in these remarks, because there seems already to be growing up in the foreign field a feeling that membership is desirable also because of the prestige and professional standing which it implies. And this something we call "prestige"—this intangible asset which we are beginning to acquire—I believe we should guard jealously and foster wisely not only abroad but at home. It is a young and tender plant which we should nurture and cultivate to vigorous maturity as time goes on, for upon it rests in large measure the power of the Institute to serve the public interest and to benefit its individual members.

In doing this, we must remember that the Institute is distinctly American in principle. It does not aim to obtain prestige by virtue of having its membership restricted to a select few who have passed many tests and have acquired sufficient professional standing to add luster to its rolls—on the contrary, it endeavors to enroll and benefit anyone who is interested in radio engineering. To be sure, its higher grades of membership necessarily represent greater proficiency in the art, but suitable grades are open to any who have a sincere desire to associate themselves with radio engineering work. It is a distinctly democratic organization.

High standing, therefore, must flow from recognized high standards and utility of its output. The actions of its governing board must be wise and sound and free from any partisan influence. The reports of its technical committees must be authoritative and valuable to the radio industry and to the public generally. The papers which are read and discussed at its meetings must be well written presentations of important matter of broad interest, avoiding the trivial and self-seeking. And perhaps most important of all, its Proceedings must be maintained at the high standard of the past and improved wherever opportunity offers.

We have grown mightily in numbers in the past few years. Let us see to it that we grow even more in utility and in the dignity of service.

THE CORRELATION OF RADIO RECEPTION WITH SOLAR ACTIVITY AND TERRESTRIAL MAGNETISM*

By

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(Communication from the International Union of Scientific Radio Telegraphy)

One of the outstanding problems today is the nature and cause of those atmospheric changes which produce such diversified effects as weather, magnetic storms and disturbances of radio reception. The problem is meteorological; if this earth has no atmosphere there could be no weather, on an airless planet there could be no long distance communication at broadcasting and higher frequencies and I think magneticians will agree that the phenomena of terrestrial magnetism would be altered if the atmosphere were removed.

The only known important force which acts upon the atmosphere is the complex radiation and emission from that variable star which we call the sun. Changes in this force are caused in two ways; first by the movements of the earth with respect to the sun, and second by actual variations in solar radiation. If the sun maintained a constant radiation, we should have only to consider the earth's rotation on its axis, which gives us night and day, and its movement in an orbit around the sun, which by the changing angle of the solar rays gives us the seasons. If these movements were the only factors involved, weather, terrestrial magnetism and radio reception would follow the calendar to a far greater extent than our measurements indicate.

But in the scheme of things as they are, we find that weather does not go according to the calendar, nor does radio reception. The visual evidence of sunspots, faculae

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and prominences tells us that the sun is periodically disturbed, and measurements of the light and heat received by the earth have shown that this varies in general correspondence with visible changes on the sun's disk. Through the work of Abbott and Clayton[1] definite relations have been established between solar changes and weather, which have already been usefully applied to weather forecasting.

Less definite today is our knowledge of the short wave and corpuscular radiation from the sun, which cause ionization and electrical currents in the atmosphere, and even chemical changes. Our only direct indices of these radiations are such things as disturbances of terrestrial magnetism, atmospheric electricity and radio reception, atlhough over long periods they are highly related to sunspots and other visible changes of the sun's surface. And as radio research has not yet become a pure science, we do not have such systematic records to study as those gathered through the years by astronomical and magnetic observatories.

One purpose of this paper is to emphasize the importance of systematic long-period observations of radio reception; it is not an overstatement to say that the picture we hope to draw of the relations of earth and sun is in material part dependent upon such measurements. But at the present time I cannot present a full account of the correlation between radio reception, solar activity and magnetic disturbances, for the simple reason that I have not yet the right sort and amount of data. This paper is therefore merely a progress report setting forth the results so far obtained, which I believe are now sufficiently definite to be of interest to other workers in this field.

Several times in the past twenty years I have attempted systematic measurements of reception from distant stations, in the hope of finding some correlation with other elements. In 1906 I devised[2] a method of accurately measuring received energy from a distant spark station, and for several months in that year I made almost nightly observations at Amesbury, Massachusetts, of the energy received from station CC at South Welfreet, Massachusetts, distant 145 kilometers and operating at 167 kilocycles. I found that the received

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[1] Solar Radiation and Weather, or Forecasting Weather from Observations of the Sun. H. H. Clayton, Smithsonian Miscellaneous Collections, Vol. 77, No. 6, June 20, 1925.
Solar Activity and Long Period Weather Changes. H. H. Clayton, Smithsonian Miscellaneous Collections, Vol. 78, No. 4, September 30, 1926

[2] The Measurement of Received Energy at Wireless Stations. Electrical Review, December 15, 1906.

energy fluctuated over a wide range from day to day, but there was no apparent correlation between these values and either sunspots, terrestrial magnetism or meteorological elements. I am now satisfied that in this series the principal fluctuations were due to changes in the radiation of the transmitter, which was far from constant in either power input, frequency or antenna current.

In 1909[3] I again began a series of reception measurements in Amesbury from station GB at Glace Bay, Nova Scotia, using the audibility meter method of measuring signal intensity. These measurements were repeated at my suggestion by Messrs. Dolbear and Proctor in 1911[4]. Although this series determined an interesting sunset and dawn effect[5] at the time no correlation was found with other elements. More recently I have correlated the night measurements of this series with van Dijk's daily values for the mean magnetic character of days[6], with the result shown below.

Date		Reception Deviation From Mean	Mean Magnetic Character of Day.
January	30, 1909	1.04	1.7
July	25, "	1.37	0.3
"	27, "	0.63	0.7
March	9, 1911	1.41	0.2
"	11-12, "	1.60	0.1
"	13-14, "	0.92	0.7
"	15-16, "	0.25	0.7
"	17-18, "	0.73	0.1
"	21-22, "	1.15	1.4
"	22-23, "	0.91	1.3

The correlation here is $r = -0.214 \pm 0.204$. This is of course a mathematical equivalent of the Scotch verdict "not proven", but the negative coefficient of correlation is interesting in view of my later work, in which an inverse relation with solar and magnetic measures was definitely found. Those familiar with correlation work will realize that this series was too short for definite results, as it consisted of only ten terms. I am also doubtful of the constancy of the transmitter from night to night.

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[3] "Principles of Wireless Telegraphy" G. W. Pierce, New York, 1910, page 135.

[4] The Effects of Sunlight on the Transmission of Wireless Signals, by B. L. Dolbear and J. A. Proctor, Electrical World, N. Y., Vol. 58, No. 6, August 5, 1911, pages 321-323.

[5] The Daylight Effect in Radio Telegraphy, by A. E. Kennelly, Proceedings of the Institute of Radio Engineers, Vol. 1, No. 6, July, 1913.

[6] The Magnetic Character of the Year, by G. van Dijk, Terrestrial Magnetism and Atmospheric Electricity, Vol. XV, No. 3, page 166, and Vol. XVII, No. 3, page 150.

In one of those all-too-rare compilations of reception data, Marriott has given [7] a tabulation of the variation in nightly range of station DF at Manhattan Beach, New York, over the period October 12, 1908 to October 15, 1909. According to the records of Cheltenham Observatory, Maryland[8] this interval included 33 magnetic storms, one of these —September 25-26, 1909—being the most severe of which we have any record. An examination of Marriott's Chart No. 3, Fig. 1 will show that no less than 26 of these storms coincided with depressions in the range curve.

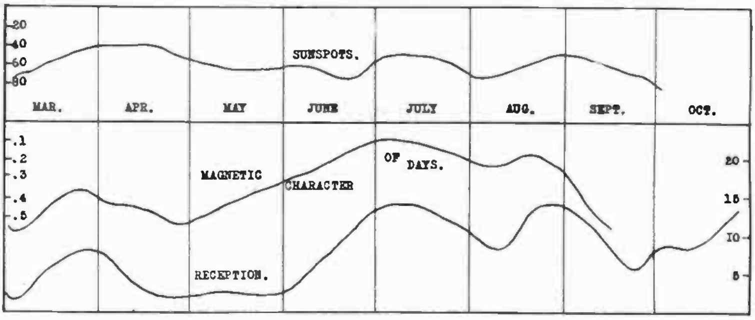

Figure 1.—Moving monthly averages of sunspots, magnetic character of days and reception from WBBM, Chicago. 1926.

Throughout 1909 I kept a reception log at Amesbury, and in this I also find unmistakable coincidences between low reception and magnetic storms. On the evening of September 25 of that year DF was heard but once, with a maximum audibility of 2. On the evening of the 26th DF was not audible at any time, and for several evenings thereafter reception from this station was abnormally low. As I shall show later, our earth recovers its magnetic equilibrium much more rapidly after a magnetic storm than does the coincident depression in radio reception. It is as if some force dealt the earth a blow, simultaneously disturbing its magnetic state and so altering the atmosphere as to lower radio reception. Long after the electrical movements which produce the magnetic storm have subsided, the atmosphere apparently remains unfavorably ionized with respect to radio wave transmission, save at the ultra high frequencies.

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[7] Radio Range Variation, by Robert H. Marriott, Proceedings of the Institute of Radio Engineers, Vol. 2, No. 1, March, 1914, pages 37-53.

[8] Terrestrial Magnetism and Atmospheric Electricity, Vol. XIV, No. 1, page 40; No. 2, page 86; No. 3, page 153; No. 4, page 182.

Beginning with the advent of broadcasting, I recommend once more a systematic measurement of field intensities from distant stations, at first (in 1922) by audibility meter, and later (in 1923) by continuous photographic recording[9]. I again began to notice coincidences between magnetic storms and depressed reception despite the fact that the first years of broadcasting fell in a happy period of minimum solar activity and a magnetically quiescent earth. For convenient reference I give below a list of the magnetic storms of the broadcasting era, with their intensities in gammas. This magnetic unit is 0.00001 gauss, and corresponds approximately to one twenty-thousandth of the earth's normal horizontal field at Cheltenham. Although in the ordinary magnetic storm the change in horizontal force is only of the order of one per cent, during the great storm of September 25-26, 1909, there was a change of nearly thirteen per cent in horizontal field.

Principal Magnetic Storms, 1922-1926 Cheltenham.

Date		Range in Gammas
April	21-22, 1922	186
September	13-15, "	179
October	5- 7, "	193
March	24-25, 1923	183
September	26-28, "	161
November	15-17, "	173
January	29-30, 1924	210
May	21-24, "	187
June	9-11, "	168
June	11-11, "	233
September	7- 8, "	96
September	23-24, "	145
October	23-24, "	119
January	19-20, 1925	103
May	3- 5, "	178
May	30-31, "	91
June	13-14, "	100
June	24-25, "	186
July	25-27, "	122
August	22-23, "	170
September	1- 2, "	153
September	14-15, "	152
September	21-21, "	400
September	24-24, "	186
October	23-24, "	144
November	8- 9, "	129
January	26-27, 1926	548
February	23-25, "	424
March	5- 6, "	286
April	14-17, "	543
May	3- 6, "	147
June	1- 2, "	265
September	9- 9, "	
September	15-15, "	
September	20-21, "	
October	14-16, "	

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[9] Short Period Variations in Radio Reception, Proceedings of the Institute of Radio Engineers, Vol. 12, No. 2, April, 1924, pages 119-158.

A search through newspaper files will show in the radio columns reports of poor broadcast reception coinciding with practically all of the above listed storms. Of this I will give one example. In the Boston Post of September 29, 1923, appeared the following:

> Unusual radio conditions are seriously hampering reception at the present time. This condition was noticeable about this time last year and continued for nearly a week, and has caused trouble for the past two nights. The cause is unknown and while there is hardly a trace of static or fading, it is impossible to tune in stations with any degree of success. While powerful stations like WGY, WMAF and others in New York and Philadelphia may be coaxed in, they seem to cause a click in the phones and disappear, leaving a hum or air rushing sound in the phones. Hundreds of fans listening in during the past two nights have started looking over their sets, changing tubes and batteries and even taking them down. Don't be surprised if this condition continues. The trouble is not with the set but with unusual receiving conditions.

The reference to "about this time last year" is to a period from September 13 to 17, 1922, when the then small but rapidly growing army of broadcast listeners found distant reception poor or impossible. There was a magnetic storm September 13-15, 1922. Similarly, there was another storm September 26-28, 1923, coinciding with the distressful interval so graphically set forth in the quotation above.

Although as I have indicated above, both the technical and popular literature of this art contain sufficient reception data to show a distinct relation to terrestrial magnetism, an accurate knowledge of this and other correlations requires systematic measurements over long periods. An ideal series would consist of daily observations covering the entire radio spectrum over at least a sunspot cycle of eleven years. But this art's solitary boast in such measurements is Austin's series beginning in 1914, of Washington reception from Nauen and other European stations. Unfortunately the low frequency end of the radio spectrum is not particularly sensitive to solar disturbances, and has a distracting relation to certain meteorological elements, so until recently I have not found solar and magnetic relations in this series.

The most sensitive part of the radio spectrum appears to be that band set apart for broadcasting, and more particularly that part from 1000 to 1500 kilocycles. From the viewpoint of this investigation, this is a most fortunate coincidence, for among the radio stations of this world only the broadcasters run "keylocked" for hours each evening, and their unbroken radiation permits a precision of measurement impossible with telegraphically modulated transmitters.

During the fall and winter of 1925 I made an extended survey of night reception from distant broadcasting stations, in which I found that in general these stations swung together, a bad night for one being usually a bad night for all. And although I had previously found that short period fading is most violent from stations at moderate distances, it appeared that the greatest differences in mean field from night to night were for stations over a thousand kilometers distant, and operating at the high frequency end of the broadcast band. Finally I selected WBBM at Chicago, operating at 1330 kilocycles, as the best available station for systematic measurements, and early in 1926 I began a series of nightly measurements of this station which have carried through to date without substantial break, and which I hope to continue for at least a year.

The night field from a distant broadcast station is far from uniform, varying in a semi-periodic manner from minute to minute. The amplitude of these fluctuations is so great that no single measurement gives any true idea of the mean field, so it is necessary to either take a large number of separated readings and average them, or, better, to make a continuous record for at least an hour, and then find the mean field by planimeter measurement. Owing to the fact that WBBM shares the frequency of 1330 kilocycles with another Chicago station, WIBO, and to the change in schedule caused by daylight saving time, there is but one hour each evening, 9-10 p. m., 75th meridian time—which can be used throughout the year. With the exception of July, August and September, the field from WBBM has been measured at Newton Centre, Massachusetts. During the summer the measurements were made at Seabrook Beach, New Hampshire, after an overlapping series of observations had shown that on the same evening there was no appreciable difference in mean field at these two points.

As I have already given the method of making such records[9] this will not be described here. An open antenna with an effective height of eight meters was used, with a superheterodyne receiver operating the recorder galvanometer. The record scale was calibrated by introducing a known radio frequency voltage in the antenna circuit, Roberts[10] ingenious arrangement being used as the source of high frequency.

The nightly field measurements from January to the

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[10] Generation and Measurement of Weak Radio-frequency Currents. W. van B. Roberts. Journal Franklin Institute, 201, pages 301-310, March 1926.

end of November, 1926, are given in an appended table. For nights when WBBM was not operating—Monday night of each week is a silent night for all Chicago stations—or when for any reason I was unable to make observations, I have included an approximate value based upon either reception from a group of other stations and designated by "*e*", or by the arithmetic mean of the values for the nights on each side, indicated by "*m*". In all cases the table gives mean field in microvolts per meter for the period 9-10 p. m., 75th meridian time. This tabulation will perhaps be useful to other investigators, particularly for correlation with solar or meteorological elements.

In the analysis of reception data I have made extensive use of moving averages[11] both for the purpose of showing general trends over long periods, or of short-time fluctuations. The principal periods which I have used are the week and the month; for all geophysical data this latter period should be made 27 days, in order to eliminate a solar period.

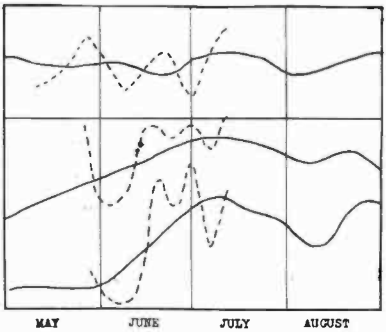

Figure 2.—Sunspots, magnetic character of days and reception, moving monthly averages in broken line. 1926.

In Fig. 1 the lower curve is a moving monthly average of WBBM reception from March to October, 1926, in microvolts per meter mean field. Above this, with inverted ordinates, is a moving monthly average of magnetic character of day numbers, as given by the Cheltenham Observatory. It is clear that these two curves march together, and a correlation between these elements on a monthly average basis gives $r = -0.89 \pm 0.06$. This is good correlation, for the

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[11] "First Course in Statistical Method" by G. Irving Gavett, New York, 1925, pages 207-209.

coefficient is high and nearly fifteen times the probable error. The upper curve of this figure is a moving monthly average of the Wolfer Provisional Sunspot Numbers, and has no clear relation to reception or magnetic measures,

Mean field in microvolts per meter at Newton Centre, Mass., from WBBM, Chicago, 9-10 P. M., 1926.

Day	Jan.	Feb.	Mar.	Apr.	May	June	July	Aug.	Sept.	Oct.	Nov.	Dec.
1	....	2.0e	0.0e	3.0	9.1	0.2	39.3	0.2	25.5	9.6	32.0e	0.0e
2	....	8.0e	4.0e	18.2	4.3	4.0e	17.0	0.2m	19.5	7.6	17.3	2.9
3	....	8.0e	2.0e	12.8	8.0e	0.0	3.2	0.2	25.0	8.8	2.8	2.4
4	....	16.0e	0.5	6.6	0.6	0.8	5.1m	6.5	25.0	8.0e	0.5x	2.4
5	....	2.0e	0.2	16.0e	16.0e	0.1	7.0	4.4	16.5	0.3	1.7	7.7
6	....	5.0e	1.0	3.7	0.0	0.8	8.8	0.4	18.3m	7.4	34.0	4.0e
7	....	8.0e	0.0	2.2m	0.2	4.0e	2.9	8.3	20.0	1.3	20.0	18.7
8	....	8.0e	8.0e	0.6	4.0	1.3	17.8	24.6	1.0	8.6	16.0e	6.4
9	....	2.0e	0.3	2.6	4.0	1.0	7.8	19.6m	7.0	35.4	24.0	3.5
10	0.0e	4.0e	3.0	2.8	2.0e	1.7	15.9	14.5	4.8	30.8	10.4	20.6
11	2.0e	2.0e	0.4	3.9	4.0	0.1	24.0	14.2	10.3	30.6m	20.0	13.1e
12	4.0e	0.0e	0.8	16.0e	5.3	0.9	22.9m	13.0	7.5	30.5	25.0	21.6
13	2.0e	0.0e	0.0	3.0	2.8	1.5	21.8	5.7	3.9m	5.6	22.0	16.0e
14	2.0e	0.0e	14.6	0.0	0.0	4.0e	0.7	2.2	0.2	0.0	18.0	6.1
15	2.0e	2.0e	0.0e	0.5	0.0	6.8	7.3m	5.2	3.9m	1.6	16.0e	....
16	1.0e	9.0	0.3	0.2	2.7	8.2	13.8	5.6m	7.5	0.2	7.8	....
17	0.0e	3.8	16.0e	0.0	8.0e	20.2	24.8	6.0	1.9	0.0	6.6	....
18	4.0e	8.3	2.0	0.3	3.3	8.8	10.2	11.9	2.4	4.0e	3.5	....
19	8.0e	2.0	4.0	2.0e	4.8	42.8	19.0m	3.9	2.8	0.0	4.5	....
20	4.0e	3.4	11.0	5.2	5.4	10.8	27.8	0.6	2.2m	0.0	9.4	....
21	16.0e	2.4	4.7	0.4	3.8	8.0e	24.1	14.5	1.6	0.1	7.5	....
22	0.0e	4.0e	8.0e	0.5	0.2	7.5	9.3	17.5	0.0	0.6	4.0e	....
23	4.0e	0.3	28.0	0.1	0.1	17.7	21.7	11.4m	0.0	5.3	20.3	....
24	8.0e	0.4	9.0	0.0	8.0e	16.0e	9.4	5.3	0.1	7.7	2.0	....
25	2.0e	0.2	15.4	9.5	2.5	12.0	13.0	15.5	4.8	8.0e	19.8	....
26	2.0e	0.0	27.0	2.0e	6.1	7.3	11.0m	10.5	6.5	8.6	9.4	....
27	2.0e	0.0	26.0	0.1	2.5	16.9	9.1m	0.5	11.1m	1.9	19.3	....
28	0.0e	0.0	3.2	2.2	7.5	16.0e	7.2	30.5	15.7	6.6	2.0e	....
29	0.0e		2.0e	7.2	4.6	8.0e	6.1	27.5	8.1	9.3	0.0e	....
30	0.0e		1.4	0.0	6.5	12.0	13.2	30.0m	12.1	50.5	4.0e	....
31	0.0e		6.7		8.0e		3.2	32.5		26.7		....
Means.	2.9	3.6	6.4	4.1	4.3	8.0	13.7	11.1	8.8	10.2	12.7	

e. Estimated from observations of other stations.

m. Mean of measurements on each side.

x. On November 4 WBBM increased power, and fields from then on are reduced to a basis of 10 amperes in the antenna, which was the current from January 10, to November 4.

either by inspection or actual correlation on a monthly basis.

But in Fig. 2, which gives in full-line a portion of the three curves of the preceding figure, I have superposed on the moving monthly averages a broken line represently moving weekly averages. Now relation can be seen between sunspots, terrestrial magnetism and reception, as the broken line curves, which represent deviations of weekly averages from monthly means, are obviously similar. An increase of solar activity, indicated by a part of the broken line below the monthly average or full-line axis, is paralleled by an increase in magnetic disturbance, and a decrease in reception.

As we are no longer concerned with the shape of the monthly average curve, this may be flattened out to a zero or base line, and the weekly deviations alone plotted. This is done in Fig. 3 for sunspots, magnetic character of days, and reception. Although these three curves do not run absolutely together, there being a marked phase displacement at times, there is the same succession of peaks and hollows; there is no question but that these three elements are related. The period so strongly shown here is 27 days. This is a solar, and not a lunar period, and is in fact due to the recurrent earthward presentation of sunspot or otherwise active areas. More than one casual observer has noted a coincidence between, say, the full of the moon and depressed reception; lunar and solar periods are so nearly alike that once started such coincidences will run through several cycles. But eventually the lunar period will get out of step with reception, and then it will be obvious that

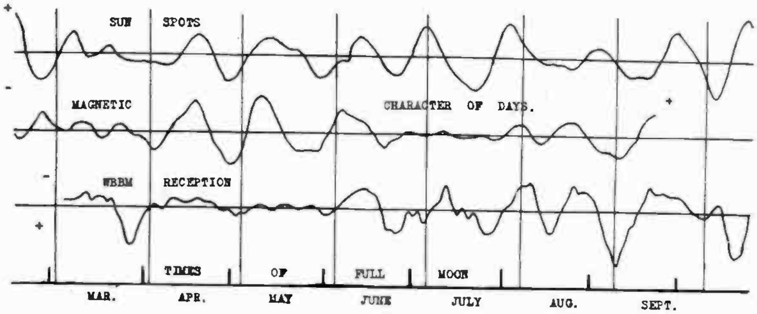

Figure 3.—Weekly departures from monthly means of sunspots, magnetic character of days and reception. 1926.

there is no real relation. This is illustrated in Fig. 3, where I have shown at the bottom the times of full moon. Those who hold lunar superstitions will note that while in August and September the full moon came at or near minimum reception, it came near periods of maximum reception in March, April and May, while in June and July it approximately coincided with normal reception!

Although the rotation of the sun never changes phase, the periodic waxing and waning of the sunspot numbers sometimes do. As the average life of a sunspot is two or three months, a spot or group of spots may persist through several rotations or cycles and then disappear, another spot or group of spots appearing at some other point, often in a quite different longitude. This gives a series having a 27-day period, but with occasional changes in phase, much as a bowed violin string vibrates.

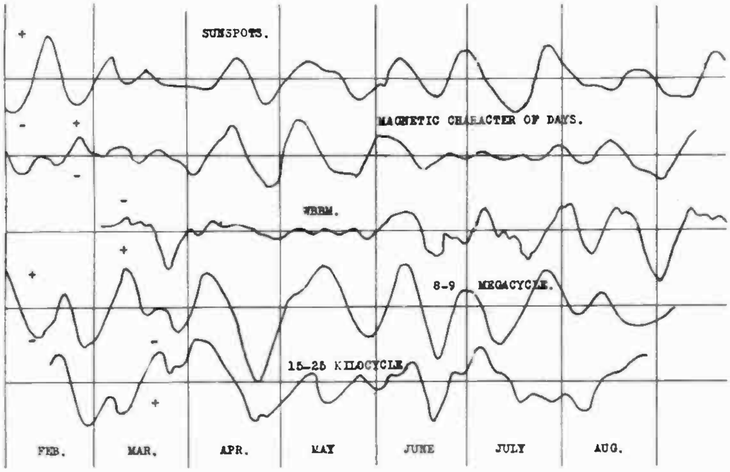

Figure 4.—Weekly averages as deviations from monthly means of sunspots, magnetic character of days, WBBM reception, 8-9 megacycle reception and 15-25 kilocycle reception. 1926.

Although as I have stated above, the broadcast band is the most sensitive to solar disturbances, effects are in fact produced over the entire radio spectrum. In Fig. 4 I have plotted five elements, sunspots, magnetic character of days, WBBM reception, 8-9 megacycle reception and 15-25 kilocycle reception. The 8-9 megacycle reception is at Washington from NPG, NPM, NBA, NBH, NPL, NIDK and from the U. S. S. Scorpion in the Mediterranean, the transmission paths being all in darkness. The 15-25 kilocycle reception is also at Washington, from GBR, LPV, FT, AGW, FU, LY, AGS, KET and LPZ, taken at 10.00 a. m., so that the transmission paths were by daylight. The three reception curves are in general alike, the peaks and hollows showing the same succession. But the 8-9 megacycle curve is inverted with respect to the other reception curves, so that this figure gives us the following relation: As solar activity, measured by sunspots, increases, reception in the low-frequency and broadcast bands decreases, but reception at 8-9 megacycles increases. In view of the other inversions which we have found in passing from low to very high-frequency transmission, such as the reversal of the day to night effects, and the inverted eclipse effect which I found in 1925[12] we are quite prepared to find that these ultra high frequencies, unlike their lower brethren, actually thrive on sunspots and magnetic storms.

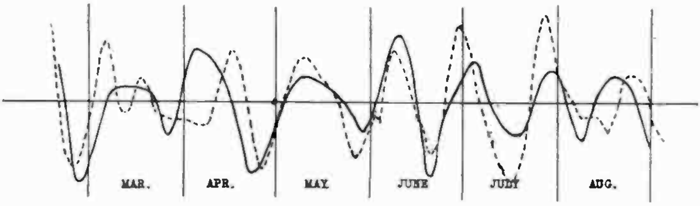

Figure 5.—Average of weekly deviations from monthly means for nine stations in the 12-25 kilocycle band, WBBM at 1330 kilocycles and five stations in 8-9 megacycle band. The broken line is the departure of weekly averages of wolfer provisional sunspot numbers from their monthly means. 1926.

Fig. 5 is an arithmetic average of all three reception curves of Fig. 4, with the sunspot curve superposed in broken line. Here is an average of fifteen stations, scattered over nearly a hemisphere, from the Hawaiian Islands, Pacific Coast, the Isthmus, South America, the North Atlantic and points in Europe. And these stations are not only geographically distributed, but they include both ends and the middle of the entire present day radio spectrum. But the averaging of such apparently diverse material has not obliterated the relation with the sun; instead this relation has been accentuated by the process.

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[12] The Effect of the Solar Eclipse of January 24, 1925, on Radio Reception. Proceedings of the Institute of Radio Engineers, Vol. 13, No. 5, October, 1925, page 567.

As I have pointed out above, certain phase differences appear in the various graphs of sunspots, magnetism and reception. Some of these differences are due to the effect which I have already mentioned, the slow recovery of reception from the original disturbance. In Fig. 6 I have plotted an average of four magnetic storms and their accompanying reception depressions, and it will be seen that although the magnetic storm and reception depression begin together, the storm reaches its maximum before reception is at a minimum, and magnetic quiescence returns two or three days before reception is normal.

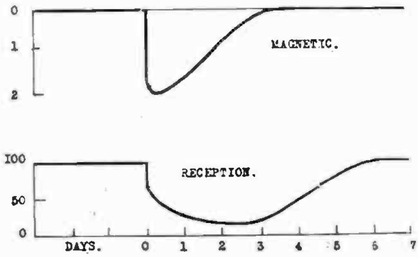

Figure 6.—Magnetic storms and reception. Average of storms of February 23-24, March 5, April 14-16 and May 4-6, 1926. Reception at Newton Centre, Massachusetts, from WBBM, Chicago, Illinois, at 1330 kilocycles, 9-10, 75th meridian time.

I have also made a preliminary analysis of WBBM reception with respect to meteorological elements, and particularly with respect to barometric gradients. So far the result has been negative, the field at Newton Centre, when solar and magnetic periods are removed, does not seem to depend in any way upon the relation of the line joining Newton and Chicago to the isobars of the weather map. But there seems to be a slight relation, which I have not yet fully investigated, between barometric activity and reception. Apparently days with great fluctuations of air pressure tend also to be days of low reception. The relation here is probably indirect; that is, barometric activity may be linked with solar disturbances, which are in turn associated with reception.

It is perhaps unlikely that any high correlation between reception and weather elements will be found. Solar disturbances and magnetic storms are world-wide events, whereas weather is rather a local matter. Analysis of weather elements over the whole earth indicate that there are areas of positive correlation with sunspots, and also areas of negative correlation. Although I have not yet collected and analyzed reception data from any such collection of receiving points as would fairly represent the earth as a whole, I have found that a bad night for reception in Newton Centre is in general a bad night anywhere in the United States. And I have also found that European reception of distant broadcast stations agrees remarkably well with my measurements of WBBM.

There is some basis for the assumption that reception is principally affected by corpuscular radiation from the sun, perhaps in the form of alpha particles. Maunder has found that magnetic disturbances seem to arise from restricted solar areas, not necessarily including sunspots, and to go out in definite directions, or rather shafts of several degrees diameter, which rotate with the sun. When such a shaft strikes the earth a magnetic storm arises. Such lines of influence are not, she thinks, necessarily radial, but may follow coronal stream lines. I find that, in general, reception is most affected when a spot or group of spots is near the center of the solar disk, that is, when they most nearly face the earth, although there are exceptions. Thus, on November 28-29, 1926, a group of three fair sized spots faced the earth, and coincidently there was a marked depression in reception. Occasionally a large spot may face earthward without any accompanying change in reception, which might be explained either by the assumption that the spot was not accompanied by corpuscular radiation, or on Maunder's hypothesis that its beam was so curved that it did not strike the earth.

But the secrets of this universe yield rather to observation than to pure speculation. When we have a sufficiency of the right kind of data we can frame stable explanations; until then we are groping in the dark. The relation of earth and sun is a dominant one to mankind, and the study of radio transmission phenomena may well throw new light upon this little-understood subject.

I must not conclude without thanking Dr. L. A. Bauer, of the Carnegie Department of Terrestrial Magnetism, for the magnetic and other data which he has supplied, as well as for his many helpful suggestions and also the U. S. Coast and Geodetic Survey for the Cheltenham data which I have so freely used. And I am indebted to Dr. A. H. Taylor of the Naval Research Laboratory for the high-frequency reception logs which he so kindly furnished. Finally I must express my appreciation of station WBBM, not only for its involuntary supply of radiation, but for its frequent readings of antenna current each night, which have enabled me to reduce my measurements to a constant basis.

IMPORTANCE OF LABORATORY MEASUREMENTS IN THE DESIGN OF RADIO RECEIVERS*

W. A. MacDonald

(Chief Engineer Hazeltine Corporation Laboratory)

The subject of this paper relates primarily to the measured characteristics of radio receivers and the importance of such measurements on the design of commercial broadcast instruments.

It is obvious that an exact knowledge of the individual and over-all characteristics of a radio receiver should be accurately known, yet experience shows that many manufacturers, including some of the largest, are practically unaware of the exact performance of the apparatus they produce.

Some time ago the Hazeltine Corporation was confronted with the problem of measuring the essential characteristics of a large variety of receivers. The first question to be answered was: What are the essential characteristics? Such a question lends itself to considerable discussion, but after careful consideration of all the possible factors it was decided that thirteen fundamental measurements were absolutely necessary in all cases. Additional measurements can be added to meet special circuits or unusual conditions. The thirteen fundamental measurements are as follows:

(1) Voltage step-up of input coupling transformer.
(2) Voltage step-up of 1st tube and coupling transformer.
(3) Voltage step-up of 2nd tube and coupling transformer.
(4) Voltage step-up of following stages.
(5) Complete R. F. amplification from input coup-coil to the detector.
(6) Resonance characteristic of input coupling transformer.

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*Received by the Editor Dec. 22, 1926. Presented before the Rochester Section November 30, 1926, and at the Convention of the Institute of Radio Engineers, New York, January 12, 1927.

(7) Resonance characteristic of 1st stage transformer.

(8) Resonance characteristic of 2nd stage transformer.

(9) Resonance characteristic of following stage transformers.

(10 Resonance characteristic of complete R. F. amplifier from input to the detector.

(11) Amplification and frequency characteristics of 1st audio transformer.

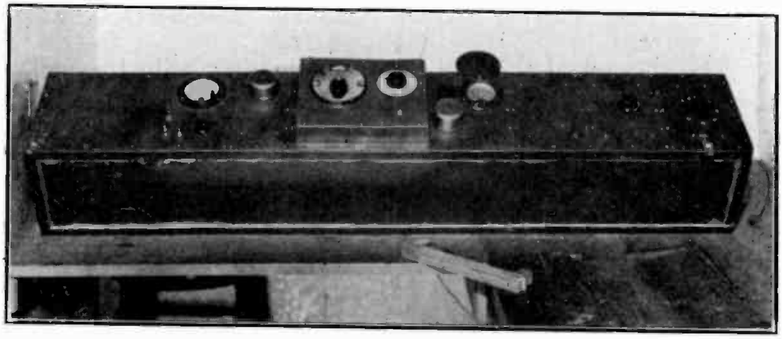

Figure 1

(12) Amplification and frequency characteristics of other audio transformers.

(13) Relative frequency characteristics of complete audio system including detector.

These measurements are selected because they are not especially difficult to make. They can be readily duplicated and the complete series standardized to a degree where the performance of a radio receiver can be accurately determined.

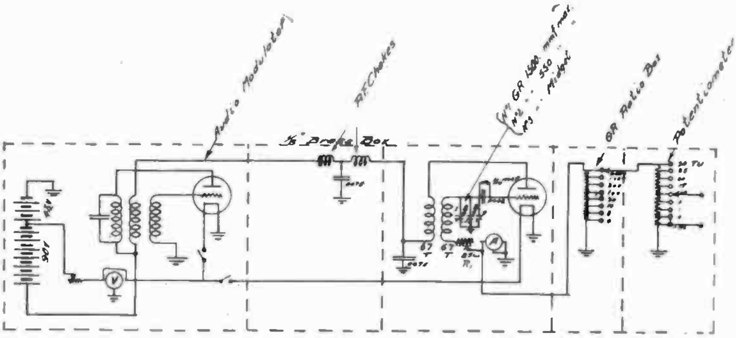

Figure 2—Laboratory Radio Frequency Oscillator

The apparatus required is not unduly complicated. It consists essentially of the following:

(1) Precision wave meter.
(2) Radio frequency oscillator.
(3) Audio frequency oscillator.
(4) Vacuum tube voltmeter.

The wave meter can be of any standard make and requires no description.

The radio frequency oscillator may consist of any convenient oscillating circuit, although one giving reasonably uniform output is preferable. If a simple oscillating circuit is employed it should preferably consist of small inductance and large capacity. Where high accuracy is required the oscillator and all batteries should be completely shielded to eliminate stray fields.

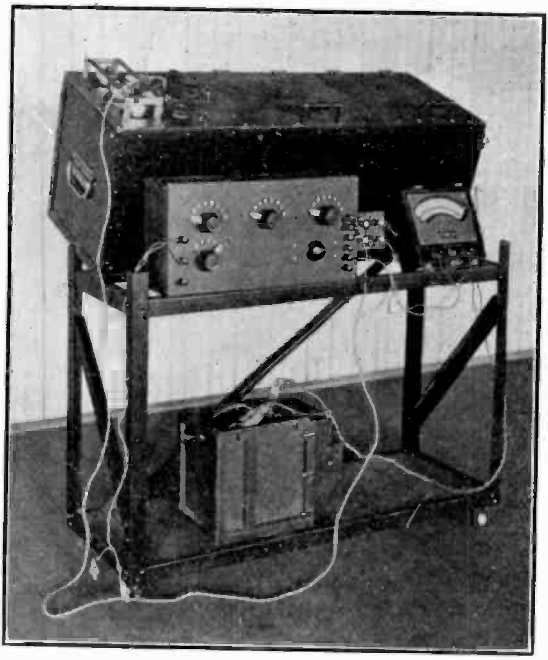

Figure 3

Figure 1 is an illustration of such an oscillator, while Figure 2 is the schematic circuit arrangement. Turning to the illustration, the dials marked $C1$ and $C2$ are the tuning and vernier adjustments. A is the output meter and $R1$ an adjustable resistance for regulating the H. F. output.

An output voltage attenuator is located at the extreme right.

The audio frequency oscillator may be of a form similar to the radio frequency oscillator. This system might preferably consist of an oscillator and power amplifier and should likewise be shielded.

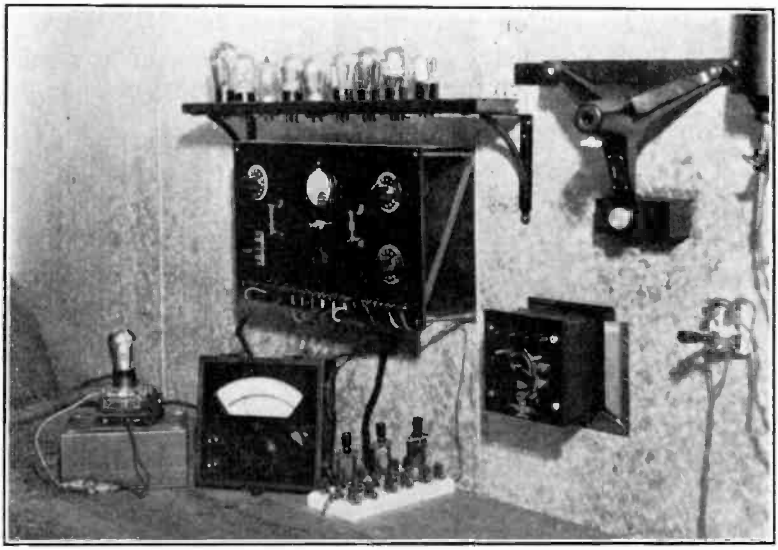

Figure 4

Figure 3 is an illustration of such a system. It is provided with an adjustable frequency range of from 32 to 32000 cycles in octave steps. This oscillator is arranged so that it has a uniform output over the entire frequency range, which of course materially assists in the speed with which measurements can be made.

The vacuum tube voltmeter is similar to that described in the Bell System Technical Journal Vol. III No. 2, Page 185. A UX 112 tube is employed and operated on the square law portion of the curve. This tube should either be debased, or else the indicated results corrected by a suitable constant to compensate for the dielectric loss in the base and socket. The output is read directly on a sensitive micro ammeter or galvanometer. The effect of the steady direct plate current is balanced out of the meter by a reverse current obtained from the heating battery and suitably adjusted by means of resistances.

Figure 4 is an illustration of a simple form of vacuum tube voltmeter and Figure 5 is the schematic circuit diagram.

Figure 4A

In order to illustrate the manner in which the measurements are made, a simple laboratory model neutrodyne receiver will be used. It has two stages of tuned radio frequency amplification.

Let us now turn back to the series of measurements previously outlined. No. 1 is a measurement of the voltage step-up in the input coupling coil. This is made by employing a dummy antenna having suitable characteristics of capacity, inductance and resistance. The dummy antenna

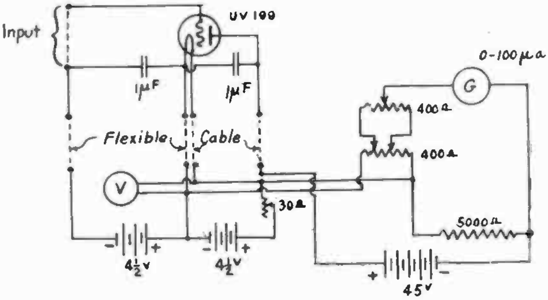

Figure 5—Vacuum Tube Voltmeter

is connected to the receiver under test. Radio frequency energy is then supplied from the oscillator by means of the output voltage attenuator. The input to the vacuum tube voltmeter is connected across the high voltage terminals of the antenna transformer and the circuit tuned to resonance. The deflection in the output meter of the voltmeter is adjusted to a suitable range by regulation of the voltage attenuator in the output of the oscillator and both the deflection in the meter and value of resistance in the voltage attenuator noted.

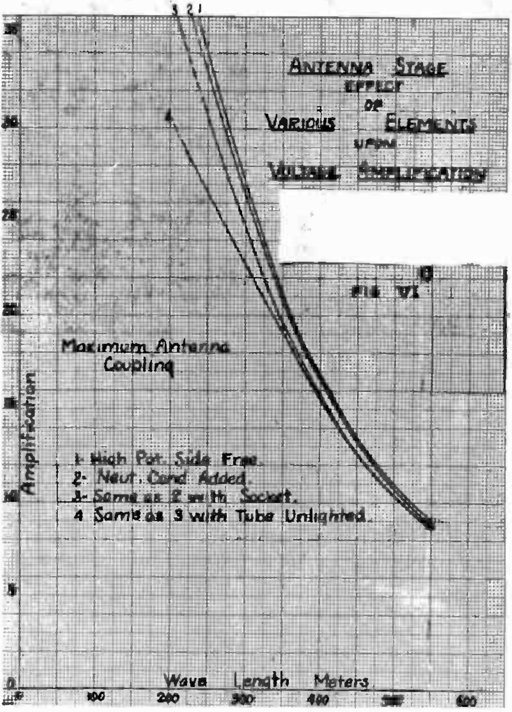

Figure 6

The expression for amplification is as follows:

$$A = \sqrt{\frac{\varphi_2}{\varphi_1}} \; \frac{R_1}{R_2}$$

where φ_2=Deflection of meter
φ_1=Calibration of meter
R_1=Total resistance of voltage attenuator
R_2=Resistance of portion used in measurement.

Figure 6 is a series of curves made on an antenna coupling transformer and shows the effect of various losses on the amplification characteristics. Curve 1 is the amplification characteristic in which the voltmeter is connected directly across the tuned circuit with other connections removed. Curve 2 is the same measurement with the neutralizing condenser added and shows the reduction in amplification due to the dielectric loss in the condenser. Curves 3 and 4 show additional losses due to the tube socket and tube base.

Figure 7 is a series of curves taken on the first stage of amplification and includes a vacuum tube with its output transformer. In this measurement the dummy antenna is removed and the signal supplied directly to the grid of the tube. Here again the various losses are well defined and consist chiefly of dielectric loss and tube loss.

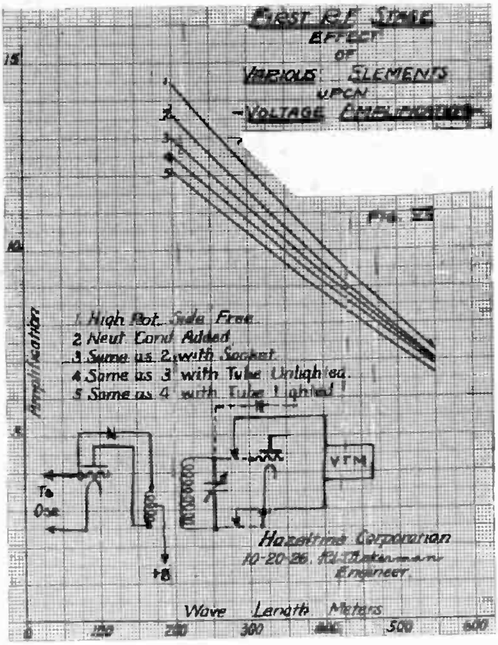

Figure 7

Figure 8 is the amplification characteristic of the second stage and is made in the same manner as the other measurements. In the case where a grid condenser and leak are employed which impose a load on the tuned circuit, the measurement is made with the detector tube in an operative condition.

Before going further it might be interesting if we examined the amplification characteristic obtained from a popular model T. R. F. receiver which depends for its stability, that is freedom from oscillation, on highly damped circuits.

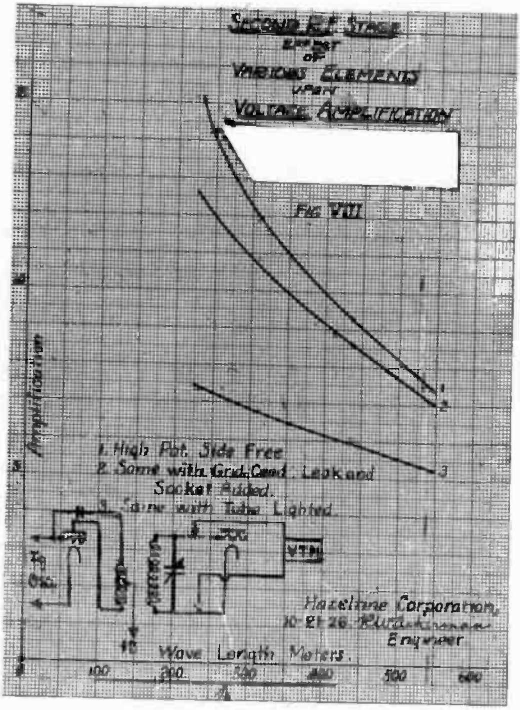

Figure 8

Figure 9 shows the amplification curves obtained from the three tunings of this receiver.

The next point of special interest is the resonance characteristics of the individual tunings. This measurement should be made at two frequencies and can be conveniently done when making the amplification measurements. The curves may be taken by tuning the signal to resonance on the receiver and the detuning the oscillator by means of the vernier adjustment which can be calibrated in kilocycles.

Figure 10 is a series of resonance curves for the various stages of the receiver under test.

Figure 11 is the resonance curve of the T. R. F. receiver previously mentioned. It is an excellent illustration of the manner in which the frequency admission band of a receiver can be altered by highly damped circuits.

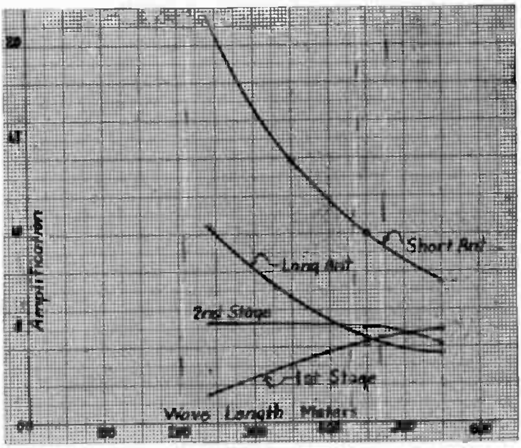

Figure 9—T. R. F. Receiver—R. F. Characteristic

From the measurements so far made the total R. F. am-

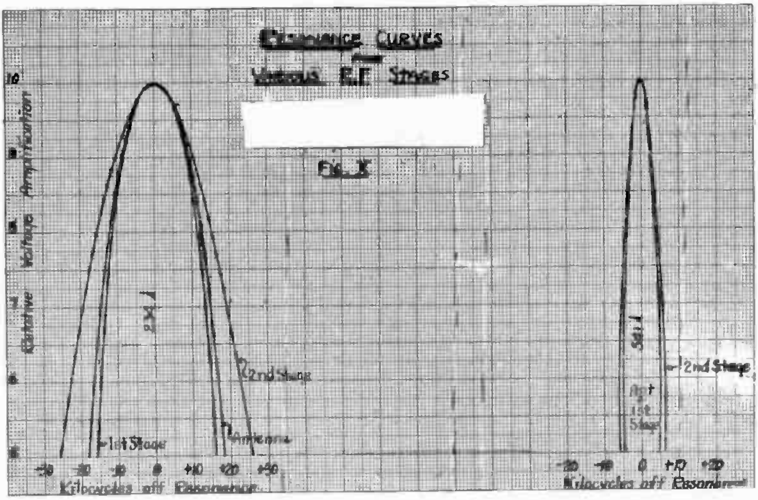

Figure 10

computed, and these are shown in Figures 12 and 13 respectively. Curves of Figures 12 and 13 show the resonance plification as well as the complete resonance curve may be characteristic and voltage amplification of the receivers previously referred to.

The next measurements are made on the audio frequency portions of the amplifier. The first involves the voltage step-up and frequency characteristic of the individual transformers. This measurement is made by connecting the primary of the transformer to the output voltage divider of the audio

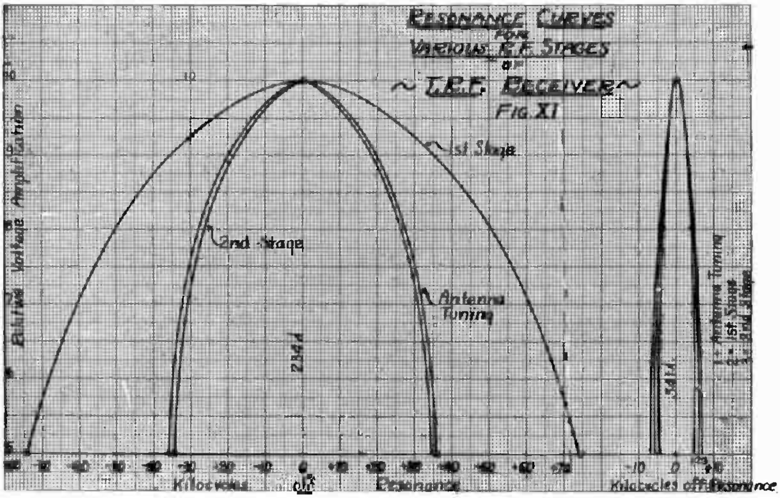

Figure 11

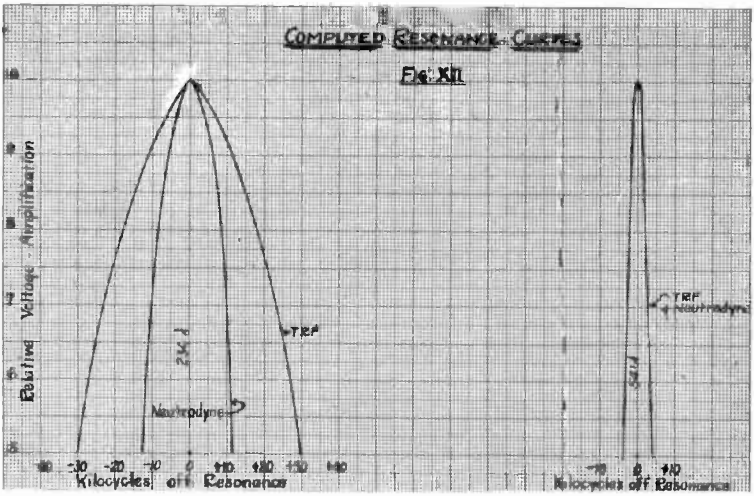

Figure 12

oscillator. The tube resistance may be simulated by a fixed resistance connected in series with the primary winding. The vacuum tube voltmeter can be connected directly across the secondary terminals of the transformer and the amplification computed as in the R. F. measurements.

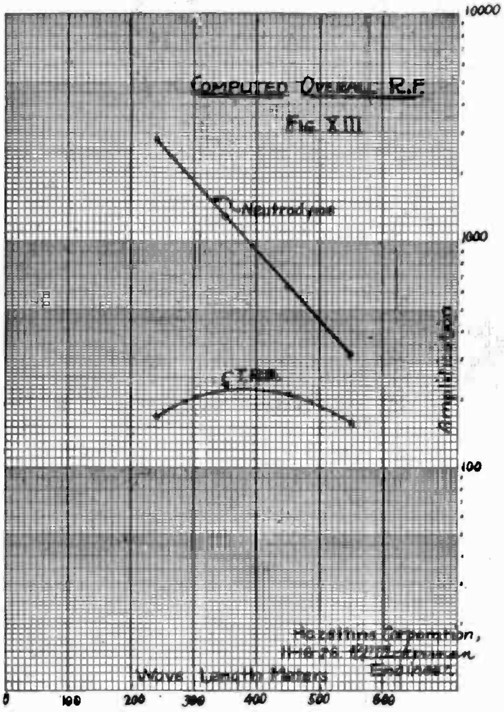

Figure 13

Figure 14 is a typical frequency characteristic of two audio transformers and shows particularly the second rise due to the effect of the leakage inductance. Curve *C* of the same figure shows the computed frequency characteristic of the complete amplifier. Unfortunately, however, this curve has little significance in a complete amplifier because of the effect of the detector grid condenser and the grid leak as well as the regenerative effects due to common battery couplings.

A relative frequency characteristic of the complete amplifier may easily be obtained, however, by supplying the A. F. oscillation to the detector tube in series with the grid leak and measuring the response between grid and filament of the last tube. This will give an amplification curve as shown in curve *A*, Figure 15 which is approximately the way a signal is heard in the loudspeaker. With regeneration present, such as due to common "B" batteries, the curve may easily be altered to that shown in curve *B* unless its effect is eliminated by suitable circuit arrangements. Curve *C* of the same figure shows the frequency characteristic of the audio system with the regeneration and second peak removed and is a good example of the present day high class audio system.

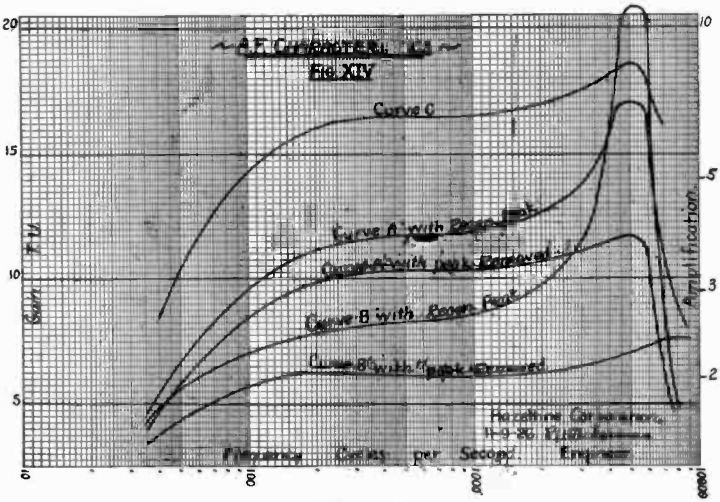

Figure 14

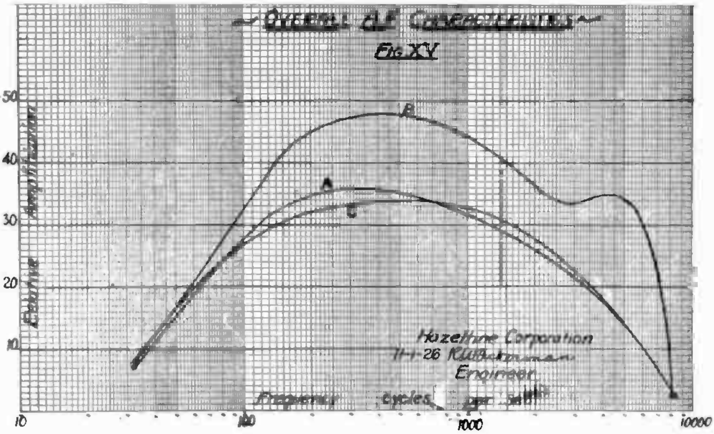

Figure 15

In analyzing the R. F. amplification and resonance curves, it must be remembered that innumerable factors enter into the results which may be obtained. Among these factors are poor transformers, excessive dielectric losses, high resistance of parts, etc. All these effects can be improved and, in many cases, entirely eliminated once their nature has been determined by suitable measurement.

The audio system responds to like treatment and as a rule can be easily corrected when the cause is known.

In conclusion, I wish to acknowledge the assistance of Mr. R. W. Ackerman, a member of the laboratory staff, in the preparation of this paper.

A THEORETICAL AND EXPERIMENTAL INVESTIGATION OF DETECTION FOR SMALL SIGNALS*

By

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Any electrically conducting device having a non-linear relation between the instantaneous values of output current and impressed voltage will act as a rectifying detector of radio-frequency signals. If the impressed voltage of the signal is a high-frequency e.m.f. of constant amplitude the detector causes a change in the steady value of the output current. If the impressed voltage is a modulated high-frequency e.m.f., *i.e.*, one the amplitude of which varies at a lower frequency, then the output of the detector contains an alternating current of modulation frequency. This process of deriving a current of modulation frequency from a modulated e.m.f. is sometimes known as *demodulation*.

The action of a rectifying detector can be expressed in terms of the impedances of the circuits and the shape of the characteristic curve or curves of the detector. In the case of a crystal detector or a two-electrode vacuum tube, often called a Fleming valve or a *diode*, the theory is very simple. When, however, a three-electrode vacuum tube or triode is used as a detector the complete theory of its action is necessarily complicated because of the larger number of variables. The detecting action can be readily determined experimentally under any prescribed condition, but the theory is of value in showing upon what variables the sensitivity of the detector depends, and also in showing under what conditions and how much distortion results.

Very little has been published on the subject of detection. The definitions of detection proposed in 1919 by Ballantine[1] and by Hazeltine[2] are both incomplete. Carson[3] in 1919 in an article called "A Theoretical Study of the Three-Element Vacuum Tube," gave a very incomplete theory of detection. In 1920,

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*Received by the Editor, September 4, 1926.

[1] Proceedings of The Institute of Radio Engineers, volume 7, page 129.

[2] Proceedings of The Institute of Radio Engineers, volume 7, page 172.

[3] Proceedings of The Institute of Radio Engineers, volume 7, page 187.

G. Breit[4] in a paper called "The Calculation of Detecting and Amplifying Properties of an Electron Tube from its Static Characteristics," presented some very general mathematical expressions for the triode from which some of the theory of detection is derived for unmodulated signals. A second paper by Hulburt and Breit[5] called "The Detecting Efficiency of the Single Electron Tube," gives experimental tests of the formulas derived in Dr. Breit's first paper for certain special cases of detection of unmodulated signals. Mr. Stuart Ballantine, in June 1924, sent to the first author of the present paper some unpublished notes on detection. Although the notes are marked unfinished by Mr. Ballantine, they contain a fairly complete treatment of a special case of grid-circuit rectification. The method of attack adopted by Ballantine is somewhat different from that used in this paper, but the results are in agreement with those here presented for the particular case treated by Ballantine. Since the completion of this paper Mr. Colebrook,[6] of England, has published a series of excellent papers dealing with the mathematical theory of detection for simple detectors and for triodes. Mr. Colebrook's method of treatment is, however, different from that used in this paper and his results differ in form from those presented here.

The present paper[7] presents the theory of detection for both two and three-electrode devices expressed in terms of the circuit impedances and the first and second partial differential coefficients of the static characteristic curves of the device taken at the points on the characteristic curves determined by the steady polarizing voltages. It is then assumed that the impressed signal is so small that for any given steady voltages these coefficients can be assumed constant within the range of the variations due to the signal voltage. A small signal voltage is defined as one less than 0.05 volts r.m.s.

Nomenclature—A brief explanation of the symbols used in this paper will be given at this point.

Small letters in general represent instantaneous values, as e and i for instantaneous e.m.f. and current, respectively. Small letters are also used to denote the slopes of curves which have

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[4] "Phys. Rev.," volume XVI, page 387.

[5] "Phys. Rev.," volume XVI, page 408.

[6] "Experimental Wireless," volume II, No. 18, 19, and 20; volume II, No. 26, 27 and volume III, No. 28 and 29.

[7] The theory and method of measuring the detection coefficient given in this paper were worked out by the first author about four years ago and have been presented since that time each year to a class in Vacuum Tubes at Harvard University

the dimensions of resistance or conductance. For example, if $i = f\,(e)$ then $k = \frac{d\,i}{d\,e}$ is the conductance for small variations.

Capital letter with a line above represents a steady value of e.m.f. or current, or a resistance for steady current.

Capital letter with a line below represents the maximum amplitude of a sinusoid.

Capital letter with no line above or below denotes the root-mean-square value of an alternating e.m.f. or current, or the value of resistance, inductance, capacity, or impedance for alternating current.

A caret (∧) *over a letter* denotes equivalent value.

A Δ *before a quantity* signifies that it is small but finite.

Subscripts in general denote location of a quantity in a circuit, or frequency, as for instance e_g is the instantaneous value of grid potential, but ω_h is the angular velocity of high frequency.

Bold-face type denotes complex and vector quantities.

I—THEORY OF TWO-TERMINAL DETECTOR

A—Unmodulated Signal

Detector With No Load. Before taking up the more complex case of the detection by a triode, the simple theory will be presented when the rectifying device is a diode, crystal detector, or electrolytic detector.

In Figure 1, T represents any two-electrode detecting device having a static characteristic given by the relation

$$i = f(e) \tag{1}$$

This function may be expressible only graphically as shown in Figure 2. $\overline{E}_B$ is a constant polarizing potential and $\Delta\, e_o = \Delta\, \underline{E}_o$ $\sin \omega_h t$ is a small impressed radio-frequency potential of frequency $\frac{\omega_h}{2\pi}$.

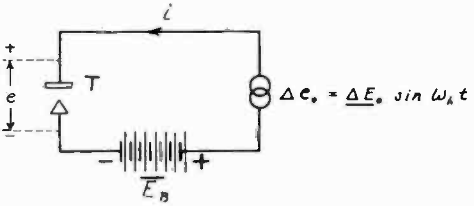

FIGURE 1

When the radio-frequency potential is not impressed, the steady current that flows is

$$\overline{I} = f(\overline{E}_B) \tag{2}$$

Now suppose Δe_o is impressed, then

$$i = f(\overline{E}_B + \underline{\Delta E}_o \sin \omega_h t)$$

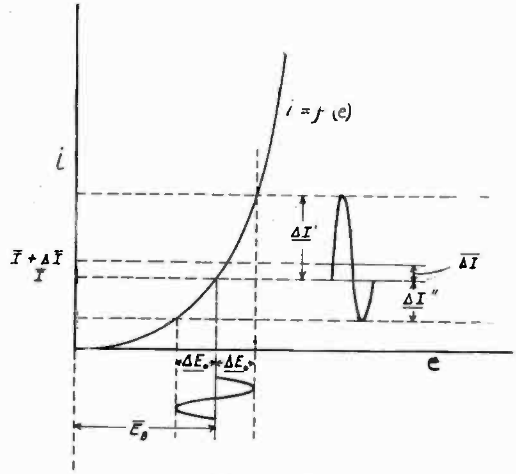

FIGURE 2

The direct current $\overline{I}$ is changed to $\overline{I} + \overline{\Delta I}$, which is the average value of the new current and given by the expression

$$\overline{I} + \overline{\Delta I} = \frac{1}{T}\int_o^T f(\overline{E}_B + \underline{\Delta E}_o \sin \omega_h t)\,dt. \tag{3}$$

Now $f(\overline{E}_B + \underline{\Delta E}_o \sin \omega_h t)$ can be developed by Taylor's Theorem, giving

$$\overline{I} + \overline{\Delta I} = \frac{1}{T}\int_o^T \left[f(\overline{E}_B) + \frac{d f(e)}{d e}\underline{\Delta E}_o \sin \omega_h t + \frac{1}{2}\frac{d^2 f(e)}{d e^2}(\underline{\Delta E}_o \sin \omega_h t)^2 + \ldots \right] dt = f(\overline{E}_B) + \frac{1}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E}_o)^2 + \ldots \tag{4}$$

Therefore, if terms of higher order than the second are neglected, the change in direct current is given by the expression

$$\overline{\Delta I} = \frac{1}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E}_o)^2 \tag{5}$$

It should be noted that in determining the increase in current, $\overline{\Delta I}$, the first and third order terms in the expansion contribute nothing since they have no average value. The change $\overline{\Delta I}$ is then given by the *second order* term. It is, therefore, more accurate to write (5) in the form

$$\overline{\Delta^2 I} = \frac{1}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E_o})^2 = (det.\ I)(\underline{\Delta E_o})^2 \qquad (5\text{-}1)$$

The factor (*det. I*), where

$$(det.\ I) = \frac{1}{4}\frac{d^2 i}{d e^2}\,\frac{amperes}{(max.\ volts)^2} \qquad (6)$$

is defined as the *current detection coefficient for the device with no load*. If equation (5-I) be multiplied by the total resistance of the circuit for a small increment in current, we have the equivalent steady voltage produced by rectification acting in the output circuit. In the case considered the total resistance is simply $r = \frac{1}{\frac{di}{de}}$ of the rectifier. Then the equivalent steady voltage is

$$\overset{\wedge}{\overline{\Delta^2 E_o}} = r(det.\ I) = \frac{r}{4}\frac{d^2 i}{d e^2}(\underline{\Delta E_o})^2 = \frac{1}{4}\,\frac{\frac{d^2 i}{d e^2}}{\frac{d i}{d e}}(\underline{\Delta E_o})^2\,\frac{volts}{(max.\ volts)^2} \qquad (7)$$

The quantity r (*det. I*) is the *voltage detection coefficient* of the *rectifier alone* and is denoted by the symbol (*det. E*).

Equation (5-1) gives the rectified current as proportional to the square of the amplitude of the impressed high-frequency alternating e.m.f. Since higher order terms than the second in Taylor's development in equation (4) have been neglected, it is thereby implicitly assumed that any *small* portion of the characteristic curve (equation (1)) of the device about the point determined by $\bar{E}_B$, can be expressed nearly enough by an equation of no higher degree than the second. This, of course, *does not* mean that the curve of equation (1) taken as a whole is expressible as a second degree equation. The assumption above puts no limitation upon the shape of the characteristic curve except that its first and second derivatives shall be everywhere finite.

It is to be noted at this point that we could have obtained the rectified current given in equation (5-1) by the following slightly different procedure. By equation (1) $i = f(e)$.
If i and e have the specific steady values $\bar{I}$ and $\bar{E}_B$, then $\bar{I} = f(\bar{E}_B)$ (see Figure 2). Let $\bar{E}_B$ be given a positive increment $\underline{\Delta E_o}$. As a consequence, $\bar{I}$ will have a positive increment $\underline{\Delta I}'$ such that

$$\bar{I} + \underline{\Delta I'} = f(\bar{E}_B + \underline{\Delta E_o}) \qquad (8)$$

Now let $\bar{E}_B$ have a negative increment $\underline{\Delta E_o}$ equal to the positive increment, and then $\bar{I}$ will have a negative increment $\underline{\Delta I''}$, such that

$$\bar{I} - \underline{\Delta I''} = f(\bar{E}_B - \underline{\Delta E_o}) \tag{9}$$

Equations (8) and (9) can be developed by Taylor's Theorem, giving

$$\bar{I} + \underline{\Delta I'} = f(\bar{E}_B) + \frac{d\,f(e)}{d\,e}\underline{\Delta E_o} + \frac{1}{2}\frac{d^2 f(e)}{d\,e^2}(\underline{\Delta E_o})^2 + \ \ldots \tag{10}$$

$$\bar{I} - \underline{\Delta I''} = f(\bar{E}_B) - \frac{d\,f(e)}{d\,e}\underline{\Delta E_o} + \frac{1}{2}\frac{d^2 f(e)}{d\,e^2}(\underline{\Delta E_o})^2 - \ \ldots \tag{11}$$

Adding equations (10) and (11) gives

$$\underline{\Delta I'} - \underline{\Delta I''} = \frac{d^2 f(e)}{d\,e^2}(\underline{\Delta E_o})^2 = \frac{d^2 i}{d\,e^2}(\underline{\Delta E_o})^2 \tag{12}$$

The difference $\underline{\Delta I'} - \underline{\Delta I''}$ is a quantity of the second order and must be proportional to the average current $\overline{\Delta^2 I}$ that is produced by the rectifying action of the device. Comparing equations (5-1) and (12) it is evident that *if the impressed potential is sinusoidal*, then

$$\frac{\underline{\Delta I'} - \underline{\Delta I''}}{4} = \overline{\Delta^2 I} \tag{13}$$

Equation (13) shows that the rectified current can be obtained from the two series developments of (10) and (11) instead of by integrating the series as is done in equation (4). This alternative method of obtaining $\overline{\Delta^2 I}$ will be made use of later.

Two-Terminal Detector in Series with Impedance. We may now consider a more general case when, as shown in Figure 3, an impedance Z is included in the circuit. This impedance Z may have a value $\bar{Z}$ for direct currents and a different value Z_h for a frequency $\frac{\omega_h}{2\pi}$. The detector has for small variations of potential a resistance of

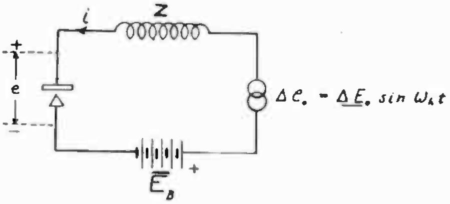

FIGURE 3

$$r = \frac{1}{\frac{d\,f(e)}{d\,e}} = \frac{1}{\frac{d\,i}{d\,e}} \tag{14}$$

Referring to Figure 3, it is obvious that when Δe_o is impressed a current of radio frequency $\frac{\omega_h}{2\pi}$ flows through the circuit, and in addition the steady current is altered by an amount to be denoted by $\overline{\Delta^2 I}$. When no alternating potential is impressed $\bar{I} = f(\bar{E})$, where $\bar{E}$ is the steady potential across the rectifier and different from $\bar{E}_B$. Refer now to Figure 4. The characteristic curve for the rectifier is shown as $i = f\,(e)$. From point $\bar{E}_B$, a point laid off on the voltage axis equal to the steady polarizing-battery potential, a line is drawn making an angle with the vertical, the tangent of which is equal to the resistance $\bar{R}$ of Z to steady currents. (For simplicity in representation by diagram, Z is taken as a pure resistance, the same for steady and alternating currents, but the mathematical derivation does not require this limitation.) The point P of intersection of this resistance line and the curve gives the values of $\bar{I}$ and $\bar{E}$. If now $\Delta e_o = \underline{\Delta E_o}\ sin\ \omega_h t$ is superposed upon $\bar{E}_B$, the new intersection points given by lines drawn parallel to the resistance line from points representing the extreme

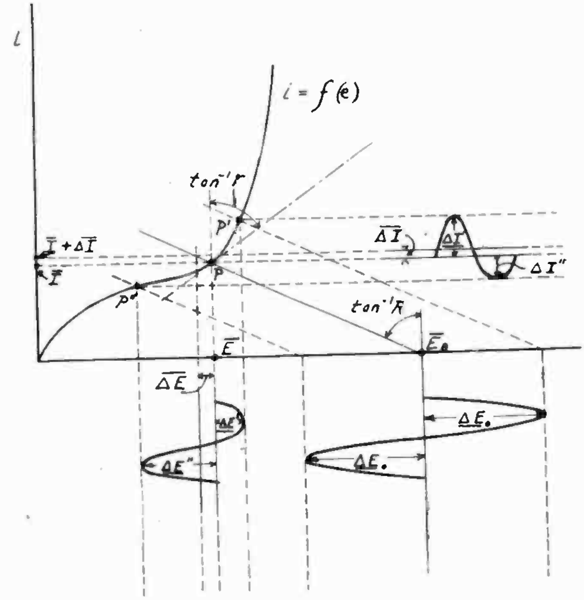

FIGURE 4

applied voltages $\bar{E}_B+\underline{\Delta E_o}$ and $\bar{E}_B-\underline{\Delta E_o}$ give the extreme currents $\bar{I}+\underline{\Delta I'}$ and $\bar{I}-\underline{\Delta I''}$, and the increments of voltage on the rectifier, *viz.*, $\underline{\Delta E'}$ and $\underline{\Delta E''}$. We can now write

$$\begin{aligned}\bar{I}+\underline{\Delta I'}&=f(\bar{E}+\underline{\Delta E'})\\&=f(\bar{E})+\frac{d\,f(e)}{d\,e}\underline{\Delta E'}+\frac{1}{2}\frac{d^2 f(e)}{d\,e^2}(\underline{\Delta E'})^2+\dots\end{aligned}\tag{15}$$

and

$$\begin{aligned}\bar{I}-\underline{\Delta I''}&=f(\bar{E}-\underline{\Delta E''})\\&=f(\bar{E})-\frac{d\,f(e)}{d\,e}\underline{\Delta E''}+\frac{1}{2}\frac{d^2 f(e)}{d\,e^2}(\underline{\Delta E''})^2-\dots\end{aligned}\tag{16}$$

Adding (15) and (16), we have

$$\begin{aligned}\underline{\Delta I'}-\underline{\Delta I''}=&\frac{d\,f(e)}{d\,e}(\underline{\Delta E'}-\underline{\Delta E''})+\\&\frac{d^2 f(e)}{d\,e^2}\left(\frac{(\underline{\Delta E'})^2+(\underline{\Delta E''})^2}{2}\right)+\dots\end{aligned}\tag{17}$$

Each term of (17) is of the second order in magnitude. By analogy with (13), we can express the differences of increments as steady components of current and voltage. The last term of (17) involves the average of the squares of the increments of voltage and is represented by $(\underline{\Delta E})^2$. Equation (17) can then be written

$$\overline{\Delta^2 I}=\frac{d\,i}{d\,e}\overline{\Delta^2 E}+\frac{1}{4}\frac{d^2\,i}{d\,e^2}(\underline{\Delta E})^2+\dots\dots\tag{18}$$

$\overline{\Delta^2 E}$ is a steady component of voltage across the rectifier and is equal to $-\bar{R}\,\overline{\Delta^2 I}$. Further, $\underline{\Delta E}$ is the peak radio voltage across the detector and is related to $\underline{\Delta E_o}$ by the obvious relation

$$\underline{\Delta E}=\frac{r}{\sqrt{(r+R_h)^2+X_h^2}}\underline{\Delta E_o}\tag{18-1}$$

where R_h and X_h are, respectively, the resistance and reactance of the series load Z at the frequency $\frac{\omega_h}{2\pi}$. Then (18) can be written

$$\overline{\Delta^2 I}=-\frac{\bar{R}\,\overline{\Delta^2 I}}{r}+\frac{r^2}{4[(r+R_h)^2+X_h^2]}\cdot\frac{d^2\,i}{d\,e^2}(\underline{\Delta E_o})^2\tag{19}$$

or

$$\begin{aligned}\overline{\Delta^2 I}&=\frac{r}{r+\bar{R}}\cdot\frac{r^2}{[(r+R_h)^2+X_h^2]}\cdot\frac{1}{4}\cdot\frac{d^2\,i}{d\,e^2}(\underline{\Delta E_o})^2\\&=(Det.\ I)(\underline{\Delta E_o})^2\end{aligned}\tag{20}$$

Hence the *complete current detection coefficient* for the system is

$$(Det.\ I)=\frac{r}{r+\overline{R}}\cdot\frac{r^2}{(r+R_h)^2+X_h{}^2}\cdot\frac{1}{4}\frac{d^2\,i}{d\,e^2} \qquad (21)$$

Capital D is used to distinguish the complete detection coefficient from the detection coefficient for the detector alone signified by small d in (*det. I*).

The equivalent voltage of rectification acting in the circuit is

$$\overset{\wedge}{\overline{\Delta^2 E}}=\frac{r^3}{(r+R_h)^2+X_h{}^2}\cdot\frac{1}{4}\cdot\frac{d^2\,i}{d\,e^2}(\underline{\Delta E_o})^2, \qquad (22)$$

and the *equivalent voltage detection coefficient* for this case is

$$(Det.\ E)=\frac{r^3}{(r+R_h)^2+X_h{}^2}\cdot\frac{1}{4}\frac{d^2\,i}{d\,e^2} \qquad (22\text{-}1)$$

It is evident from equation (21) and (22-1) that the complete detection coefficient is greater the smaller the radio resistance and reactance, R_h and X_h. If R_h and X_h are negligible, then

$$\overline{\Delta^2 I}=\frac{r}{r+\overline{R}}\cdot\frac{1}{4}\cdot\frac{d^2\,i}{d\,e^2}(\underline{\Delta E_o})^2 \qquad (22\text{-}2)$$

and the rectified power is

$$(\overline{\Delta^2 I})^2\,\overline{R}=\frac{r^2\,\overline{R}}{16(r+\overline{R})^2}\cdot\left(\frac{d^2\,i}{d\,e^2}\right)^2(\underline{\Delta E_o})^4 \qquad (23)$$

The value of R which makes this a maximum is

$$\overline{R}=r \qquad (24)$$

This stated in words is as follows:—If there is no impedance offered to the alternating current to be rectified other than that of the rectifier itself, then the maximum rectified power for small values of the impressed e.m.f. is obtained when the resistance of the load or absorbing device for the rectified current is equal to r, the variational resistance of the rectifier.

If the series load is a pure resistance which has the same value for both alternating and direct currents, then the rectified power is

$$(\overline{\Delta^2 I})^2\,R=\frac{r^6\,R}{16(r+R)^6}\left(\frac{d^2\,i}{d\,e^2}\right)(\underline{\Delta E_o})^4 \qquad (25)$$

Solving equation (25) to find the value of R to give maximum power gives

$$R=\frac{r}{5} \qquad (26)$$

B—Detection of Modulated Signal

Case 1. Series Load a Pure Resistance to the Modulation Frequency. In radio communication the detector usually rectifies a modulated radio-frequency e.m.f. giving rise to a current having a frequency equal to the modulation frequency. Figure 5 represents a sinuosoidally modulated radio-frequency e.m.f., which is expressed mathematically as follows:

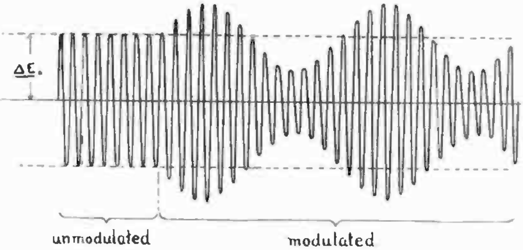

FIGURE 5

$$\Delta e_o = \underline{\Delta E}_o(1 + m \sin \omega_l t) \sin \omega_h t \qquad (27)$$

In this expression $\underline{\Delta E}_o$ is the average amplitude of the alternating e.m.f. or the amplitude of the unmodulated radio-frequency e.m.f. of angular velocity ω_h; ω_l is the angular velocity corresponding to the modulation frequency; and m is a factor ranging from 0 for no modulation to 1 for complete modulation, and is a factor giving the fraction of complete modulation and known as the *degree of modulation.*

Suppose now that the modulated e.m.f. of expression (27) is impressed on a rectifying system containing a pure resistance load and having a current detection coefficient (*Det. I*). Then in place of the $(\underline{\Delta E}_o)^2$ of equation (20) must be substituted the square of the coefficient of $\sin \omega_h t$ in equation (27), giving

$$\begin{aligned} \Delta^2 i &= (Det.\ I)(\underline{\Delta E}_o)^2(1 + m \sin \omega_l t)^2 \\ &= (Det.\ I)(\underline{\Delta E}_o)^2(1 + 2\, m \sin \omega_l t + m^2 \sin^2 \omega_l t) \\ &= (Det.\ I)(\underline{\Delta E}_o)^2\left(1 + \frac{m^2}{2} + 2\, m \sin \omega_l t - \frac{m^2}{2} \cos 2\, \omega_l t\right) \qquad (28) \end{aligned}$$

In the above expression the $\overline{\Delta^2 I}$ of equation (20) is changed to $\Delta^2 i$, to signify that now the rectified component is periodic and has an instantaneous value $\Delta^2 i$. Equation (28) shows that when a modulated e.m.f. is rectified by a system having a curved characteristic and terms of higher order than the second are neglected, three rectified components are obtained as follows:

1. *Direct current* $= \overline{\Delta^2 I} = \left(1 + \frac{m^2}{2}\right)(Det.\ I)(\underline{\Delta E_o})^2 \qquad (29)$

2. *Current of modulation frequency having effective value*

$$= \Delta^2 I_l = \sqrt{2}\, m (Det.\ I)(\underline{\Delta E_o})^2 \tag{30}$$

3. *Current of double modulation frequency having effective value*

$$= \Delta^2 I_{2l} = \frac{m^2}{2\sqrt{2}} (Det.\ I)(\underline{\Delta E_o})^2 \tag{31}$$

In the above special case the complete detection coefficient is the same in all three expressions, 29, 30, and 31. This will not be the case when the load is other than a pure resistance.

It may now be assumed that instead of but one frequency of modulation, there are *two*, as denoted by the following equation which replaces (27).

$$\Delta e_o = \underline{\Delta E_o}(1 + m' \sin \omega_{l'} t + m'' \sin \omega_{l''} t) \sin \omega_h t \tag{31-1}$$

The resulting current in a rectifying system having a complete detection coefficient of (*Det. I*), due to this doubly-modulated high-frequency potential of (31-1) contains the following components as is evident on substituting the square of the amplitude of $\sin \omega_h t$ of (31-1) in equation (20).

1. *Direct current*

$$= \overline{\Delta^2 I} = \left(1 + \frac{m'^2}{2} + \frac{m''^2}{2}\right)(Det.\ I)(\underline{\Delta E_o})^2 \tag{31-2}$$

2. *Current of first modulation frequency of effective value*

$$= \Delta^2 I_{l'} = \sqrt{2}\, m' (Det.\ I)(\underline{\Delta E_o})^2 \tag{31-3}$$

3. *Current of second modulation frequency of effective value*

$$= \Delta^2 I_{l''} = \sqrt{2}\, m'' (Det.\ I)(\underline{\Delta E_o})^2 \tag{31-4}$$

4. *Current of double first modulation frequency of effective value*

$$= \Delta^2 I_{2l'} = \frac{m'^2}{2\sqrt{2}} (Det.\ I)(\underline{\Delta E_o})^2 \tag{31-5}$$

5. *Current of double second modulation frequency of effective value*

$$= \Delta^2 I_{2l''} = \frac{m''^2}{2\sqrt{2}} (Det.\ I)(\underline{\Delta E_o})^2 \tag{31-6}$$

6. *Current of frequency equal to difference of modulation frequencies of effective value*

$$= \Delta^2 I_{(l'-l'')} = \frac{m'\, m''}{\sqrt{2}} (Det.\ I)(\underline{\Delta E_o})^2 \tag{31-7}$$

7. *Current of frequency equal to sum of modulation frequencies of effective value*

$$= \Delta^2 I_{(l'+l'')} = \frac{m'\, m''}{\sqrt{2}}\,(Det.\ I)\,(\underline{\Delta E_o})^2 \qquad (31\text{-}8)$$

Equations (29), (30), and (31) show that a rectifying detector used to demodulate a sinusoidally modulated high-frequency signal inevitably introduces a double modulation frequency component which is small in comparison with the main component of modulation frequency only when the modulation is small.

Equations (31-2) to (31-8) further show that if there is more than one frequency of modulation there are introduced beside the main terms of the modulation frequencies and their second harmonics, currents of frequencies equal to both the sum and difference of each pair of modulation frequencies. These sum and difference tones are extremely undesirable as they are dissonant tones. They can be made small only by making the modulation coefficients small.

Case 2—Series Load Not a Pure Resistance at Modulation Frequency. In the preceding treatment of Case 1 the impedance Z of Figure 3 was restricted to a pure resistance for the audio or rectified current, but was not restricted for the modulated or radio-frequency current. The treatment will now be extended to the case where Z may have reactance for the rectified current. We must return to equations (18) and (18-1), both of which hold with no change for the present case. Substituting (18-1) in (18) gives

$$\overline{\Delta^2 I} = \frac{\overline{\Delta^2 E}}{r} + \frac{r^2}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{de^2}(\underline{\Delta E_o})^2 \qquad (32)$$

Now suppose that $(\underline{\Delta E_o})$ has the significance given to it by equation (27). Then $(\underline{\Delta E_o})$ must be replaced by $\underline{\Delta E_o}\,(1+m\ sin\ \omega_l t)$ in (32), which then becomes

$$\Delta^2 i_l = \frac{\Delta^2 e_l}{r} + \frac{r^2}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d\,e^2}(\underline{\Delta E_o})^2(1+m\ sin\ \omega_l t)^2$$

$$= \frac{\Delta^2 e_l}{r} + \frac{r^2}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d\,e^2}(\underline{\Delta E_o})^2 \left(1+\frac{m^2}{2}+2\,m\ sin\ \omega_l t-\frac{m^2}{2}\ cos\ 2\,\omega_l t\right) \qquad (33)$$

In equation (33) $\Delta^2 i_l$ is now the total instantaneous value of the rectified current which is made up of the constant-current component and the currents of frequencies $\frac{\omega_l}{2\pi}$ and $\frac{2\,\omega_l}{2\pi}$. $\Delta^2 e_l$ is the total instantaneous rectified voltage across the rectifying device, and also comprises the instantaneous values of the same three rectified conponents as for $\Delta^2 i_l$. Equation (33) can then be broken up into three equations, the first giving the value of the constant portion of the rectified current, the second the maximum amplitude of the rectified current of modulation frequency $\frac{\omega_l}{2\pi}$, and the third the maximum amplitude of the rectified current of double modulation frequency. These three equations follow:

$$\overline{\Delta^2 I} = \frac{\overline{\Delta^2 E}}{r} + \frac{r^2}{(r+R_h)^2+X_h{}^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E}_o)^2 \left(1+\frac{m^2}{2}\right) \qquad (34)$$

$$\begin{aligned} \underline{\Delta^2 I}_l \, sin(\omega_l t + \alpha_l) &= \frac{\underline{\Delta^2 E}_l \, sin\,(\omega_l t + \beta_l)}{r} \\ &\quad + \frac{r}{(r+R_h)^2+X_h{}^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E}_o)^2 \, 2\, m \, sin\, \omega_l t \qquad (35) \end{aligned}$$

$$\begin{aligned} \underline{\Delta^2 I}_{2l} \, cos\,(2\,\omega_l t + \alpha_{2l}) &= \frac{\underline{\Delta^2 E}_{2l} \, cos\,(2\,\omega_l t + \beta_{2l})}{r} \\ &\quad + \frac{r^2}{(r+R_h)^2+X_h{}^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} (\underline{\Delta E}_o)^2 \frac{m^2}{2} cos\, 2\,\omega_l t \qquad (36) \end{aligned}$$

Here the phase angles α and β are determined by r and the impedance of the load.

Since the voltage across the rectifying device is the negative of the voltage across the impedance Z, equations (34), (35), and (36) can be rewritten, the last two being expressed in complex notation for root-mean-square values. Complex quantities are signified by bold-face type.

$$\begin{aligned} \overline{\Delta^2 I} &= \frac{r}{r+\overline{R}} \cdot \frac{r^2}{(r+R_h)^2+X_h{}^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} \cdot \left(1+\frac{m^2}{2}\right) (\underline{\Delta E}_o)^2 \\ &= (\overline{Det.\ I}) \left(1+\frac{m^2}{2}\right) (\underline{\Delta E}_o)^2 \qquad (37) \end{aligned}$$

$$\begin{aligned} \Delta^2 \boldsymbol{I}_l &= \frac{r}{r+\boldsymbol{Z}_l} \cdot \frac{r^2}{(r+R_h)^2+X_h{}^2} \cdot \frac{1}{4} \frac{d^2 i}{d e^2} \cdot \sqrt{2}\, m\, (\underline{\Delta E}_o)^2 \\ &= (\boldsymbol{Det.\ I})_l \sqrt{2}\, m (\underline{\Delta E}_o)^2 \qquad (38) \end{aligned}$$

$$\begin{aligned} \Delta^2 \boldsymbol{I}_{2l} &= \frac{r}{r+\boldsymbol{Z}_{2l}} \cdot \frac{r^2}{(r+R_h)^2+X_h{}^2} \cdot \frac{1}{4} \cdot \frac{d^2 i}{d e^2} \cdot \frac{m}{2\sqrt{2}} (\underline{\Delta E}_o)^2 \\ &= (\boldsymbol{Det.\ I})_{2l} \frac{m}{2\sqrt{2}} (\underline{\Delta E}_o)^2 \qquad (39) \end{aligned}$$

Obviously in the above equation $(\boldsymbol{Det.\ I})_l$ and $(\boldsymbol{Det.\ I})_{2l}$ are complex quantities and different because of the difference in frequency. Considerable advantage arises in expressing the equivalent voltage-detection coefficient, for as will be seen in this case, this quantity is *independent of frequency* and *is not a complex quantity*. Equations (37), (38), and (39) are rewritten, expressed in terms of equivalent rectified voltage, as follows:

$$\overset{\wedge}{\overline{\Delta^2 E}} = (r+\overline{R})\overline{\Delta^2 I} = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\cdot\frac{d^2 i}{d e^2}\left(1+\frac{m^2}{2}\right)(\underline{\Delta E_o})^2$$

$$= (Det.\ E)\left(1+\frac{m^2}{2}\right)(\underline{\Delta E_o})^2 \qquad (37\text{-}1)$$

$$\overset{\wedge}{\Delta}{}^2 E_l = (r+\mathbf{Z}_l)\Delta^2 \mathbf{I}_l = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d e^2}\sqrt{2}\,m(\underline{\Delta E_o})^2$$

$$= (Det.\ E)\sqrt{2}\,m(\underline{\Delta E_o})^2 \qquad (38\text{-}1)$$

$$\overset{\wedge}{\Delta}{}^2 E_{2l} = (r+\mathbf{Z}_{2l})\Delta^2 \mathbf{I}_{2l} = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d e^2}\cdot\frac{m^2}{2\sqrt{2}}(\underline{\Delta E_o})^2$$

$$= (Det.\ E)\frac{m^2}{2\sqrt{2}}(\underline{\Delta E_o})^2 \qquad (39\text{-}1)$$

In the above expressions $(Det.\ E)$ is the same for all and is a real quantity given by the following equation:

$$(Det.\ E) = \frac{r^3}{(r+R_h)^2+X_h^2}\cdot\frac{1}{4}\frac{d^2 i}{d e^2} \qquad (40)$$

The complex e.m.fs. of (38-1) and (39-1) do not actually exist, but when used as though they exist give the correct actual values of the rectified currents. The phase of the e.m.f. given by (38-1) is the same as that of the modulation envelope or the modulating e.m.f. The phase of the e.m.f. given by (39-1) is the same as the phase of the double-frequency term in the square of equation (27).

It is obvious that the theory of this case can be extended as in Case 1 to give the magnitude of the difference and sum components where there are two or more frequencies of modulation.

II—THEORY OF DETECTION OF A TRIODE

The method used above may now be extended to the three-electrode vacuum tube used as a detector. The connections are shown in Figure 6.

The instantaneous plate current i_p is a function of the instantaneous plate potential e_p and the instantaneous grid potential e_g, and is expressed by the functional equation:

$$i_p = \phi(e_p,\ e_g) \qquad (41)$$

Similarly the grid current i_g is dependent upon the instantaneous grid potential and to a slight degree upon e_p, but, to a fair degree of approximation, particularly if the grid is not polarized greatly positively, the grid current can be expressed as a function only of e_g, thus:

$$i_g = \psi(e_g) \tag{42}$$

Unmodulated Signal. We will at first assume that the small unmodulated alternating e.m.f. $\Delta e_o = \underline{\Delta E}_o \, sin \; \omega_h t$ is impressed in the grid circuit of the detector of Figure 6. Because both characteristic equations (41) and (42) are curves, the changes of grid and plate voltage, due to the impressed e.m.f. Δe_o, will be made up of steady rectified components as well as variations at frequency $\frac{\omega_h}{2\pi}$. We may then designate the increases of e_g and e_p due to $+\underline{\Delta E}_o$ as $\underline{\Delta E}_g'$ and $\underline{\Delta E}_p'$, respectively, and the decreases in the same quantities as $\underline{\Delta E}_g''$ and $\underline{\Delta E}_p''$, respectively. Using Taylor's Theorem we then have for the increment of plate current,

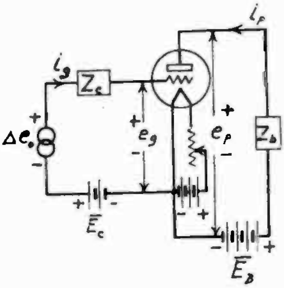

FIGURE 6

$$\begin{aligned}\overline{I}_p + \underline{\Delta I}_p' &= \phi(\overline{E}_p + \underline{\Delta E}_p', \; \overline{E}_g + \underline{\Delta E}_g') \\ &= \phi(\overline{E}_p, \overline{E}_g) + \frac{\partial i_p}{\partial e_p}\underline{\Delta E}_p' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p^2}(\underline{\Delta E}_p')^2 + \dots\dots \\ &\quad + \frac{\partial i_p}{\partial e_g}\underline{\Delta E}_g' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_g^2}(\underline{\Delta E}_g')^2 + \dots\dots \\ &\quad + \frac{\partial^2 i_p}{\partial e_p \partial e_g}\underline{\Delta E}_p' \; \underline{\Delta E}_g' + \dots\dots\end{aligned} \tag{43}$$

Similarly the decrease in plate current is given by

$$\begin{aligned}I_p - \underline{\Delta I}_p'' &= \phi(\overline{E}_p - \underline{\Delta E}_p'', \; \overline{E}_g - \underline{\Delta E}_g'') \\ &= \phi(\overline{E}_p, \overline{E}_g) - \frac{\partial i_p}{\partial e_p}\underline{\Delta E}_p'' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p^2}(\underline{\Delta E}_p'')^2 - \dots\dots \\ &\quad - \frac{\partial i_p}{\partial e_g}\underline{\Delta E}_g'' + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_g^2}(\underline{\Delta E}_g'')^2 - \dots\dots \\ &\quad + \frac{\partial^2 i_p}{\partial e_p \partial e_g}\underline{\Delta E}_p'' \; \underline{\Delta E}_g'' + \dots\dots\end{aligned} \tag{44}$$

Adding equations (43) and (44) we have

$$\underline{\Delta I}_p{}' - \underline{\Delta I}_p{}'' = \frac{\partial i_p}{\partial e_p}(\underline{\Delta E}_p{}' - \underline{\Delta E}_p{}'') + \frac{\partial i_p}{\partial e_g}(\underline{\Delta E}_g{}' - \underline{\Delta E}_g{}'')$$
$$+ \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p{}^2}[(\underline{\Delta E}_p{}')^2 + (\underline{\Delta E}_p{}'')^2] + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_g{}^2}[(\underline{\Delta E}_g{}')^2$$
$$+ (\underline{\Delta E}_g{}'')^2] + \frac{\partial^2 i_p}{\partial e_p \partial e_g}(\underline{\Delta E}_p{}' \, \underline{\Delta E}_g{}'$$
$$+ \underline{\Delta E}_p{}'' \, \underline{\Delta E}_g{}'') + \dots\dots \tag{45}$$

In equations (43), (44), and (45), higher order terms are neglected as being in most cases too small to be important.

Equation (45) can be more simply expressed, remembering that the first three terms are expressible as constant values by equation (13), and that the average of the squares of the increments of the grid and plate voltages may be replaced by single expressions as $(\underline{\Delta E}_p)^2 = \dfrac{(\underline{\Delta E}_p{}')^2 + (\underline{\Delta E}_p{}'')^2}{2}$, etc.

$$\overline{\Delta^2 I}_p = \frac{\partial i_p}{\partial e_p}\overline{\Delta^2 E}_p + \frac{\partial i_p}{\partial e_g}\overline{\Delta^2 E}_g + \frac{1}{4}\frac{\partial^2 i_p}{\partial e_p{}^2}(\underline{\Delta E}_p)^2$$
$$+ \frac{1}{4}\frac{\partial^2 i_p}{\partial e_p{}^2}(\underline{\Delta E}_g)^2 + \frac{1}{2}\frac{\partial^2 i_p}{\partial e_p \partial e_g}\underline{\Delta E}_p \, \underline{\Delta E}_g + \dots\dots \tag{46}$$

A further simplification of expression (46) is afforded by the well-known definitions of the tube coefficients:

$$\frac{\partial i_p}{\partial e_p} = \textit{plate-to-filament variational conductance}$$
$$= k_p = \frac{1}{r_p} \tag{47}$$

$$\frac{\partial i_p}{\partial e_g} = \textit{grid-plate variational conductance} = \sigma_p. \tag{48}$$

Rewriting (46), we have

$$\overline{\Delta^2 I}_p = \frac{\overline{\Delta^2 E}_p}{r_p} + \sigma_p \overline{\Delta^2 E}_g + \frac{1}{4}\frac{\partial k_p}{\partial e_p}(\underline{\Delta E}_p)^2$$
$$+ \frac{1}{4}\frac{\partial \sigma_p}{\partial e_g}(\underline{\Delta E}_g)^2 + \frac{1}{2}\frac{\partial \sigma_p}{\partial e_p}\underline{\Delta E}_p \, \underline{\Delta E}_g + \dots\dots \tag{49}$$

In equation (49) $\underline{\Delta E}_p$ is the maximum amplitude of the radio-frequency alternating potential between the plate and filament

Now the plate current of frequency $\dfrac{\omega_h}{2\pi}$ is:

$$\underline{\Delta I}_p = \frac{\mu_p \underline{\Delta E}_g}{\sqrt{(r_p + R_{bh})^2 + X_{bh}{}^2}} \tag{50}$$

where μ_p is the amplification factor of the tube and equal to $-\dfrac{\partial e_p}{\partial e_g}$, and R_{bh} and X_{bh} are respectively the resistance and reactance of the plate impedance Z_b at frequency $\dfrac{\omega_h}{2\pi}$. From (50) we can get $\underline{\Delta E}_p$ by multiplying by $-Z_{bh}$, giving

$$\underline{\Delta E}_p = -\frac{\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\cdot\underline{\Delta E}_g \tag{51}$$

Substituting (51) in (49) gives

$$\overline{\Delta^2 I_p} = \frac{\overline{\Delta^2 E_p}}{r_p} + \sigma_p\overline{\Delta^2 E_g} + \left[\frac{1}{4}\frac{\partial\sigma_g}{\partial e_g} + \frac{1}{4}\frac{\mu_p^2(R_{bh}^2+X_{bh}^2)}{(r_p+R_{bh})^2+X_{bh}^2}\cdot\frac{\partial k_p}{\partial e_p}\right.$$
$$\left. - \frac{\mu_p\sqrt{R_{bh}^2+X_{bh}^2}}{2\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\cdot\frac{\partial\sigma_p}{\partial e_p}\right](\underline{\Delta E}_g)^2 \tag{52}$$

The values of $\overline{\Delta^2 E_g}$ and $\underline{\Delta E}_g$ depend upon $\underline{\Delta E}_o$ and the characteristics of the grid circuit. Following the same method for the increments in grid current as for the plate circuit we have

$$\begin{aligned}\bar{I}_g+\underline{\Delta I}_g' &= \psi(\bar{E}_g+\underline{\Delta E}_g')\\ &= \psi(\bar{E}_g)+\frac{\partial i_g}{\partial e_g}\underline{\Delta E}_g'+\frac{1}{2}\frac{\partial^2 i_g}{\partial e_g^2}(\underline{\Delta E}_g')^2+\ldots\ldots\end{aligned} \tag{53}$$

and

$$\begin{aligned}\bar{I}_g-\Delta I_g'' &= \psi(\bar{E}_g-\underline{\Delta E}_g'')\\ &= \psi(\bar{E}_g)-\frac{\partial i_g}{\partial e_g}\underline{\Delta E}_g''+\frac{1}{2}\frac{\partial^2 i_g}{\partial e_g^2}(\underline{\Delta E}_g'')^2+\ldots\ldots\end{aligned} \tag{54}$$

Adding (53) and (54) gives

$$\begin{aligned}\underline{\Delta I}_g'-\underline{\Delta I}_g'' = \frac{\partial i_g}{\partial e_g}(\underline{\Delta E}_g'-\underline{\Delta E}_g'')+\frac{1}{2}\frac{\partial^2 i_g}{\partial e_g^2}[(\underline{\Delta E}_g')^2\\ +(\underline{\Delta E}_g'')^2]+\ldots\ldots\end{aligned} \tag{55}$$

Converting (55) by equation (13) and replacing $\dfrac{\partial i_g}{\partial e_g}$ by the grid variational conductance $k_g=\dfrac{1}{r_g}$, we have

$$\overline{\Delta^2 I_g} = \frac{\overline{\Delta^2 E_g}}{r_g}+\frac{1}{4}\frac{\partial k_g}{\partial e_g}(\underline{\Delta E}_g)^2+\ldots\ldots \tag{56}$$

We now wish to express the maximum amplitude of the alternating e.m.f. $\underline{\Delta E}_g$ between grid and filament in terms of $\underline{\Delta E}_o$, the amplitude of the impressed e.m.f. The amplitude of the high-frequency alternating current flowing in the grid circuit is

$$\underline{\Delta I}_g = \frac{\underline{\Delta E}_o}{\sqrt{(R_{ch}+r_{gh})^2+(X_{ch}+x_{gh})^2}} \tag{57}$$

where R_{ch} and X_{ch} are respectively the resistance and reactance to frequency $\frac{\omega_h}{2\pi}$ of the series impedance Z_c in the grid circuit, and $\hat{r}_{gh}$ and $\hat{x}_{gh}$ are, respectively, the equivalent grid input resistance and reactance of the vacuum tube at frequency $\frac{\omega_h}{2\pi}$. (We have up to this point neglected any effects caused by capacity between the tube elements. It is well known, however, that the tube capacities, notably the capacity between the grid and plate, give the tube an equivalent input resistance different from r_g and an input reactance which may be either positive or negative, dependent upon the load in the plate circuit.) But

$$\underline{\Delta E}_g = \underline{\Delta I}_g \sqrt{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}$$
$$= \frac{\sqrt{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}}{\sqrt{(R_{ch}+\hat{r}_{gh})^2 + (X_{ch}+\hat{x}_{gh})^2}} \cdot \underline{\Delta E}_o \tag{58}$$

Equation (58) is now to be substituted into (52) and (56), and we have

$$\overline{\Delta^2 I_p} = \frac{\overline{\Delta^2 E_p}}{r_p} + \sigma_p \overline{\Delta^2 E_g} + \left[\frac{1}{4} \frac{\partial \sigma_p}{\partial e_g} + \frac{\mu_p^2 (R_{bh}^2 + X_{bh}^2)}{4[(r_p+R_{bh})^2 + X_{bh}^2]} \cdot \frac{\partial k_p}{\partial e_p} - \frac{\mu_p \sqrt{R_{bh}^2 + X_{bh}^2}}{2\sqrt{(r_p+R_{bh})^2 + X_{bh}^2}} \cdot \frac{\partial \sigma_p}{\partial e_p} \right] \frac{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}{(R_{ch}+\hat{r}_{gh})^2 + (X_{ch}+\hat{x}_{gh})^2} \cdot (\underline{\Delta E}_o)^2 \tag{59}$$

and

$$\overline{\Delta^2 I_g} = \frac{\overline{\Delta^2 E_g}}{r_g} + \frac{1}{4} \frac{\partial k_g}{\partial e_g} \cdot \frac{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}{(R_{ch}+\hat{r}_{gh})^2 + (X_{ch}+\hat{x}_{gh})^2} (\underline{\Delta E}_o)^2 \tag{60}$$

As an abbreviation let the square bracket of (59) be denoted by A

$$A = \frac{1}{4} \frac{\partial \sigma_p}{\partial e_g} + \frac{\mu_p^2 (R_{bh}^2 + X_{bh}^2)}{4[(r_p+R_{bh})^2 + X_{bh}^2]} \cdot \frac{\partial k_p}{\partial e_p} - \frac{\mu_p \sqrt{R_{bh}^2 + X_{bh}^2}}{2\sqrt{(r_p+R_{bh})^2 + X_{bh}^2}} \cdot \frac{\partial \sigma_p}{\partial e_p}, \tag{61}$$

and let B stand for

$$B = \frac{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}{(R_{ch}+\hat{r}_{gh})^2 + (X_{ch}+\hat{x}_{gh})^2} \tag{62}$$

Modulated Signal. So far we have assumed that the impressed alternating e.m.f. Δe_o has a constant amplitude $\underline{\Delta E}_o$. Now, following the method previously adopted, we will suppose that the impressed e.m.f. of frequency $\frac{\omega_h}{2\pi}$ is modulated at a frequency $\frac{\omega_l}{2\pi}$ as indicated by equation (27). The maximum amplitude $\underline{\Delta E}_o$, assumed constant in the preceding equations, is now a function of t and the value $\underline{\Delta E}_o$ is to be replaced in equations (59) and (60) by

$$\underline{\Delta E}_o(1+m\,sin\,\omega_l t)$$

Because $\underline{\Delta E}_o$ in these equations is squared, three component terms result, one of constant amplitude, one of frequency $\frac{\omega_l}{2\pi}$, and the third of double this frequency. The quantities $\overline{\Delta^2 I_p}$, $\overline{\Delta^2 I_g}$, $\overline{\Delta^2 E_p}$, $\overline{\Delta^2 E_g}$ must now be considered functions of time and each will be made up of three components. With this in mind we may again rewrite (59) and (60) as follows:

$$\overline{\Delta^2 I_p}+\underline{\Delta I}_{pl}\,sin(\omega_l t+\alpha_l)-\underline{\Delta I}_{p2l}\,cos(2\,\omega_l t+\alpha_{2l})$$
$$=\frac{\overline{\Delta^2 E_p}+\underline{\Delta E}_{pl}\,sin(\omega_l t+\beta_l)-\underline{\Delta E}_{b2l}\,cos\,(2\omega_l t+\beta_l)}{r_p}$$
$$+\sigma_p(\overline{\Delta^2 E_g}+\underline{\Delta E}_{g_l})sin(\omega_l t+\gamma_l)-\underline{\Delta E}_{g2l}\,cos(2\,\omega_l t+\gamma_{2l})$$
$$+A\,B\left(1+\frac{m^2}{2}+2\,m\,sin\,\omega_l t-\frac{m^2}{2}cos\,2\,\omega_l t\right)(\underline{\Delta E}_o)^2 \tag{63}$$

$$\overline{\Delta^2 I_g}+\underline{\Delta I}_{g_l}\,sin(\omega_l t+\theta_l)-\underline{\Delta I}_{g2l}\,cos(2\omega_l t+\theta_{2l})$$
$$=\frac{\overline{\Delta^2 E_g}+\underline{\Delta E}_{gl}\,sin(\omega_l t+\theta_l)-\underline{\Delta E}_{p2l}\,cos(2\,\omega_l t+\theta_{2l})}{r_g}$$
$$+\frac{B}{4}\frac{\partial k_g}{\partial e_g}\left(1+\frac{m^2}{2}+2\,m\,sin\,\omega_l t-\frac{m^2}{2}cos\,2\,\omega_l t\right)(\underline{\Delta E}_o)^2 \tag{64}$$

Each of equations (61) and (62) now breaks up into three equations, one for each component. If now these component equations be expressed in effective complex values we have:

$$\left.\begin{aligned}\overline{\Delta^2 I_p}&=\frac{\overline{\Delta^2 E_p}}{r_p}+\sigma_p\overline{\Delta^2 E_g}+A\,B\left(1+\frac{m^2}{2}\right)(\underline{\Delta E}_o)^2 && (63\text{-}1)\\ \overline{\Delta^2 I_g}&=\frac{\overline{\Delta^2 E_g}}{r_g}+\frac{B}{4}\frac{\partial k_g}{\partial e_g}\left(1+\frac{m^2}{2}\right)(\underline{\Delta E}_o)^2 && (63\text{-}2)\end{aligned}\right\}\ \textit{constant term}$$

$$\left.\begin{aligned}\Delta^2 I_{pl}&=\frac{\Delta^2 E_{pl}}{r_p}+\sigma_p\,\Delta^2 E_{gl}+A\,B\sqrt{2}\,m(\underline{\Delta E}_o)^2 && (64\text{-}1)\\ \Delta^2 I_{gl}&=\frac{\Delta^2 E_{gl}}{r_g}+\frac{B}{4}\frac{\partial k_g}{\partial e_g}\sqrt{2}\,m(\underline{\Delta E}_o)^2 && (64\text{-}2)\end{aligned}\right\}\ \textit{modulation frequency}$$

$$\left.\begin{aligned}
\Delta^2 \boldsymbol{I}_{p2l} &= \frac{\Delta^2 \boldsymbol{E}_{p2l}}{r_p} + \sigma_p \underline{\Delta^2 \boldsymbol{E}}_{g2l} + A\,B\frac{m^2}{2\sqrt{2}}(\underline{\Delta E}_o)^2 \qquad (65\text{-}1)\\
\Delta^2 \boldsymbol{I}_{g2l} &= \frac{\Delta^2 \boldsymbol{E}_{g2l}}{r_g} + \frac{B}{4}\frac{\partial k_g}{\partial e_g}\frac{m^2}{2\sqrt{2}}(\underline{\Delta E}_o)^2 \qquad (65\text{-}2)
\end{aligned}\right\}\ \textit{double modulation frequency}$$

The two equations for each component can now be combined and simplified by means of the following relations:

$$\overline{\Delta^2 E_{pl}} = -\overline{R}_b\,\overline{\Delta^2 E_p} \qquad \overline{\Delta^2 E_g} = -\overline{R}_c\,\overline{\Delta^2 I_g} \tag{66}$$

$$\Delta^2 \boldsymbol{E}_{pl} = -\boldsymbol{Z}_{bl}\,\Delta^2 \boldsymbol{I}_{pl} \qquad \Delta^2 \boldsymbol{E}_{gl} = -\boldsymbol{Z}_{cl}\,\Delta^2 \boldsymbol{I}_{gl} \tag{67}$$

$$\Delta^2 \boldsymbol{E}_{p2l} = -\boldsymbol{Z}_{b2l}\,\Delta^2 \boldsymbol{I}_{p2l} \qquad \Delta^2 \boldsymbol{E}_{g2l} = -\boldsymbol{Z}_{c2l}\,\Delta^2 \boldsymbol{I}_{g2l} \tag{68}$$

giving the final expression for the three rectified components as follows:

$$\begin{aligned}
\overline{\Delta^2 I_p} &= \frac{r_p}{r_p+\overline{R}_b}\left[-\frac{\sigma_p r_g \overline{R}_c}{4(r_g+\overline{R}_c)}\cdot\frac{\partial k_g}{\partial e_g}+A\right]B\left(1+\frac{m^2}{2}\right)(\underline{\Delta E}_o)^2\\
&= (\overline{Det.\ I})\left(1+\frac{m^2}{2}\right)(\underline{\Delta E}_o{}^2)
\end{aligned}\tag{69}$$

$$\begin{aligned}
\Delta^2 \boldsymbol{I}_{pl} &= \frac{r_p}{r_p+\boldsymbol{Z}_{bl}}\left[-\frac{\sigma_p r_g \boldsymbol{Z}_{cl}}{4(r_g+\boldsymbol{Z}_{cl})}\cdot\frac{\partial k_g}{\partial e_g}+A\right]B\sqrt{2}\,m(\underline{\Delta E}_o)^2\\
&= (\boldsymbol{Det.\ I})_l\sqrt{2}\,m(\underline{\Delta E}_o)^2
\end{aligned}\tag{70}$$

$$\begin{aligned}
\Delta^2 \boldsymbol{I}_{p2l} &= \frac{r_p}{r_p+\boldsymbol{Z}_{b2l}}\left[-\frac{\sigma_p r_g \boldsymbol{Z}_{c2l}}{4(r_g+\boldsymbol{Z}_{c2l})}\cdot\frac{\partial k_g}{\partial e_g}+A\right]B\frac{m^2}{2\sqrt{2}}(\underline{\Delta E}_o)^2\\
&= (\boldsymbol{Det.\ I})_{2l}\frac{m^2}{2\sqrt{2}}(\underline{\Delta E}_o)^2
\end{aligned}\tag{71}$$

The values of the equivalent current and voltage detection coefficients are listed below for convenient reference.

$$\left.\begin{aligned}
(\overline{Det.\ I}) &= \frac{r_p}{4(r_p+\overline{R}_b)}\left[-\frac{r_g\overline{R}_c}{r_g+\overline{R}_c}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}\right.\\
&\qquad +\frac{R_{bh}{}^2+X_{bh}{}^2}{(r_p+R_{bh})^2+X_{bh}{}^2}\mu_p{}^2\frac{\partial k_p}{\partial e_p}\\
&\qquad \left.-\frac{2\sqrt{R_{bh}{}^2+X_{bh}{}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}{}^2}}\mu_p\frac{\partial \sigma_p}{\partial e_p}\right]\\
&\qquad \left[\frac{\hat{r}_{gh}{}^2+\hat{x}_{gh}{}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]\\
(\overline{\boldsymbol{Det.\ E}}) &= (r_p+\overline{R}_b)(\overline{Det.\ I})
\end{aligned}\right\}\ \textit{steady component} \qquad (72)$$

$$\left.\begin{aligned}
(\boldsymbol{Det.\,I})_l &= \frac{r_p}{4(r_p+\mathbf{Z}_{bl})}\Bigg[-\frac{r_g\mathbf{Z}_{cl}}{r_g+\mathbf{Z}_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}\\
&\quad+\frac{R_{bh}^2+X_{bh}^2}{(r_p+R_{bh})^2+X_{bh}^2}\mu_p^2\frac{\partial k_p}{\partial e_p}\\
&\quad-\frac{2\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\mu_p\frac{\partial \sigma_p}{\partial e_p}\Bigg]\\
&\quad\left[\frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]\\
(\boldsymbol{Det.\,E})_l &= (r_p+\mathbf{Z}_{bl})(\boldsymbol{Det.\,I})_l
\end{aligned}\right\}\begin{array}{c}\textit{component}\\ \textit{of modula-}\\ \textit{tion}\\ \textit{freqency}\end{array}\quad(73)$$

$$\left.\begin{aligned}
(\boldsymbol{Det.\,I})_{2l} &= \frac{r_p}{4(r_p+\mathbf{Z}_{b2l})}\Bigg[-\frac{r_g\mathbf{Z}_{c2l}}{r_g+\mathbf{Z}_{c2l}}\sigma_p\frac{\partial k_g}{\partial \epsilon_g}\\
&\quad+\frac{\partial \sigma_p}{\partial e_g}+\frac{R_{bh}^2+X_{bh}^2}{(r_p+R_{bh})^2+X_{bh}^2}\mu_p^2\frac{\partial k_p}{\partial e_p}\\
&\quad-\frac{\sqrt{R_{bh}^2+X_{bh}^2}}{\sqrt{(r_p+R_{bh})^2+X_{bh}^2}}\mu_p\frac{\partial \sigma_p}{\partial e_p}\Bigg].\\
&\quad\left[\frac{\hat{r}_{gh}^2+\hat{x}_{gh}^2}{(\hat{r}_{gh}+R_{ch})^2+(\hat{x}_{gh}+X_{ch})^2}\right]\\
(\boldsymbol{Det.\,E})_{2l} &= (r_p+\mathbf{Z}_{b2l})(\boldsymbol{Det.\,I})_{2l}
\end{aligned}\right\}\begin{array}{c}\textit{component}\\ \textit{of double}\\ \textit{modulation}\\ \textit{frequency}\end{array}\quad(74)$$

Equations (72), (73), and (74) give the final and complete values, to the approximation allowed by the assumptions made, of the change in plate current and the currents of modulation frequency and of double modulation frequency when the impressed signal is very small. It is evident that the equivalent current and voltage detection coefficients $(\boldsymbol{Det.\,I})_l$ $(\boldsymbol{Det.\,I})_{2l}$, $(\boldsymbol{Det.\,E})_l$, and $(\boldsymbol{Det.\,E})_{2l}$ are, in the case of the triode, all complex quantities. Attention is called to the fact that the value of $(\boldsymbol{Det.\,I})$ depends upon the load in the plate circuit, whereas the voltage detection coefficient $(\boldsymbol{Det.\,E})$ is independent of the plate load. $(\boldsymbol{Det.\,E})$ gives the equivalent voltage of audio frequency which is considered to be introduced into the plate circuit by detection and from which the current in the plate circuit can be calculated when the total impedance of the plate circuit is known. The voltage detection coefficient is, therefore, a better quantity to express the detecting ability of a triode.

It is evident that the theory can be easily extended to give the amplitudes of the sum and difference components when there is more than one frequency of modulation.

A very simple transformation makes somewhat easier the interpretation of the equations (69) to (74). Since $\sigma_p = \mu_p k_p$, then

$$\frac{\partial \sigma_p}{\partial e_p} = \mu_p \frac{\partial k_p}{\partial e_p} + k_p \frac{\partial \mu_p}{\partial e_p} \tag{75}$$

This relation can now be substituted in equations (69) to (74). This simplification, however, is expressed only in the case of equation (70), for this equation is the one of greatest interest. The other equations can, by comparison with the following new form for (70), be easily changed.

$$\begin{aligned}\Delta^2 \boldsymbol{I}_{pl} = \frac{r_p}{4(r_p + \boldsymbol{Z}_{bl})}\Bigg[& - \frac{r_g \boldsymbol{Z}_{cl}}{r_g + \boldsymbol{Z}_{cl}} \sigma_p \frac{\partial k_g}{\partial e_g} + \frac{\partial \sigma_p}{\partial e_g} \\ & - \frac{2\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}} \sigma_p \frac{\partial \mu_p}{\partial e_p} - \frac{\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}} \\ & \left(2 - \frac{\sqrt{R_{bh}^2 + X_{bh}^2}}{\sqrt{(r_p + R_{bh})^2 + X_{bh}^2}}\right) \mu_p^2 \frac{\partial k_p}{\partial e_p}\Bigg] \cdot \\ & \left[\frac{\hat{r}_{gh}^2 + \hat{x}_{gh}^2}{(\hat{r}_{gh} + R_{ch})^2 + (\hat{x}_{gh} + X_{ch})^2}\right] \sqrt{2}\, m(\underline{\Delta E_o})^2 \end{aligned} \tag{76}$$

An examination of equation (73) or the transformed form given by equation (76), giving the audio current of modulation frequency, shows that detection or demodulation by a triode is a very complex process depending upon the audio and radio impedances in both the grid and place circuits as well as upon the tube coefficients and several of their derivatives. It is impossible to form any idea of the magnitude of the several terms of the expression without knowing the values of the tube coefficients and their derivatives. For a hard tube the grid-plate conductance σ_p is usually positive, but its derivative $\frac{\partial \sigma_p}{\partial e_g}$ may be either positive or negative. The second term of equation (76) is positive for the part of the plate current curve which is concave upward and negative for the part which is concave downward. This second term is the most important one which contributes to plate-circuit rectification. k_g and its derivative with respect to e_g are usually positive, so that the first term is usually negative. This is the principal term when a grid leak and stoppage condenser or the equivalent are used. The other terms are usually of less importance than the first two and are in general negative because of the negative signs before them. The relative magnitudes of the several terms will be given in a special numerical case. If a gas tube is used the relative magnitudes and signs of the several terms may be very different from those for a hard tube. A more complete discussion of the equations will be given when the experimental results are considered.

When a detector is used in practice some of the impedances included in the general expression (76) may be negligible or absent. At this point, therefore, several special but practical cases of (76) will be examined. In order to make the study more definite the various characteristics of a certain typical triode which contribute to the detection coefficient were experimentally determined and are here presented. The tube used is a Cunningham type 301-A. Figure 7 is a plot of the plate and grid currents. Figures 8, 9, 10, and 11 are, respectively, the values of μ_p, k_p, σ_p, and k_g, of the tube. The values of the derivatives of these curves were obtained graphically. Evidence of a trace of gas is apparent in the irregularities in the course of most of the curves.

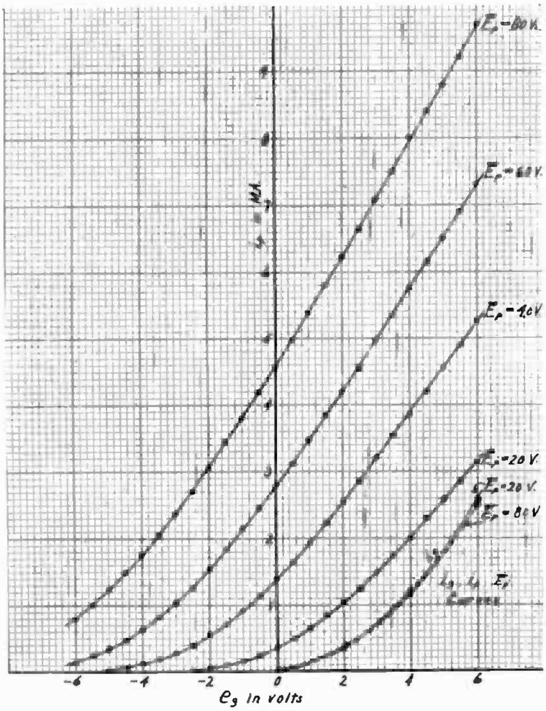

Figure 7—Plate Current and Grid Current of Cunningham 391-A Tube

Case 1. Simple Plate Circuit Rectification. To examine the simplest case of rectification we may assume that there are no high-frequency impedances in the plate circuit and no low-frequency impedances in the grid circuit. Figure 12 illustrates this case where it is to be noted that C_b is large enough to be

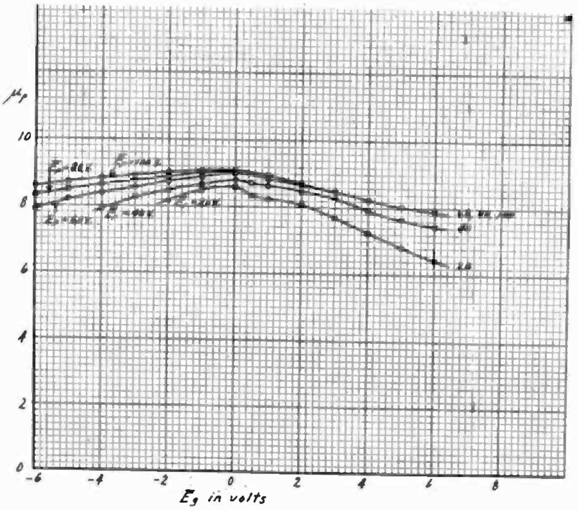

FIGURE 8—Amplification Factor of Cunningham 301-A Tube

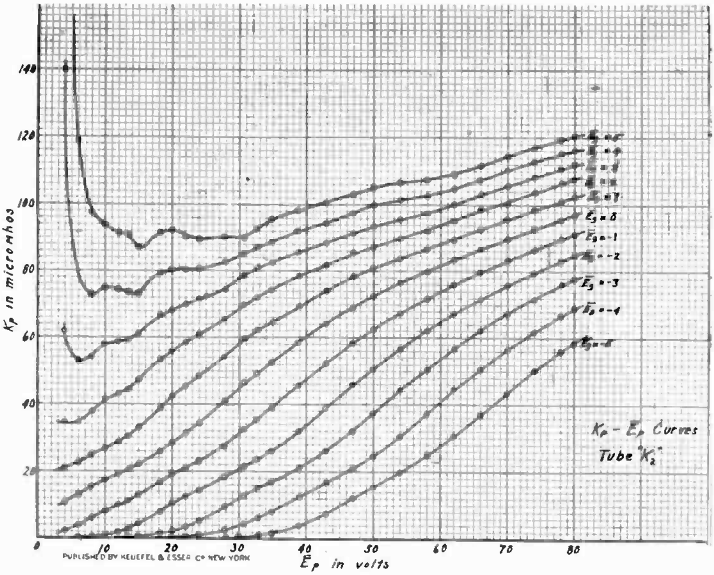

FIGURE 9—Plate Conductance of Cunningham 301-A Tube

of negligible impedance in comparison with r_p to the high-frequency current. Expressions (72)–(74) now reduce to the following:

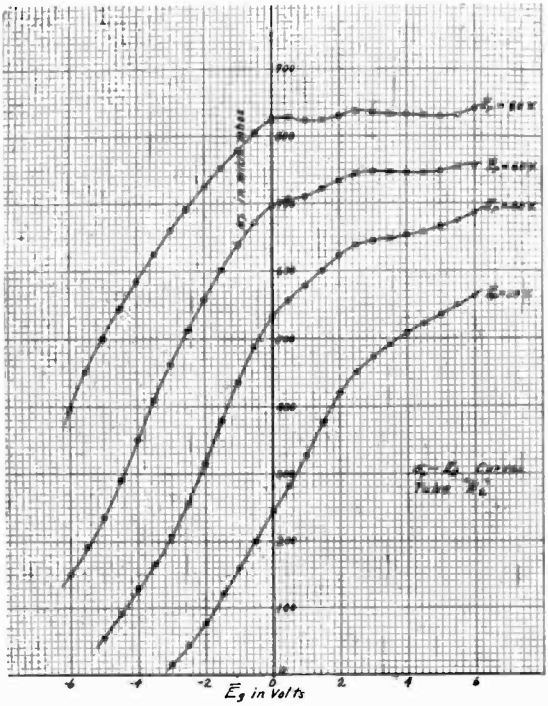

FIGURE 10—Grid-Plate Conductance of Cunningham 301-A Tube

$$
\left.
\begin{aligned}
(\overline{Det.\ I}) &= \frac{r_p}{r_p+\overline{R}_b}\cdot\frac{1}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\\
(\overline{Det.\ E}) &= \frac{r_p}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}
\end{aligned}
\right\}
\qquad (77)
$$

$$
\left.
\begin{aligned}
(\mathbf{Det.\ I})_l &= \frac{r_p}{r_p+\mathbf{Z}_{bl}}\cdot\frac{1}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\\
(Det.\ E)_l &= \frac{r_p}{4}\cdot\frac{\partial\,\sigma_p}{\partial\,e_g}
\end{aligned}
\right\}
\qquad (78)
$$

$$
\left.
\begin{aligned}
(\mathbf{Det.\ I})_{2l} &= \frac{r_p}{r_p+\mathbf{Z}_{b2l}}\cdot\frac{1}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}\\
(Det.\ E)_{2l} &= \frac{r_p}{4}\frac{\partial\,\sigma_p}{\partial\,e_g}
\end{aligned}
\right\}
\qquad (79)
$$

The factor B (see (62)) is practically unity, because R_{ch} and X_{ch} are usually small.

An examination of equations (77)–(79) indicates that in this simple case detection depends upon the value of $\frac{\partial \sigma_p}{\partial e_g}$ or $\frac{\partial^2 i_p}{\partial e_g^2}$. Figure 13 gives the values of $\frac{\partial \sigma_p}{\partial e_g}$ derived from Figure 10. The irregularities in the σ_p curves due to a trace of gas are much more pronounced in the curves of $\frac{\partial \sigma_p}{\partial e_g}$.

Case 2. Plate Circuit Rectification with Tickler Coil. This case differs from Case 1 only in that a radio impedance is added in the plate circuit in the form of a tickler coil. As an approximation, the radio impedance may be considered purely reactive of value $L_p \omega_h$.

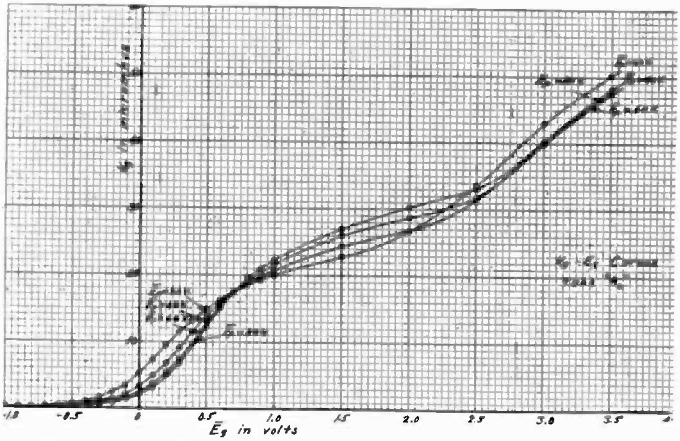

FIGURE 11—Grid Conductance of Cunningham 301-A Tube

The coupling of L_p to the grid circuit reduces the effective resistance of the oscillatory circuit, thereby increasing the value of

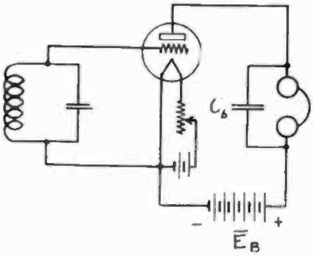

FIGURE 12

the applied voltage $\underline{\Delta E_o}$, and in addition, introduces a resistance component into the plate circuit. The value of the resistance added to the plate circuit is usually small in comparison with r_p and is here neglected so that *the coupling does not change the value of the detection coefficient but increases the impressed signal voltage.*

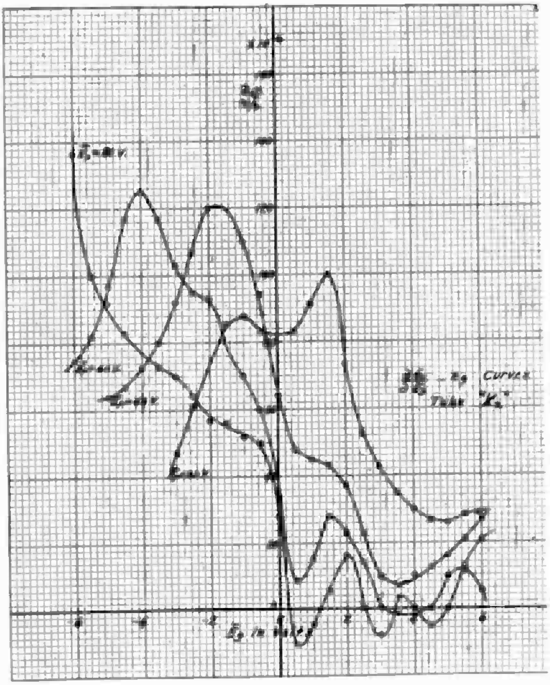

FIGURE 13—$\dfrac{\partial \sigma_p}{\partial e_o}$ for a Cunningham 301-A Tube

Equation (76) is in this case the best expression to use and reduces to the following form

$$(\boldsymbol{Det.\ I})_l = \frac{r_p}{4(r_p+Z_{bl})}\left[\frac{\partial \sigma_p}{\partial e_g} - 2\frac{L_p\,\omega_h}{\sqrt{r_p^2+L_p^2\,\omega_h^2}}\sigma_p\frac{\partial \mu_p}{\partial e_p} - \frac{L_p\,\omega_h}{\sqrt{r_p^2+L_p^2\,\omega_h^2}}\left(2-\frac{L_p\,\omega_h}{\sqrt{r_p^2+L_p^2\,\omega_h^2}}\right)\mu_p^2\frac{\partial k_p}{\partial e_p}\right] \quad (80)$$

Here again B (see (62)) is assumed to differ little from unity.

We will now examine the relative magnitudes of the three terms of (80) for the particular tube tested. The values of $\dfrac{\partial \sigma_p}{\partial e_g}$ are shown in Figure 13. The magnitude of $\sigma_p\dfrac{\partial \mu_p}{\partial e_p}$ plotted against grid voltage is given in Figure 14, which shows that the second term is of negligible importance for negative grid voltages. The values of the detection coefficient for positive grid voltages greater than about one are of little interest because, although the magnitude of the detection coefficient may be large for these voltages, the input resistance of the grid circuit is so low that the applied voltage $\underline{\Delta E_o}$ across the oscillatory circuit is much reduced. Figure 15, which gives the values of $\mu_p{}^2 \frac{\partial k_p}{\partial e_p}$, shows that for negative grid potentials $\mu_p{}^2 \frac{\partial k_p}{\partial e_p}$ is of the same order of magnitude as $\frac{\partial \sigma_p}{\partial e_p}$, and because of the negative sign before this third term, operates to reduce the detection coefficient. Evidently then a radio-frequency impedance in the plate circuit of a detector

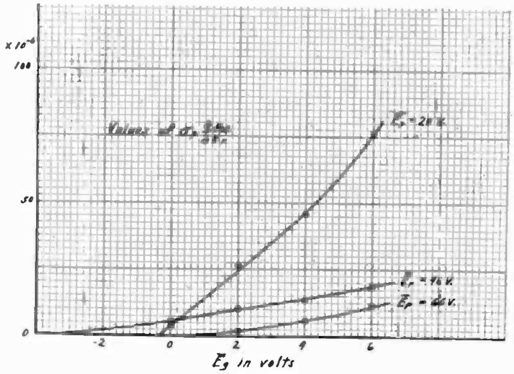

FIGURE 14—$\sigma, \frac{\partial \mu_p}{\partial e_p}$ for a Cunningham 301-A Tube

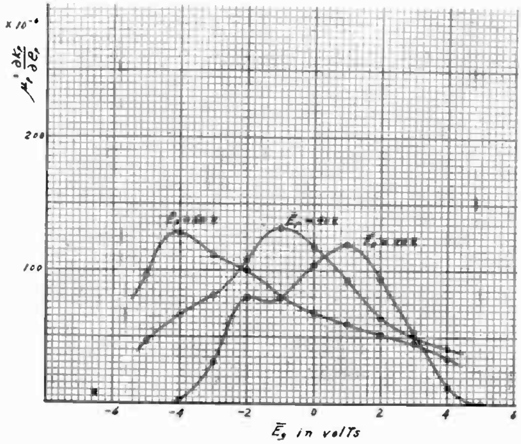

FIGURE 15—$\mu_p{}^2 \frac{\partial k_p}{\partial e_p}$ for a Cunningham 301-A Tube

operating with plate circuit rectification reduces the effective detection coefficient.

Case 3. Simple Grid-Circuit Rectification. This case, illustrated in Figure 16, applies to the most common use of a triode as a detector. The grid leak R_1 is shunted by the grid condenser C_1. R_1 is usually of the order of from 1 to 10 megohms and C_1 from .00005 to .00025 microfarads. The equivalent series resistance and reactance in the grid circuit are

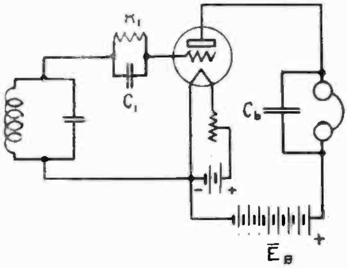

FIGURE 16

$$R_c = R_1 \frac{1}{R_1^2 C_1^2 \omega^2 + 1}$$

$$X_c = -\frac{1}{C_1 \omega} \cdot \frac{R_1^2 C_1^2 \omega^2}{R_1^2 C_1^2 \omega^2 + 1} \qquad (81)$$

At radio frequencies the value of R_{ch} is negligible, and $X_{ch} = \frac{-1}{C_1 \omega_h}$, is also usually negligible compared with r_g.

At audio frequencies R_{cl} and X_{cl} have values which vary much with frequency dependent upon the value of $R_1^2 C_1^2 \omega_l^2$ as compared with unity. The table below gives the values of this quantity for different frequencies n when $R_1 = 1$ megohm and $C_1 = .0001$ μf, and also when $R_1 = 2$ megohms and $C_1 = .0002$ μf. The value of $R_1^2 C_1^2 \omega_l^2$ can readily be deduced from the table for other values of R_1 or C_1.

$R_1^2 C_1^2 \omega_l^2$

n	for $R_1 = 10^6$ ohms $C_1 = .0001$ $\mu.f.$	for $R_1 = 2 \times 10^6$ ohms $C_1 = .0002$ $\mu.f.$
100	.00395	.0631
500	.0988	1.58
1000	.395	6.31
2000	1.58	25.3
4000	6.32	101.
8000	25.3	405.
10000	39.5	631

Equation (73) reduces in the case now considered to the following:

$$\left.\begin{aligned}(\boldsymbol{Det.\ I})_l &= \frac{r_p}{4(r_p+\boldsymbol{Z}_{tl})}\left[-\frac{r_g\boldsymbol{Z}_{cl}}{r_g+\boldsymbol{Z}_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_g}\right]\\ (\boldsymbol{Det.\ E})_l &= \frac{r_p}{4}\left[-\frac{r_g\boldsymbol{Z}_{cl}}{r_g+\boldsymbol{Z}_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}+\frac{\partial \sigma_p}{\partial e_p}\right]\end{aligned}\right\} \quad (82)$$

Factor B is again assumed equal to unity.

Evidently when $\frac{\partial \sigma_p}{\partial e_g}$ is positive the second term deducts from the effect of the first term, but since the maximum value of the first term is usually more than ten times greater than the maximum value of the second term, the latter is of little importance. Let us assume that the plate voltage is so adjusted that $\frac{\partial \sigma_p}{\partial e_g}$ is small so that it may be neglected. Then (82) becomes

$$\left.\begin{aligned}(\boldsymbol{Det.\ I})_l &= -\frac{r_p}{4(r_p+\boldsymbol{Z}_{bl})}\cdot\frac{r_g\boldsymbol{Z}_{cl}}{r_g+\boldsymbol{Z}_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}\\ (\boldsymbol{Det.\ E})_l &= -\frac{r_p r_g\boldsymbol{Z}_{cl}}{r_g+\boldsymbol{Z}_{cl}}\sigma_p\frac{\partial k_g}{\partial e_g}\end{aligned}\right\} \quad (83)$$

The magnitudes of $(Det.\ I)_l$ and $(Det.\ E)_l$ in terms of R_1 and C_1 are

$$\left.\begin{aligned}(Det.\ I)_l &= \frac{r_p}{4\sqrt{(r_p+R_{bl})^2+X_{bl}^2}}\cdot\\ &\quad \frac{R_1}{\sqrt{(R_1^2C_1^2\omega_l^2+1)+R_1k_g(2+R_1k_g)}}\sigma_p\frac{\partial k_g}{\partial e_g}\\ (Det.\ E)_l &= \frac{r_p}{4}\frac{R_1}{\sqrt{R_1^2C_1^2\omega_l^2+1+R_1k_g(2+R_1k_g)}}\sigma_p\frac{\partial k_g}{\partial e_g}\end{aligned}\right\} \quad (84)$$

Figure 17 shows the manner in which $\sigma_p\frac{\partial k_g}{\partial e_g}$ varies with $\bar{E}_g$ for a type 301-A tube.

The shape of the detection coefficient curve as plotted against $\bar{E}_g$ is as much dependent upon the fraction

$$F=\frac{R_1}{\sqrt{(R_1^2C_1^2\omega_l^2+1)+R_1k_g(2+R_1k_g)}} \quad (85)$$

as upon the shape of the curves of $\sigma_p\frac{\partial k_g}{\partial e_g}$. Furthermore, the fraction (85) contains the term $R_1^2C_1^2\omega_l^2$, which makes the detection coefficient a function of frequency.

The rapid manner in which F varies with grid polarizing potential for particular plate potentials and values of R_1 and C_1, is shown in Figure 18, and the variation of F with frequency is indicated in Figure 19.

The detection coefficient due only to grid-current rectification is equal to

$$(\textbf{Det. } I)_l = \frac{r_p}{4(r_p + Z_{bl})}\left[F \times \sigma_p \frac{\partial k_g}{\partial e_g}\right] \qquad (87)$$

The value of $(Det.\ I)_l$ for $\overline{E}_p = 60$ volts has been calculated from (87) and is later shown compared with an experimental determination of the detection coefficient for the 301-A tube.

Case 4. Grid-Circuit Rectification with Tickler Coil. The diagram of connections for this case is the same as shown in Figure 16 except that a tickler coil is added in the plate circuit. The back coupling, as in *Case* 2, in effect decreases the effective resistance of the oscillatory circuit resulting in an increased value of the impressed signal voltage. The tickler coil also introduces a radio reactance in the plate circuit.

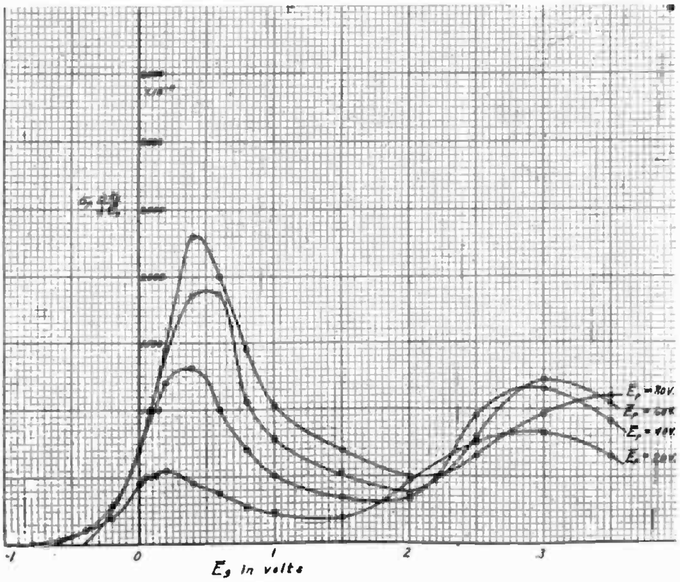

FIGURE 17—$\sigma_p \dfrac{\partial k_g}{\partial e}$ for a Cunningham 301-A Tube

The detection coefficient, therefore, depends upon every term of the complete equation (73) or of the better form (76). The first term is much the largest of the four terms. The third term is usually negligible. The second and fourth terms are of the same order of magnitude, but of opposite sign. Therefore, the radio load in the plate circuit is in this case an advantage because it increases the value of the fourth term which acts to help out the first or main term.

III—EXPERIMENTAL DETERMINATION OF DETECTION COEFFICIENT AND CHECK OF THEORY

Method of Measurement of Detection Coefficient. The current detection coefficient can easily be measured by a null balance method best explained by reference to the diagram of connections shown in Figure 20. The plate circuit of the detector under test contains the primary winding of an aircore mutual inductance M_b. The primary winding L_b is shunted by a radio by-pass condenser C_b. The grid circuit of the detector tube is connected through the stoppage condenser C_1 and leak R_1, when these are used, and through a potential divider, supplying the grid polarizing potential $\overline{E}_C$, to the potential terminals of a small known four-terminal resistance R_o. The resistance R_o is included in the oscillatory circuit of a radio oscillator so that the oscillatory current multiplied by the resistance R_o gives the radio

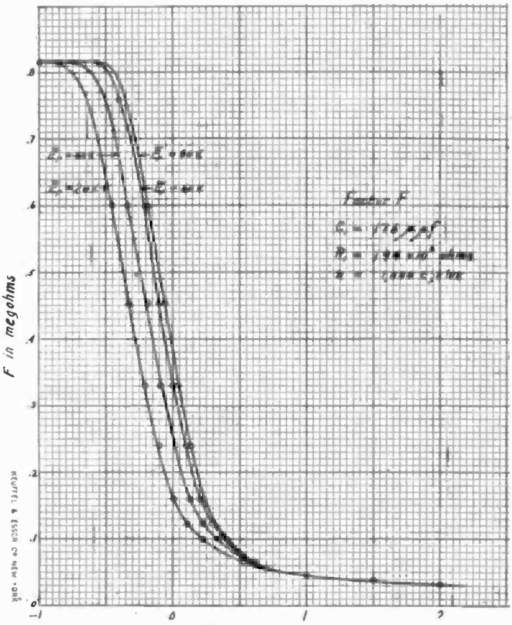

FIGURE 18—Factor F for a Cunningham 301-A Tube

signal impressed on the detector. The plate circuit of the oscillator is fed by a plate battery $\bar{E}_B{}'$ in series with a source of 1,000-cycle alternating current of known voltage read by voltmeter E_m. A radio by-pass condenser C' is connected across E_m.

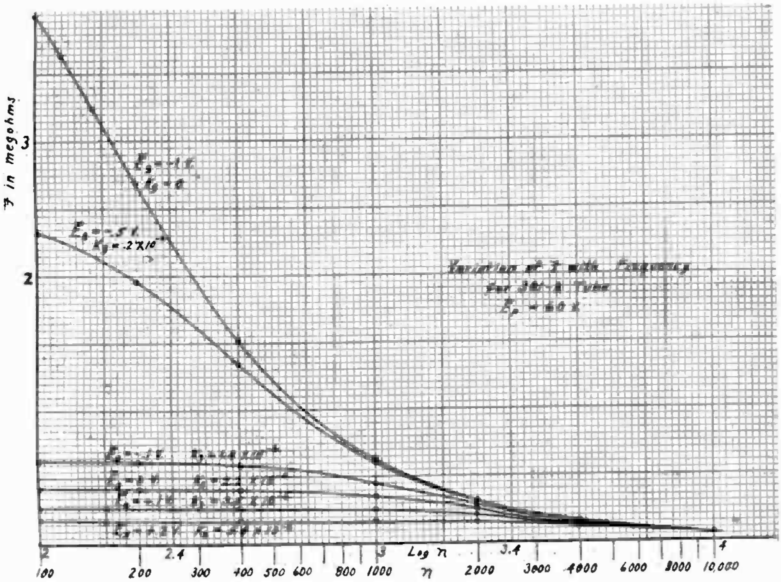

FIGURE 19—Variation of Factor F with Frequency—$E_p = 60$ Volts

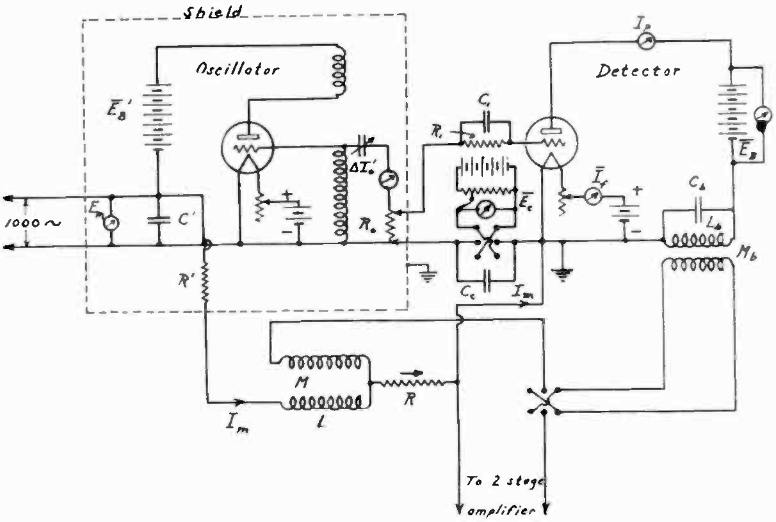

FIGURE 20—Diagram of Connections for Measurement of Detection Coefficient

This alternating current modulates the radio-frequency current through the resistance R_o. The modulated radio signal acting through the detector gives rise to a demodulated 1,000-cycle current in the primary winding L_b. The induced voltage in the secondary winding of M_b is measured by balancing it in phase and magnitude by the voltage $I_m\sqrt{R^2+M^2\omega_l^2}$ where both R and M are variable. The current I_m is taken from source E_m through a very high resistance R' so that I_m is practically equal to $\frac{E_m}{R'}$. A grounded shield encloses the oscillator as shown in Figure 20.

The values of various quantities as used in the measurements to be described are as follows:

$L_b = .6524\ h$
$M_b = .6538\ h$
$C_1 = 176\ \mu\mu f$
$R_o = .914$ ohms
$E_B' = 30$ volts
$R' = 60{,}240$ ohms

$R_b = 537$ ohms
$C_b = .002\ \mu f$
$R_1 = 1.937$ megohms
$\lambda = 250$ meters
$E_m = 12$ volts
$C' = .010\ \mu f$

M variable up to 10,000 microhenrys (Campbell Mutual Inductance)

R variable in hundredths, tenths, units, and tens.

$C_c = 1\ \mu f$

$\Delta I_o'$ about 30 milliamperes

Oscillator 201-A tube

In order to determine if the oscillator modulates properly, the oscillatory current was measured for various values of $\bar{E}_B'$, while E_m was zero. The back coupling of the oscillator was adjusted until the output current plotted against $\bar{E}_B'$ was a straight line over the range of $\bar{E}_B'$ from 10 to 50 volts, although the plot extended did not pass through the origin. The plate voltage was then set at 30 and it was assumed that when a 12-volt alternating voltage was superimposed upon the steady 30 volts the amplitude of the radio-frequency current varied linearly as the total plate voltage varied from $31-12\sqrt{2}$ or 13 volts to $30+12\sqrt{2}$ or 47 volts. The degree of modulation was obtained by reading from the linear plot the current corresponding to 30 volts and that corresponding to 13 volts. Then the difference between the two currents divided by the former gives the degree of modulation m. This factor was in these experiments about .6. The unmodulated radio current ΔI_o was taken from the plot for $\bar{E}_B'$ equal to 30.

The value of the audio current in the plate circuit of the detector is obviously

$$\Delta^2 I_p = \frac{E_m \sqrt{R^2 + M^2(2\pi \,.\, 1{,}000)^2}}{R' \; M_b(2\pi \,.\, 1{,}000)} \tag{88}$$

Then referring to equation (30), the final expression from which the detection coefficient can be computed is

$$\begin{aligned}(Det.\ I)_l &= \frac{\Delta^2 I_p}{\sqrt{2}\, m\, R_o^2(\sqrt{2}\,\Delta I_o)^2} \\ &= \frac{E_m \sqrt{R^2 + M^2(2\pi \,.\, 1{,}000)^2}}{4{,}000 \,.\, \pi \sqrt{2}\, m\, R_o^2\, R^1\, M_b(\Delta I_o)^2}\end{aligned} \tag{89}$$

Since m varies nearly proportionally with E_m, the ratio $\frac{E_m}{m}$ remains practically constant if E_m varies slightly.

The phase of the demodulated plate current can also be determined. If $L\,\omega_l$ is small in comparison with R', then I_m is in phase with the modulation. The mutual inductance M_b introduces a 90° lag so that the angle between the audio plate current and the modulation envelope is

$$\phi = tan^{-1} \frac{R}{M\,\omega_l}$$

It is sometimes instructive to plot separately the two components of $(\boldsymbol{Det.\ I})_l$ obtained from the following relations.

$$(Det.\ I)_{l\,\alpha} = \frac{E_m M}{2\sqrt{2}\, m\, R_o^2\, R'\; M_b(\Delta I_o)^2} \;\; \begin{matrix} \textit{In-phase} \\ \textit{component} \end{matrix} \tag{90}$$

$$(Det.\ I)_{l\,\beta} = \frac{E_m R}{4{,}000 \,.\, \pi \sqrt{2}\, m\, R_o^2\, R^1\, M_b(\Delta I_o)^2} \;\; \begin{matrix} \textit{quadrature} \\ \textit{component} \end{matrix} \tag{91}$$

Experimental Results. In Figure 21 are plotted the experimentally determined values of current detection coefficient for the Cunningham 301-A tube previously used in the measurement of tube characteristics. These curves are similar in shape to the curves of Figure 13. The solid points shown on the dotted curve of Figure 21 were calculated by equation (78) for $\bar{E}_B = 60$ volts. These calculated points fall on a curve similar in shape to the observed curve but somewhat lower. This discrepancy is not surprising considering the difficulty in deriving graphically accurate values of $\frac{\partial\, \sigma_p}{\partial\; e_g}$.

Figure 22 presents the experimental curves of the current detection coefficient for the 301-A tube used with a stoppage condenser of 176 $\mu\mu f$. and grid leak of 1.937 megohms. In taking these curves the actual grid potential was determined by short circuiting the grid leak and finding the grid potential which would reproduce the plate current observed when making a determination of detection coefficient. The solid points in Figure 22 were calculated from equation (84) for $\overline{E}_B$ equal to 60 volts.

Figures 23 and 24 show the effect of ionization upon the current detection coefficient. The curves of Figure 23 were taken with a Cunningham Gas Tube type C-300, operating somewhat below normal filament current in order to show temperature saturation of emission. No grid impedance was used so that the detection results from bends and kinks in the plate current curve. The two components of $(\boldsymbol{Det.\ I})_l$ are plotted, $(Det.\ I)_\alpha$ corresponding to the component of audio plate current in phase with the modulating voltage, and $(Det.\ I)_\beta$ corresponding to the component ninety degrees in advance of the modulating voltage. The numerical value of detection coefficient is of course

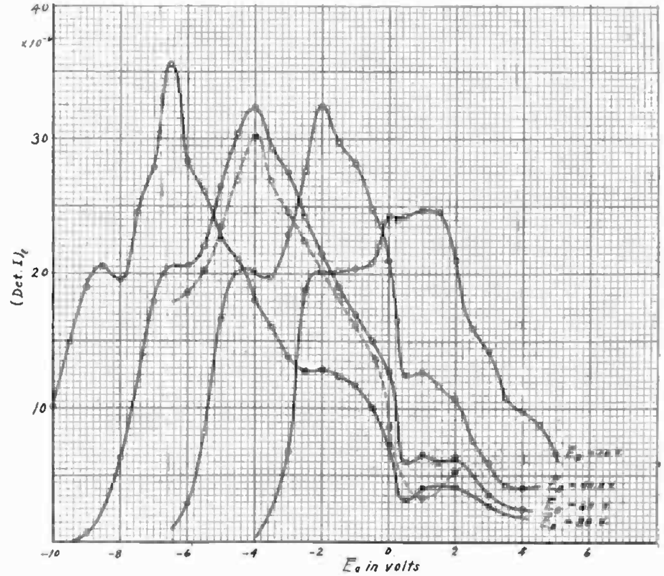

FIGURE 21—Detection Coefficient Curves without Grid Impedance for Cunningham 301-A-Tube

Full lines observed Dotted lines calculated

$$(Det.\ I)_l = \sqrt{(Det.\ I)_\alpha^2 + (Det.\ I)_\beta^2} \qquad (92)$$

Referring to Figure 23, it will be noticed that the plate current curves, particularly the one for 31.5 volts, have peculiar shapes. The first rapid rise may be attributed to the neutralization of space change by the positive ions produced by ionization of the gas. The plate current does not continue to rise at the first rapid rate because, due to the small amount of gas present, the supply of positive ions is insufficient to neutralize the rapidly increasing negative space charge and the current flattens out. The second bend, which is concave downward, is due to temperature saturation of emission. The detection coefficient reflects

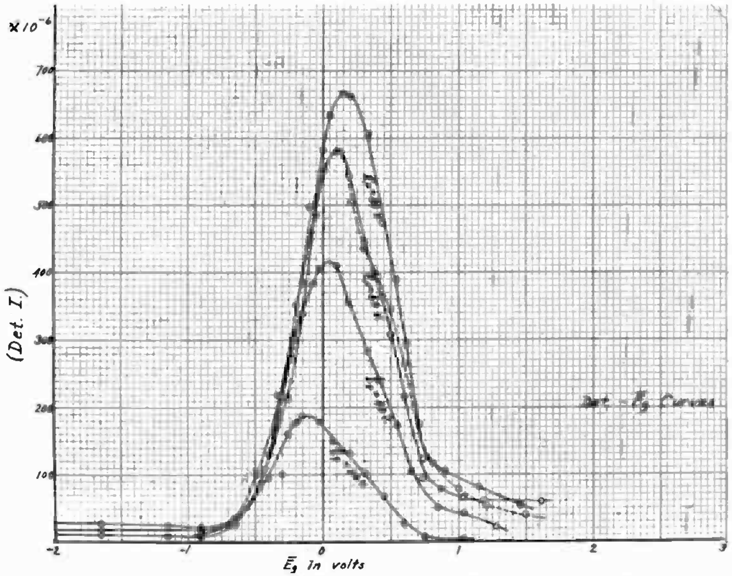

FIGURE 22—Detection Coefficient Curves with Stopage Condenser for Cunningham 301-A Tube

Full lines observed

Dotted lines calculated

the several bends in the plate current curves, the $(Det.\ I)a$ being positive for bends which are concave upward and negative for bends which are concave downward. The very large values of detection coefficient corresponding to the bends which are concave downward can hardly be explained by the curvative of the plate current curves shown. The large values may be due to a time lag in ionization causing the dynamic plate current curve to be different from the static characteristic. The only evidence of kinks in the plate current curves is the small hump on the $(Det.\ I)_a$ curve for 31.5 volts. The curves were taken expressly to show the effect of ionization in changing the rates of curvature of the plate current curves.

The curves of Figure 24 were taken to show the effects of definite ionization kinks in the plate-current curve. The tube used contains caesium vapor at low pressure and is known as a UX-200-A. It was operated without grid impedance. One set of curves corresponds to a plate potential of twenty volts, but the other set is for a plate potential of -1.5 volts. The latter curves were taken not because of any practical value in using a negative plate potential, but because of the interest in the fact that electrons do sometimes flow to a negatively charged plate due to contact potentials, and because of the definite ionization kink obtained at these low potentials as evidenced by the shape of the detection coefficient curves. This tube is supposed to be used with a stoppage condenser and grid leak. Under this condition of use, the maximum detection coefficient was found to have the very large value of $7{,}390 \times 10^{-6} \dfrac{amps}{(volts)^2}$

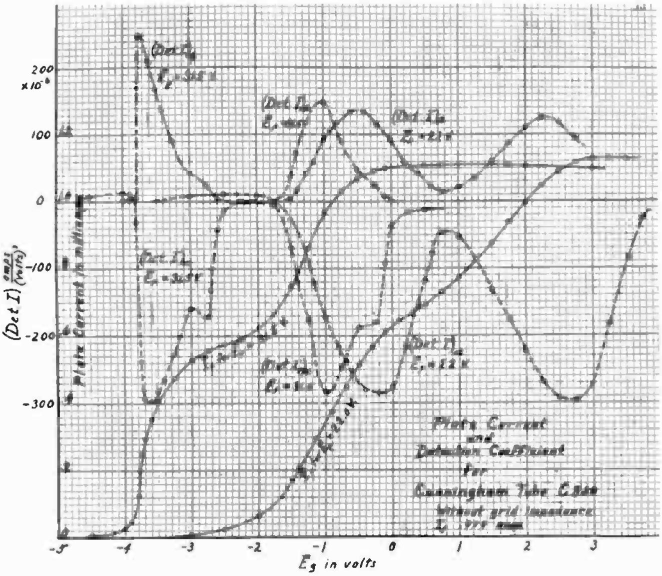

FIGURE 23

Although gas tubes have very large values of detection coefficient there is always present more or less hiss, which, to some extent, diminishes their apparent superiority as detectors. Furthermore, their input impedance is usually very low due to their large grid conductance.

IV—CONCLUSIONS

A rectifying detector depends in its action upon a non-linear relation between the instantaneous output current and the instantaneous applied voltage.

If the impressed modulated signal voltage is small, the output contains the following components: a constant current; a current of modulation frequency; a current of double modulation frequency; and currents of frequencies equal to the sum and difference of the several modulation frequencies. All of these component currents are proportional to the square of the impressed signal voltage. The component of modulation frequency is proportional to the degree of modulation; all others depend upon the square of the modulation or upon the product of the two modulation factors, and hence are small in comparison with the current of modulation frequency, if the degree of modulation is small.

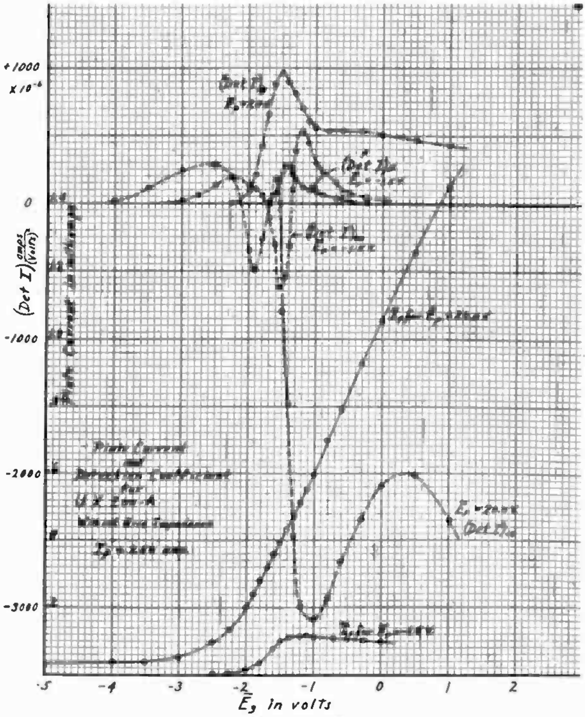

FIGURE 24

The voltage detection coefficient is a better measure of the detecting action of a rectifying detector because its value is independent of the impedance interposed in the plate circuit during the measurement of the detection coefficient. The voltage detection coefficient gives an equivalent voltage which, considered as acting in the plate circuit, gives the audio current which flows through the circuit containing r_p and the plate impedance normally used with the detector. A knowledge of r_p is thus necessary.

A hard triode used without a grid-circuit impedance, that is, stoppage condenser and leak or the equivalent, and with no radio impedance in the plate circuit, depends for its detecting action entirely upon the bends of the plate current curve. The resulting detection is usually very small. A gas tube thus used gives much greater detection than a similar tube highly exhausted because ionization increases both the lower and upper bends of the plate current curve. Ionization also causes kinks in the plate current curve resulting in a high sensitivity at such points due to the very large values of $\frac{\partial \sigma_p}{\partial e_g}$. A radio-frequency impedance in the plate circuit of a hard tube used without a grid impedance decreases the detection coefficient due to the lower bend of the plate-current curve. A tickler usually more than makes up for this decrease in detection coefficient by increasing the strength of the impressed signal. All audio output devices in the plate circuit should be shunted by a condenser whose reactance is small for the radio frequency.

The sensitivity of a hard triode used with a grid impedance depends upon the product $\sigma_p \frac{\partial k_g}{\partial e_g}$ and upon a factor F, which is equal to the equivalent parallel impedance of the grid impedance and the grid-to-filament resistance. The greatest sensitivity obtained for this case when using a hard tube is usually much greater than the maximum sensitivity obtainable without a grid impedance. σ_p should be made large by using the proper plate voltage. The value of $\sigma_p \frac{\partial k_g}{\partial e_g}$ is a maximum for polarizing voltages of a few tenths positive, but F falls so rapidly for positive grid voltages, due to the increase of k_g, that the point of maximum sensitivity is found at a grid voltage more negative than that which gives a maximum value of $\sigma_p \frac{\partial k_g}{\partial e_g}$. The detection coefficient curves are very narrow peaks so it is necessary to adjust the grid polarizing potential to the proper value, usually one or two-tenths of a volt positive. Because of the steady rectified component of current in the grid circuit, a strong signal unfortunately alters the polarizing potential.

The ordinary grid leak and stoppage condenser is not the best form of grid impedance because of its variation with frequency and its large value at zero frequency. The ideal impedance is one having negligible resistance to steady currents, a high impedance for frequencies from 100 to 10,000, and low impedance for the radio frequency current.

A tickler coil used with a detector provided with a grid impedance increases the detection coefficient.

The last two terms in the detection-coefficient equation (76) are usually of small importance. At low plate voltages and with gas tubes these terms may, however, contribute largely to the detection-coefficient.

DISCUSSION ON

LONG DISTANCE RADIO RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1925

(L. W. AUSTIN)

K. Sreenivasan: The curves given by Dr. Austin as a result of his work over a number of years established quite definitely that there are well marked seasonal variations of field strengths of radio signals. The similarity of the curves of monthly averages over a period of four years is certainly striking. But I am a little puzzled to observe the unmistakable gradual increase of field strength of Bordeaux (LY) from year to year as shown in Fig. 5. Can this be a part of a cyclic change extending over a number of years corresponding to any terrestrial or solar phenomenon? Has the author observed similar occurrences with any other stations, say Nauen or Lyons?

The strongest intensity changes of Bordeaux seem to occur anywhere between November and January at Washington and at about the same time at Meudon. That this curve is not an isolated instance is borne out by the observations of R. Mesny and J. Hollingworth. But while the Washington observations on Bordeaux always show an increase in field strength, the curves obtained by Hollingworth for a member of stations show in some cases as marked a reduction in field intensity.

The one point of difference in the two cases is that while the Washington observations represent monthly averages, those of Hollingworth show weekly averages. In some of my observations on Madras (Fort) Radio (295 km. from Bangalore) the following results were obtained for the weekly averages, the working wave length being 4,000 km.

Week ending with	field strength in μv/m
10th October	765
17th ”	833
24th ”	992
31 st ”	611
7th November	664
14th ”	779
21st ”	982
28th ”	788

I have given these figures to show the marked changes that occur during this period even on 4 km. specially the low values of 611 and 664 μv/m for the weeks ending with 31st October and 7th November. While sudden and rapid changes in intensity take place during this period, I do not think, in view of the foregoing, that it will always be an increase in every case.

In this connection, the suggested explanation of Dr. Austin that the agreement of seasonal variations between the observations of Washington and Meudon or Bordeaux is due to causes in the neighborhood of the transmitting station is worthy of note. But in view of Hollingworth's observations and to a little extent of the results on Madras, I am inclined to agree with Hollingworth that this change might be attributed to a rise in the height of the Kennelly-Heaviside layer.

It would have been helpful if Dr. Austin had given us the weekly averages and if and where necessary the daily values of field intensity. The monthly averages have, I think, a tendency to hide the comparatively transient changes which may afford interesting information about the propagation phenomena. But for the derivation of a working formula, the arithmetical work involved in using weekly values would no doubt be very great and in a sense unprofitable.

Quite recently there was reported a severe magnetic storm which had the effect of completely blocking communication from England to Canada by the beam system of short waves. On the other hand, the reception of Annapolis in England showed, it appears, no signs of any departure from the normal. I should like to know if the author observed any abnormality in his readings of the 2nd and 3rd week of October, when the magnetic storm seems to have been strongest. If no abnormality was observed, the explanation for the difference in the case of long and "beam" short waves does not seem to be easy.

L. W. Austin: In commenting on Mr. Sreenivasan's discussion of my paper I should like to point out that there are too many varying factors entering into the changes in signal intensity to permit us to expect to find anything like uniformity from year to year except by averaging over rather long periods.

The causes of the day by day or week by week variations may be quite distinct from the prevailing causes of the longer period changes.

The observations made in Washington since 1922 on the European stations at about 3 P. M., E. S. T., when during much of the year part of the signal path lies in darkness, indicate in most cases a slight but regular increase in yearly average intensity, which is most marked in the case of Bordeaux.

These changes, I agree, are probably due to changes in the Kennelly-Heaviside layer either in average height or in ionization, and it seems possible that they may be connected with changes in solar activity. The reality of this relationship can probably only be settled with certainty by regular observations on signal intensity covering at least two sunspot cycles.

In regard to Mr. Sreenivasan's question concerning long wave reception during the second and third weeks of last October, our records do not show any exceptional peculiarities.

Of course we are always glad to send mimeograph copies of our daily signal measurements to anyone interested, as was announced in the "Proceedings," when the regular publication of these measurements was discontinued some years ago.

DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued November 2, 1926—November 23, 1926

By John B. Brady

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,604,552—HAMMARLUND, L. A., Floral Park, New York. Filed Jan. 10, 1925, issued Oct. 26, 1926. Assigned to The Hammarlund Manufacturing Company, Incorporated.

CONDENSER of variable plate construction where the plates are supported in slotted bearing members which are in turn carried by grooved insulation members constituting the condenser frame.

1,604,986—W. E. GARITY, New York, N. Y. Filed October 7, 1921, issued Nov. 2, 1926. Assigned to De Forest Radio Telegraph & Telephone Co.

ALTERNATING CURRENT GENERATOR in which a construction of tubes is employed having two separated cold electrodes with a restricted electron discharge path therebetween for use in variable light controlling systems.

1,604,987—J. S. GARVIN, New York, N. Y. Filed Dec. 13, 1920, issued Nov. 2, 1926. Assigned to Western Electric Co. Inc.

ENERGIZING CIRCUITS FOR VACUUM TUBES in which direct current generators are employed for supplying the filament and plate circuits of electron tubes and resistances arranged in circuit for controlling the potentials.

1,605,001—F. SCHROTER, Berlin-Schmargendorf, Germany. Filed Mar. 30, 1921, issued Nov. 2, 1926. Assigned to American Telephone & Telegraph Company.

VACUUM VALVE WITH GLOW DISCHARGE where a pair of electrodes are spaced apart by a relatively small unobstructed gap by which a glow discharge may be produced.

1,605,121—F. LOWENSTEIN, New York, N. Y.; John C. Wait, Administrator of said F. Lowenstein, deceased. Filed May 23, 1921, issued Nov. 2, 1926. Assigned to Radio Patents Corp.

RADIO APPARATUS consisting of a wave generator system with simultaneously operative control members for adjusting the several circuits of a signaling system through single actuator.

1,605,230—W. F. HENDRY, New York, N. Y. Filed July 14, 1920, issued Nov. 2, 1926. Assigned to Western Electric Co., Inc.

ELECTRON DISCHARGE DEVICE AND METHOD OF MANUFACTURING THE SAME where a collar is provided for gripping a stem within a tube with blocks of insulating material attached to the collar for supporting the electrodes within the tube.

1,605,333—F. W. DUNMORE, Washington, D. C. Filed Sept. 4, 1925, issued Nov. 2, 1926.

RADIO RECEIVING APPARATUS in which several circuits are tuned simultaneously through a series of independent adjusters operated by a single actuator arranged to move a series of condensers in cooperation with a system of cams.

1,605,607—P. O. PEDERSEN, Frederiksberg, Denmark. Filed June 21, 1919, issued Nov. 2, 1926. Assigned to Poulsen Wireless Corp.

ARC CONVERTOR wherein a pair of arc electrodes are arranged at a distance equal to the most favorable distance of travel of the arc along the cathode.

1,605,608—P. O. PEDERSEN, Frederiksberg, Denmark. Filed June 21, 1919, issued Nov. 2, 1926. Assigned to Poulsen Wireless Corp.

ARC CONVERTOR in which a pair of electrodes are provided for establishing an arc with a plurality of blow pipes for imparting cooling blast to the cathode at a point where the arc is to be extinguished.

Received by the Editor January 26, 1927.

1,605,627—R. E. THOMPSON, Nyack, N. Y. Filed April 17, 1922, issued Nov. 2, 1926. Assigned One-half to Wireless Improvement Co., and one-half to Wireless Improvement Co., Inc.

RADIO RECEIVING APPARATUS in which a signaling receiving circuit is adjusted over a definite frequency range with an electron tube system having a non-resonant grid circuit with a high resistance winding therein for preventing oscillation while permitting regeneration.

1,605,723—G. W. HEATH, East Orange, N. J. Filed Oct. 11, 1922, issued Nov. 2, 1926. Assigned one-half to Alfred C. Heath.

VARIABLE CONDENSER in which vernier movement for obtaining accurate variation engages with the periphery of one of the rotor plates.

1,605,735—C. W. HOUGH, Boonville, N. Y. Filed Jan. 16, 1926, issued Nov. 2, 1926. Assigned to Wired Radio, Inc.

ELECTRON TUBE SYSTEM wherein the usual house lighting current may be employed for exciting the circuits of the electron tubes, the same tube serving as a rectifier and an amplifier in a signal receiving circuit.

1,605,748—J. L. MARTIN, Amarillo, Texas. Filed Nov. 20, 1925, issued Nov. 2, 1926.

AERIAL AND GROUND CONNECTION FOR RADIO SETS where connections may be made between a radio cabinet through contacts established between the cabinet and connections upon a supporting structure on which the cabinet may rest.

1,605,804—D. F. ASBURY, Washington, D. C. Filed Jan. 18, 1924, issued Nov. 2, 1926.

OPERATING MECHANISM FOR PLURAL TUNING UNIT RADIO APPARATUS where a plurality of condensers may be simultaneously actuated or the condensers may be independently moved through a series of independent adjusters.

1,606,008—H. J. WIEGAND, Milwaukee, Wis. Filed May 29, 1922, issued Nov. 9, 1926. Assigned to The Cutler-Hammer Mfg. Co.

CAPACITANCE CONTROL DEVICE where sets of rollers are arranged for the movement of a flexible conductive sheet from one roller system to another in the variation of electrical capacity.

1,606,184—V. L. Ronci, Brooklyn, N. Y. Filed Dec. 17, 1924, issued Nov. 9, 1926. Assigned to Western Electric Co.

ELECTRON DISCHARGE DEVICE where a collar is provided around a tubular member within an electron tube structure with a bifurcated member straddling the collar and providing an electron support within the tube.

1,606,283—R. WOLF and A. A. WADSWORTH, Jr., of New York, N. Y. Filed Sept. 22, 1922, issued Nov. 9, 1926. Assigned to Hollan S. Duell and Kenneth O'Brien.

RADIO RECEIVING SYSTEM in which an electron tube circuit has its input and output circuits coupled by a free ended inductance connected with the cathode circuit.

1,606,476—J. O. MAUBORGNE and GUY HILL, of Washington, D. C. Filed May 24, 1920, issued Nov. 9, 1926.

RADIO SIGNALING SYSTEM in which a wave coil is arranged for operation as a direction finder.

1,606,660—H. E. MOTT and J. P. MAURER, of Montreal, Canada. Filed Nov. 20, 1925, issued Nov. 9, 1926.

ELECTRICAL CONDENSER of variable plate construction in which the rotor plates are driven through a system of gears for obtaining micrometer adjustment of the condenser.

1,606,773—E. L. NELSON, of East Orange, New Jersey. Filed Aug. 3, 1923, issued Nov. 16, 1926. Assigned to Western Electric Company, Inc.

TRANSMISSION SYSTEM for duplex operation wherein carrier waves of different frequencies are employed and while the frequencies may vary during duplex operation, the difference between the frequencies is maintained constant.

1,606,775—ALEXANDER NYMAN, of Wilkinsburg, Pennsylvania. Filed July 15, 1920, issued Nov. 16, 1926. Assigned to Westinghouse Electric & Manufacturing Company.

COMBINED RADIO SENDING AND RECEIVING SYSTEM where electron tube circuits are arranged in balanced relationship in such manner that mechanical switching between the transmission and receiving periods is unnecessary.

1,606,791—J. W. HORTON, of Bloomfield, New Jersey. Filed July 18, 1924, issued Nov. 16, 1926. Assigned to Western Electric Company, Inc.

OSCILLATION GENERATOR comprising electron tube circuits arranged for the production of oscillations with means for preventing reaction of the load circuit upon the electron tube circuits comprising a path including resistance, capacity, inductance elements in series and resistance elements in shunt therewith.

1,606,792—F. W. ISLES, of Brooklyn, N. Y. Filed Oct. 6, 1919, issued Nov. 16, 1926. Assigned to Western Electric Company, Inc.

OSCILLATION GENERATOR FOR CURRENT OF CONTINUOUSLY VARYING FREQUENCY consisting of a variable conductor having a rotatable core portion for varying its impedance. The inductance of the inductor is cyclically varied by uniformly and continuously rotating the core portion.

1,606,809—B. Rosenbaum, of Berlin, Germany. Filed Oct. 31, 1921, issued Nov. 16, 1926. Assigned to Westinghouse Electric & Manufacturing Co.

RADIO APPARATUS in which the tuning elements are interlinked by a system of levers for simultaneous operation for controlling the apparatus from a single actuator.

1,606,929—V. A. R. FERNANDES, of Cambridge, Mass. Filed June 6, 1925, issued Nov. 16, 1926.

ARRANGEMENT FOR PRODUCING OSCILLATIONS BY MEANS OF VACUUM TUBES where a vacuum tube path is provided for the current in shunt with the main vacuum tube path for varying the distribution of the current between the two paths at the period of the oscillations.

1,606,940—ROBERT HERZOG, of Berlin, and LEO PUNGS, of Berlin-Charlottenburg, Germany. Filed Aug. 23, 1921, issued Nov. 16, 1926. Assigned to C. Lorenz Aktiengesellschaft.

MEANS FOR KEYING IN RADIO TELEGRAPHY wherein an electron tube circuit is arranged to be controlled for in turn controlling the magnetic condition of an inductance device in the antenna circuit of a transmitter for the formation of signals.

1,606,958—E. PICARD, of Paris, France. Filed June 19, 1925, issued Nov. 16, 1926.

VARIABLE CONDENSER WITH SMALL LOSS wherein silicious insulating material is provided for supporting the frame members of a condenser which in turn carries the plates of the condenser.

1,607,119—O. C. DEUTSCHER, of Brooklyn, N. Y. Filed Aug. 10, 1925, issued Nov. 16, 1926.

CONDENSER of variable construction where a hinged plate of triangular shape may be swung toward or away from a fixed plate.

1,607,158—J. H. HAMMOND, JR., Gloucester, Mass. Original filed Feb. 15, 1917; renewed July 24, 1923; issued Nov. 16, 1926.

MULTIPLEX METHOD AND SYSTEM FOR THE TRANSMISSION OF RADIANT ENERGY in which a series of oscillations are transmitted and impressed upon selectively tuned circuits which operate independently for separating the signaling energy into different channels.

1,607,158—J. H. HAMMOND, Jr., Gloucester, Mass. Original filed Feb. 15, 1918; renewed Oct. 31, 1922; divided Aug. 2, 1923; renewed Mar. 27, 1926; issued Nov. 16, 1926.

ELECTRIC DEVICE in the form of a vapor tube particularly adapted for the generation of high frequency oscillations.

1,607,278—P. C. HEWITT, Ringwood Manor, N. J. Original filed April 23, 1920; renewed July 21, 1926; issued Nov. 16, 1926. Assigned to Cooper Hewitt Electric Co.

SYSTEM FOR THE TRANSMISSION AND RECEPTION OF RADIANT ENERGY for securing secrecy in transmission and reception where a series of wave trains are simultaneously emitted to obtain a predetermined difference in phase relationship between the different waves for selective reception at a correspondingly arranged receiver.

1,607,277—P. C. HEWITT, Ringwood Manor, N. J. Filed Rec. 1, 1920; issued Nov. 16, 1926. Assigned to Cooper Hewitt Electric Co.

METHOD AND SYSTEM FOR THE TRANSMISSION OF RADIANT ENERGY for secret signaling where a plurality of successive series of periodic impulses are transmitted and received for controlling by their conjoint action, the receiving apparatus.

1,607,277—P. C. HEWITT, Ringwood Manor, N. J. Filed Dec. 1, 1920; issued 1921; renewed July 31, 1926; issued Nov. 16, 1926. Assigned to Cooper Hewitt Electric Co.

SYSTEM FOR ELECTRICAL DISTRIBUTION in which a vapor tube is employed in a circuit for generating high frequency oscillations where the time period of operation of the tube is fixed by a special reactance circuit.

1,607,456—J. H. HAMMOND, JR., Gloucester, Mass. Original filed May 8, 1918; renewed May 1, 1925; issued Nov. 16, 1926.

METHOD OF AND APPARATUS FOR PRODUCING ALTERNATING CURRENTS in which a vapor tube is arranged for the transformation of direct or low frequency current into high frequency oscillations.

1,607,467—M. LATOUR, Paris, France. Filed Dec. 5, 1923, issued Nov. 16, 1926. Assigned to Latour Corporation.

HIGH POWER THERMIONIC VALVE in which the cathode is heated by currents induced therein by the control winding for obtaining electron emission which may be varied for effecting desired control in an associated circuit.

1,607,485—H. SCHMIDT, Berlin, Germany. Filed Feb. 2, 1924, issued Nov. 16, 1926.

WIRELESS SECRET TELEPHONY wherein the frequency of the transmitter is variably controlled and the receiving circuits at distant receiving stations correspondingly controlled in synchronism with the transmitter.

1,607,682—DE LOSS K. MARTIN, West Orange, N. J. Filed Mar. 22, 1924, issued Nov. 23, 1926. Assigned to American Telegraph and Telephone Company.

RADIO REPEATER SYSTEM where the frequency of the received waves is changed at each repeater station and undesirable singing between the repeater stations positively prevented.

1,607,683—DE LOSS K. MARTIN, West Orange, N. J. Filed Mar. 22, 1924, issued Nov. 23, 1926. Assigned to American Telephone and Telegraph Company.

RADIO REPEATER SYSTEM where two way communication may be conducted on different signaling frequencies without interference between the channels.

1,607,837—DE LOSS K. MARTIN, West Orange, N. J. Filed Mar. 22, 1924, issued Nov. 23, 1926. Assigned to American Telegraph and Telephone Company.

RADIO REPEATER SYSTEM in which a plurality of repeater stations are located between terminal stations and communications transmitted therebetween without interference between the channels.

1,607,856—ERNEST E. YAXLEY, Chicago, Ill. Filed Mar. 10, 1926, issued Nov. 23, 1926.

ELECTROSTATIC CONDENSER where a plurality of circular plates are built as a unit in a condenser structure and riverted at the center thereof.

1,607,952—G. HOLST and A. BOUWERS, Eindhoven, Netherlands. Filed Aug. 25, 1925, issued Nov. 23, 1926. Assigned to N. V. Philips' Gloeilampenfabrieken.

ELECTRIC DISCHARGE TUBE where the outer wall of the tube is built up of alternately positioned sections of conductive insulated material.

1,608,003—F. SCHAFFER, Berlin, Germany. Filed June 5, 1924, issued Nov. 23, 1926. Assigned to Gesellschaft fur Drahtlose Telegraphie.

ARRANGEMENT FOR GENERATING AUDIBLE FREQUENCIES IN HIGH FREQUENCY SIGNALING in which an electro-acoustic device is arranged in circuit with a signaling system for controlling the high frequency signaling energy.

Geographical Location of Members Elected January 5, 1927

Transferred to Fellow Grade

Winfield, L. I., N. Y.,	Pacent, L. G.

Transferred to Member Grade

Bloomfield, N. J., 216 Berkeley Ave.,	Mitchell, C. D.
New York City, 41 Park Row,	Aull, Wilson, Jr.

Elected to Member Grade

Cleveland, Ohio Case School Applied Science,	Martin, J. R.
Melbourne, Australia, 44 Lincoln Road, Essenden,	Bain, J. L.
London, England, General Post Office,	Lee, A. G.
Tonga, Mukualofa,	Land, John R.
U. S. S. R. (Russia), Moscow, Gagarensky Perenlok 2B,	Bashenoff, V. T.

Elected to Associate Grade

El Granada, California,	Kerr, M. F.
Glendale, California, 118 S. Brand Blvd.,	Spencer, B. M.
Los Angeles, California, 757 So. Main Street,	Reed, G. H.
Tujunga, California, P. O. Box 456,	Feringer, W.
Van Muys, California, 14707 Victory Blvd.,	Marsden, G. K.
Hartford, Conn., 46 So. Marshall St.,	Hatry, L. W.
Hartford, Conn., 314 Collins St.,	Page, A. C.
New Haven, Conn., 224 St. Ronan St.,	Weed, C. B.
Stamford, Conn., c-o Electric Specialty Co.,	Berry, E. W.
Stamford, Conn., c-o Electric Specialty Co.,	Haines, W. H.
Wethersfield, Conn., P. O. Box 108, Nott St.,	Mitchell, J. A.
Washington, D. C., Bellevue,	Carboni, A.
Washington, D. C., The Chevy Chase Apts.,	Gross, G. C.
Washington, D. C., Radio Sec., Bureau Standards,	Kirby, S. S.
Washington, D. C., 2633-15th Street, N. W.,	Lutz, J. A.
Washington, D. C., 633 Raleigh Place, S. E.	Merryman, E. A.
Washington, D. C., 2633-15th Street, N. W.,	Taylor, M. B.
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Chicago, Illinois, 4737 Ingleside Ave.,	Byrnes, W. C.
Chicago, Illinois, 821 Railway Exchange Bldg.,	Crossett, E. C.
Chicago, Illinois, 1076 Roosevelt Road,	Esmaker, J. B.
Chicago, Illinois, 5787 Ridge Ave.,	Gerstle, C.
Chicago, Illinois, 8729 Harvard Ave.,	Kramer, A. W.
Chicago, Illinois, 115 N. Elizabeth St.,	Kruger, L. H.
Chicago, Illinois, 5833 Circle Ave.,	Leverett, G. H.
Chicago, Illinois, 6729 Bosworth Ave.,	Levy, H.
Chicago, Illinois, 5638 S. Washtenaw Ave.,	McCarthy, F. M.
Chicago, Illinois, 4158 Argyle St.,	Moaz, E.
Chicago, Illinois, 3532 W. Adams St.,	O'Byrne, L. C.
Chicago, Illinois, 1647 East 69th St.,	Rose, J. K.
Chicago, Illinois, 2136 Indiana Ave.,	Roscoe, R.
Chicago, Illinois, 230 East Ohio St.,	Slauson, W. T.
Chicago, Illinois, 4734½ Woodlawn Ave.,	Sowell, G. O.
Chicago, Illinois, 7609 East Lake Terrace,	Tyrman, E.
Elmhurst, Illinois, 250 South St.,	Hercher, F. J.
Evanston, Illinois, 804 Ridge Terrace,	Conklin, E. H.
Freeport, Illinois, 130 East Main St.,	O'Connell, G.
Mt. Morris, Illinois, 1½ N. Wesley Ave.,	Tilden, E. R.
Oak Park, Illinois, 815 N. Lombard Ave.,	Downs, K. D.
Urbana, Illinois, 101 S. Busey Ave.,	Chen, C. M.
Clarinda, Iowa, Radio Station KSO,	Maxwell, P. D.
Dubuque, Iowa, 1228 N. Booth St.,	Smith, C. M.
Sabula, Iowa,	Kunua, R. L.
West Oelwein, Iowa, 211-5th Ave.,	Hartley, A. V.
Lawrence, Kansas, Univ. of Kansas,	Kent, P. N.
Wichita, Kansas, 1437 Ellis Ave.,	Andrews, A. E.
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Louisville, Ky., 316 E. Caldwell Ave.,	Marshall, A. B.
New Orleans, La., 217 Custom House Bldg.,	Deiler, T. G.
Baltimore, Md., Room 13, Custom House Bldg.,	Herndon, L. C.
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New Bedford, Mass., 162 North St.,	Pierce, C. G.
Roslindale, Mass., 570 Hyde Park Ave.,	Hobart, R. F.
Springfield, Mass., Amer. Bosch Magneto Co.,	Cotter, W. F.
Springfield, Mass., 887 Chestnut St.,	Fox, F. J.
Springfield, Mass., 2787 Main St.,	Laverty, F. W.
Springfield, Mass., 10 Cloran St.,	Vail, H. A.
Wellesley, Mass., Grove Street,	Russell, R. D.
Wollaston, Mass., 55 Franklin Ave.,	Dickman, C. T.
Royal Oak, Mich., 113 W. 4th St.,	Mogredge, J. N.
Detroit, Mich., 3885 Burns Ave.,	Baxter, C. E.
Detroit, Mich.,4464 Cass Ave.,	Robinson, P. S.
Detroit, Mich., 218 Manistique Ave.,	Seymour, F. M.
Detroit, Mich., ss Peter Reiss, Marine P. O.,	Strout, E. M.
Duluth, Minn., 529-4th Ave., East,	England, L. J.
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Lyndhurst, N. J., 246 Oriental Place,	Gamblin, S.
Orange, N. J., 125 Main St.,	Spencer, R. C.
Westmont, N. J., 112 Ardmore Ave.,	Densham, H. W.
W. Orange, N. J., 2 Harrison Ave.,	Smith, F. L.
Brooklyn, N. Y., 55 Hanson Place,	Bristow, F. R.
Brooklyn, N. Y., 192 Lincoln Place,	Brown, G. G.
Brooklyn, N. Y., 7107-14th Ave.,	Burdick, A. B.
Brooklyn, N. Y., 315 East 96th St.,	Lifschitz, J. L.
Brooklyn, N. Y., 87 Macon St.,	Scofield, R. W.
Buffalo, N. Y., 829 Michigan Ave.,	Brown, S. W.
Buffalo, N. Y., 77 Depew Ave.,	Moore, C. G.
Buffalo, N. Y., 169 Mariner St.,	Morley, W. L.
Jamaica, N. Y., 286 Amherst Ave.,	Steinberger, A. W.
New York, N. Y., 627 W. 43rd St.,	Cockaday, L. M.
New York, N. Y., 1 East 213th St.,	Geise, H.
New York, N. Y., 60 Beaver St.,	Harley, J. B.
New York, N. Y., 133 East 84th St.,	Loscalzo, P. A.
New York, N. Y., 463 West St.,	Mullett, C. B.
New York, N. Y., 120 East 86th St.,	Powley, A. T.
New York, N. Y., 120 Broadway,	Ridgway, J. L.
New York, N. Y., uss Cincinnati,	Riley, C. D.
New York, N. Y., uss Dale,	Wilcox, H. L.
Rochester, N. Y., 1060 University Ave.,	Hanover, E. A.
Rocky Point, N. Y., Radio Corp'n. America,	Barbor, V. H.
Schenectady, N. Y., 177 Nott Terrace,	Roys, H. E.
Schenectady, N. Y., 525 Liberty St.,	Stanton, C. F.
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Oklahoma City, Okla., 423 East 14th St.,	Cole, R. H.
Atglen, Penn.,	Cowan, H. B.
Drexel Hill, Penn., 3232 Marshall Rd.,	Stobbart, A. J.
Johnston, Penn., c-o Star Radio Co.,	Hanson, H. M.
Johnston, Penn., 17 Cox St.,	Reid, A. J.
Philadelphia, Penn., 1533 Pine St.,	Benner, H. S.
Philadelphia, Penn., 1915 S. 12 St.,	Eaton, G. W.
Philadelphia, Penn., 1218 Overington St.,	Eby, J. B.
Philadelphia Penn., Bourse Bldg.,	Hutchinson, S. J.
Philadelphia, Penn., 5442 Sanson St.,	Mahon, F. A.
Philadelphia, Penn., 4422 Chestnut St.,	Wiley, P. F.
York, Penn., 135 W. Jackson St.,	Stewart, J. W.
Aiken, So. Carolina,	Terry, L. M.
Dallas, Texas, 2122 Jackson St.,	Braden, L. O.
Dallas, Texas, 834 No. Marsalis,	Richards, F. O.
Clarendon, Va., 424 Latterner Ave.,	Looney, Don B.
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Casper, Wyoming, 1022 So. Ash St.,	Hood, N. R.
Laramie, Wyoming, Station KFBU,	Walker, G. D.

Foreign Associates

Brisbane, Queensland, Australia,	Stevens, F. W.
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Toronto, Ontario, Canada, National Building, Bay St.,	Linnell, A. P.
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Toronto, Ontario, Canada, 120 Avenue Rd.,	Williams, J. C.
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Victoria, B. C., 1809 Second St.,	Gilbert, G.
Ladowitz, Czechoslovakia,	Kafka, T. H.
Kent, England, Whitstable, "Hilder Crest,"	Pollard, A. E.
London, SW4, England, 235 Elm Hurst Mansions,	Hawkins, J.
London, SE19, England, The Vicarage, Gipsy Hill,	Wilson, W. P.
Soest, Holland, Middlewijh St.,	Roorda, G.
Bangalore, India, Indian Inst. of Science, Hebbal P. O.,	Sreenivasan, K.
Marton, New Zealand, Mirfield,	Wilde, S. F. M.
Wellington, New Zealand, 47 Austin St.,	Harrison, W. L.

Elected to Junior Grade

Carthage, Illinois, 116 So. First St.,	Alley, K. G.
Highland Park, Illinois, Log House,	Millard, E. L.
Greensburg, Ind., P. O. Box 136,	Fogg, W. C.
Formoso, Kansas,	Compton, R. D.
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New York, N. Y., 224 Bay 10th St.,	Cardea, A.
Rochester, N. Y., 478 Maplewood Ave.,	O'Brien, B. C.
Rochester, N. Y., 189 Fairgate Ave.,	Studier, W. A.
Potosi, Wisc., P. O. Box 165,	Roser, C. O.
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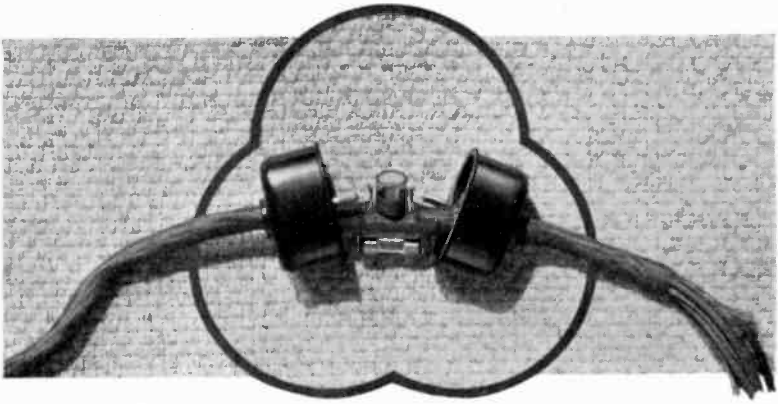
Toronto, Ontario, Canada, 1358 Ossington Ave.,	Mitchell, L. W.
Glasgow, Scotland, Buchanan House, Greenhead,	Ramsay, J. F.
Wellington, New Zealand, Lancaster St., Karori,	Grant, R. L. C.



A fuse housing of Bakelite Molded

Dielectric strength and fine appearance were required in the material used for the housing of the Belden fused radio battery cord. Both were obtained in Bakelite Molded.

The two parts of this housing are completed in a single molding operation, with metal friction pins solidly embedded in the Bakelite. The high lustre is acquired in the mold, and no buffing nor finishing is necessary.

Radio engineers and manufacturers are constantly finding new uses for Bakelite, because it not only possesses the insulating properties essential for radio service, but is also readily molded with extreme accuracy to any desired form.

Bakelite engineers and research laboratories welcome opportunities to cooperate in applying the advantages and economies of Bakelite to the particular needs of any manufacturer. Write for Booklet 33, "Bakelite Molded."

BAKELITE CORPORATION

247 Park Ave., New York, N. Y. Chicago Office: 636 W. 22nd St.
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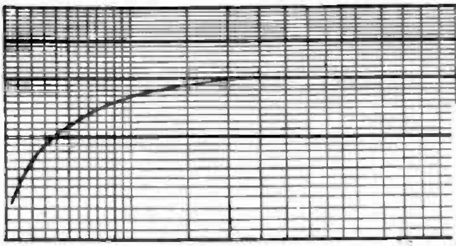
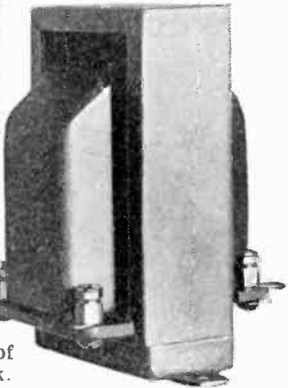
THE MATERIAL OF A THOUSAND USES

"The registered Trade Mark and Symbol shown above may be used only on products made from materials manufactured by Bakelite Corporation. Under the capital "B" is the numerical sign for infinity, or unlimited quantity. It symbolizes the infinite number of present and future uses of Bakelite Corporation's products."



Average voltage amplification curve of the Pacent Superaudioformer No. 27-A in a 201-A amplifier circuit.



Pacent Superaudioformer
No. 27-A

Designed and built by audio frequency specialists

Pacent Superaudioformers are the result of years specializing in audio frequency work. They have been designed to meet in a most efficient manner present day audio amplifying requirements—a high uniform energy amplification covering the entire musical scale without distortion and for use in modern power amplifying circuits where high plate potentials are necessary.

Our engineering department will gladly cooperate with the engineers of set manufacturers toward meeting the audio amplification requirements of their particular circuits. We solicit and welcome this opportunity.

SPECIFICATIONS OF TRANSFORMERS

Input Type No. 27A	Output Type No. 27B	Audio Choke No. 29
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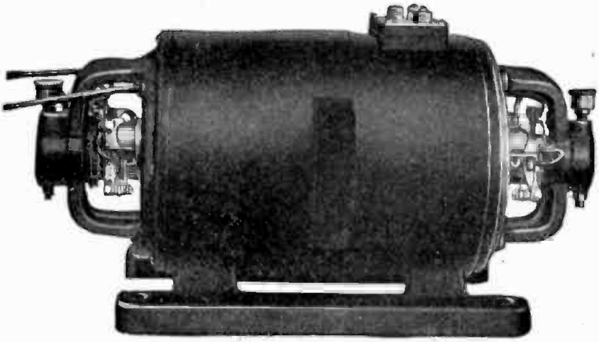

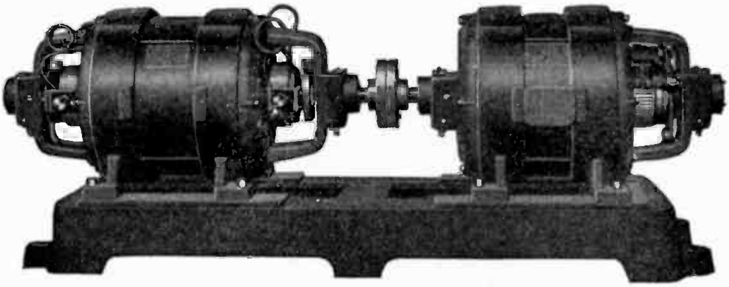

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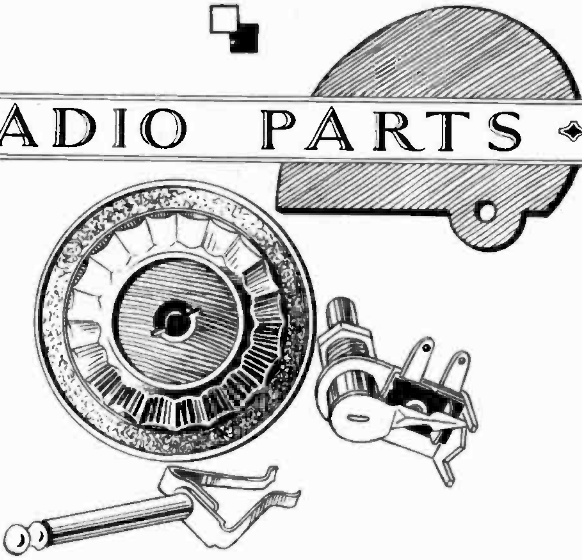

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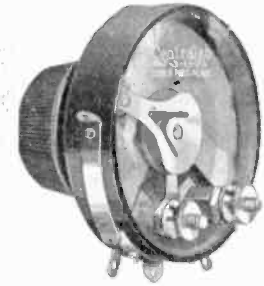

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Alphabetical Index to Advertisements

A	
American Transformer Company	IV
B	
Bakelite Corporation	IX
Burgess Battery Company	VII
C	
Cardwell, The Allen D., Mfg. Corp	XXI
Cunningham, E. T., Inc.,	Inside Front Cover
Central Radio Laboratories	XVIII
D	
Dubilier Condenser and Radio Corp.	Back Cover
Dudlo Manufacturing Corporation	XIII
Duskis Sales Company	XIV
E	
Electric Specialty Company	XII
Electrical Testing Laboratories	XIV
F	
Federal Telegraph Company	X
Farrand Manufacturing Company	XX
Ferranti, Inc.	XIX
G	
General Radio Company	III
Grebe. A. H. & Company	V
H	
Harper, W. W.	XIV
J	
Jewell Electrical Instrument Co.	VI
L	
Larsen, Paul J.	XIV
M	
Minton, John	XIV
P	
Pacent Electric Company	XI
Q	
Q. R. V. Radio Service	XIV
R	
Radio Corporation of America	VIII
Roller-Smith Company	II
S	
Scientific Radio Service	XV
Scovill Manufacturing Company	XVI
T	
Tobe-Deutschmann Co.	XVII
W	
Wireless Specialty Apparatus Company	I
White, J. G., Engineering Corporation	XIV

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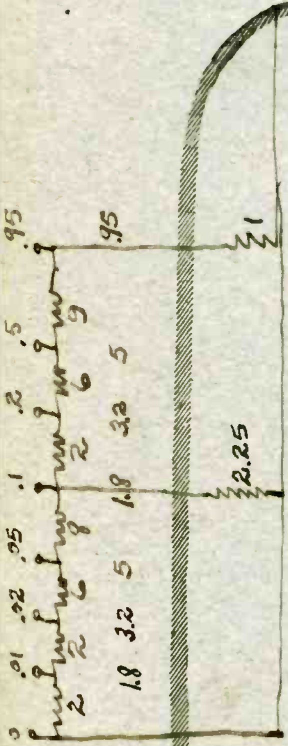

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