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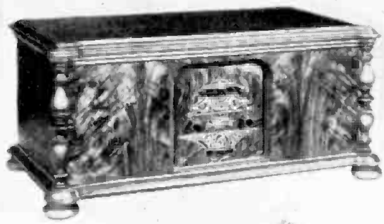

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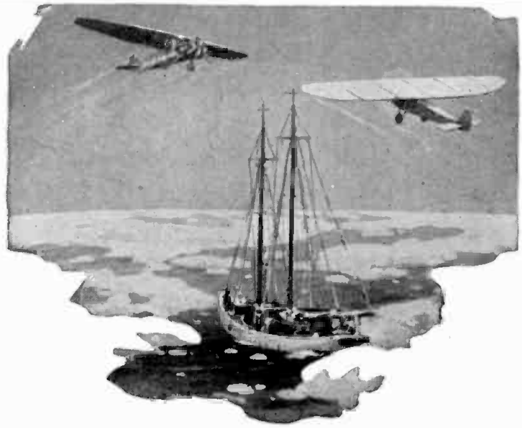

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Volume 15 SEPTEMBER, 1927 Number 9

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The Institute learns with regret of the death on August 16th, 1927 of

Charles V. Logwood

Mr. Logwood, who was born in California in 1882, has long been identified with the development of continuous wave radio reception and the vacuum tube. He was an active worker in the art, and the grantee of a number of radio patents. The sudden curtailment of his efforts is a distinct loss to the radio art and industry.

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

By

GREENLEAF W. PICKARD

(Consulting Engineer, the Wireless Specialty Apparatus Company, Boston, Massachusetts.)

(Communication from the International Union of Scientific Radio Telegraphy.)

This paper is a continuation of an earlier one[1] on the same subject, in which day reception at 15.25 kilocycles, night reception at 1330 kilocycles and night reception at 8-9 megacycles were compared with sunspot numbers and magnetic measures. Although in the former paper definite relations were shown, there were also discrepancies, particularly in the relation of sunspots to reception and magnetism. When the three elements were compared over any considerable period, the maxima and minima swung alternately in and out of step in an irregular seeming manner, although preserving in a general way a 27-day period.

It must be confessed at the outset that the relation between individual sunspots, reception and magnetism is still far from definite. In general it appears that the terrestrial elements are most disturbed when the sunspots are large and numerous, and particularly when they face most nearly earthward. But often large spots cross the meridian without either accompanying magnetic disturbances or changes in reception, and it is well-known that severe magnetic storms sometimes occur when the only sunspots visible are near the sun's limbs.

Undoubtedly both magnetism and reception are subject to disturbances which are not of solar origin, or which at

*Presented before the American Geophysical Union, Section of Terrestrial Magnetism and Electricity, Washington, D. C., April 28, 1927.

Received by the Institute April 20, 1927.

[1]The Correlation of Radio Reception with Solar Activity and Terrestrial Magnetism, Proceedings of the Institute of Radio Engineers, Vol. 15, No. 2, February, 1927.

least do not follow either solar rotation periods or the longer swing of the sunspot cycle. Thus, diurnal periods are found in both reception and magnetism, and seasonal changes are also well known. A solar eclipse distinctly affects radio reception[2] and less definitely disturbs terrestrial magnetism. There also appear to be relations between certain meteorological elements and reception which are not paralleled by magnetic disturbances, probably

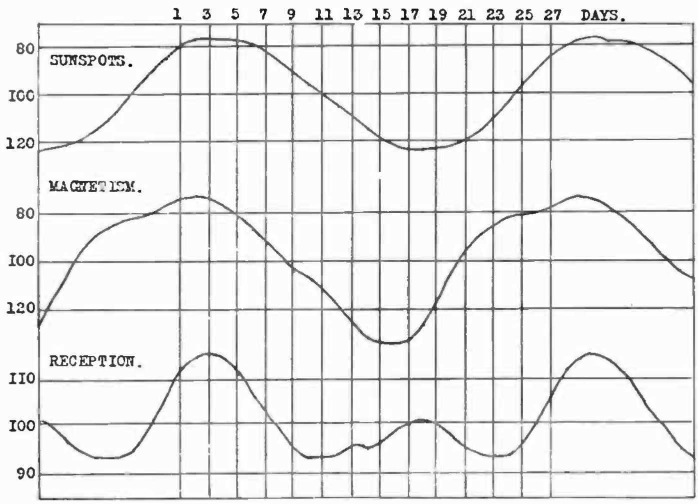


Fig. 1—Mean of 12 Solar Rotations, October 25, 1908 to September 17, 1909. Sunspots, Magnetic Character of Day and Maximum Night Ranges of "DF" Working.

because reception and weather are rather local matters, while terrestrial magnetism is quite literally a world-wide affair.

While reception disturbances of non-solar periods are important and fertile fields for investigation, the present paper must largely confine itself to magnetic and reception changes associated with the solar rotation period, and so these other effects must be eliminated rather than analyzed. This is readily done by taking a sufficiently long series of observations, dividing this series into 27.3 day periods, and finding the daily means. This operation averages out anything which does not recur in each rotation, and gives a

[2]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.

clearer picture of the interrelation than that presented in the former paper.

It is not necessarily true that the solar centers responsible for terrestrial disturbances are the sunspots themselves, although the evidence is now strong that these centers are at least closely associated with sunspots or sunspot groups. If it be assumed that these centers, like

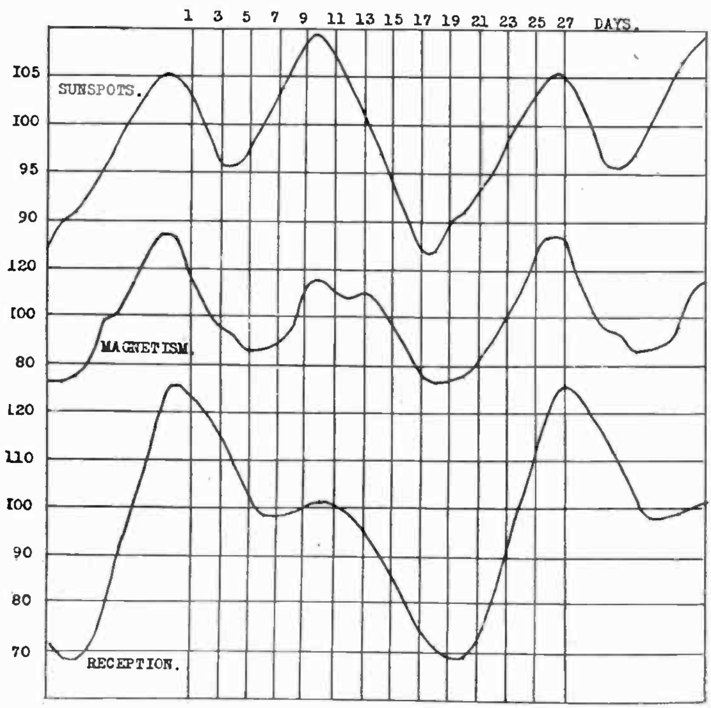


Fig. 2—Mean of 5 Solar Rotations, January 11 to May 30, 1917. Sunspots, Magnetic Character of Day and Washington Day Reception of POZ.

the accompanying sunspots, are confined within definite though rather wide zones for long periods, but shift about within these areas from month to month, some of the discrepancies found can be quite simply explained. And even if, as at present appears, the wanderings of the active centers are entirely at random, the mean of a number of solar rotations should show definite disturbance maxima corresponding with the meridian crossing of the centers of the active zones, instead of the somewhat irregularly spaced disturbances accompanying the transit of the individual centers.

Before periodic means of sunspot numbers, magnetic measures, or reception observations are made, it is usually necessary to reduce the numerical values to such a basis that periods of high absolute values will not dominate the means of a long series. This is easily done by converting the daily values into ratios with a moving mean of the

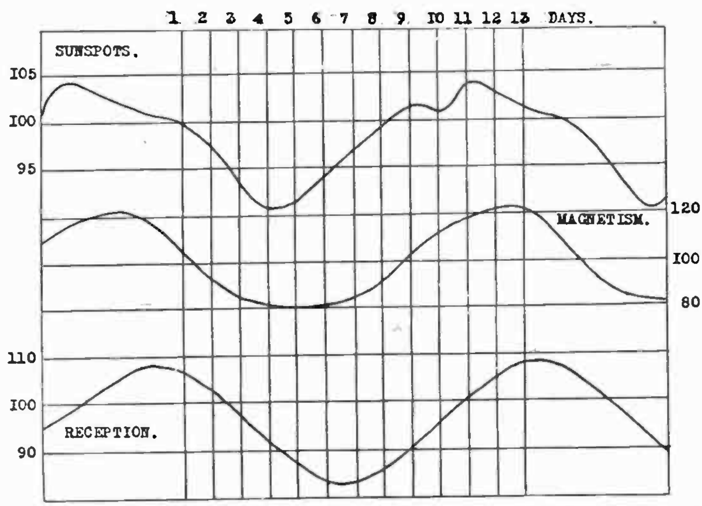


Fig. 3—Mean Within 5 Solar Rotations of Ten 13.64 Day Periods, January 14 to May 30, 1917. Sunspots, Magnetic Character of Day and Washington Day Reception of POZ.

period under investigation, thus obtaining a measure of interperiod activity. By taking a sufficient number of periods, even rough measurements of reception can be made to show definite relations to sunspots and terrestrial magnetism.

In Fig. 1 the reception values are the maximum nightly ranges of two-way working between station DF at Manhattan Beach, New York, and various ships at sea[3], which are compared with the Wolfer Final Sunspot Numbers and van Dijk's Magnetic Character of Day Numbers. The interval taken is from October 25, 1908 to September 17, 1909, corresponding to 12 solar rotations of 27.3 days. It will be seen that the sunspot curve is nearly sinusoidal, indicat-

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

ing that in the sunspot numbers the only frequency of appreciable amplitude is that of the solar rotation, but both magnetic and reception curves show the presence of shorter periods. Owing to the roughness of the reception data, no attempt has been made to investigate the harmonics of the solar period, although this has been done for some of the later and more accurate reception measurements. The

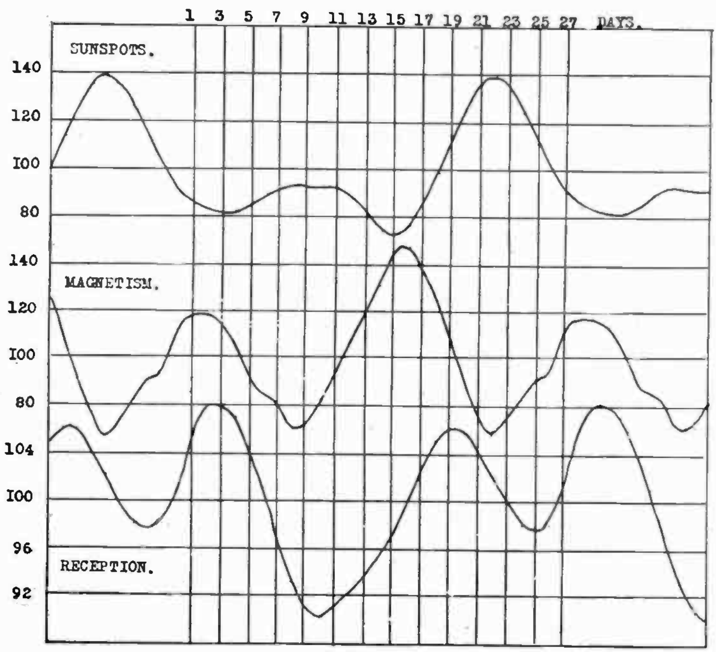


Fig. 4—Mean of 9 Solar Rotations, March 14 to November 15, 1922. Sunspots, Magnetic Character of Day and Washington Day Reception from POZ and LY.

relation of reception to the other elements is inverse, which appears characteristic of night reception save at ultra high frequencies. It is of interest to note that this transmission was all at frequencies within the present broadcasting band, so that the results may be compared with recent measurements, allowance being made for the fact that in 1908-1909 spark transmitters of relatively high decrement were used.

During 1917, a sunspot maximum year, the U. S. Naval Radiotelegraphic Laboratory made audibility meter measurements at Washington of day reception from station POZ at Nauen, Germany. Owing to war conditions, Dr.

Austin considers these the least accurate of his long series of measurements of this station; nevertheless Fig. 2 clearly shows relation between this reception, sunspots and

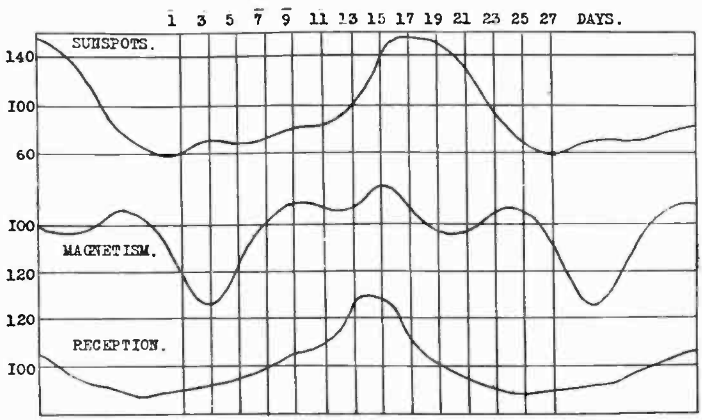


Fig. 5—Mean of 7 Solar Rotations, May 14 to November 21, 1923. Sunspots, Magnetic Character of Day and Washington Day Reception of POZ.

magnetism. All three curves show shorter periods of considerable amplitude, and despite the roughness of the reception measurements, an analysis has been attempted for the second harmonic of the solar rotation period, or 13.6 days. While in Fig. 1 the relation of night reception to sunspots

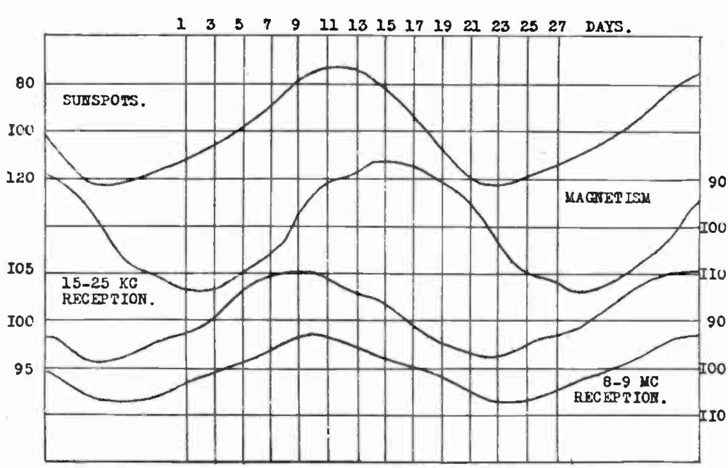


Fig. 6—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham, 15—25 Kilocycle Day Reception at Washington and 8—9 Megacycle Night Reception at Washington.

and magnetism is inverse, in Fig. 2 reception is found to be directly correlated with the other elements; this seems to be the normal relation of day reception at low frequency.

In Fig. 3 means of 1917 day reception, sunspots and magnetism are taken for periods of 13.6 days, or the second harmonic of the solar rotation period. Considering the roughness of the reception data, the curves are in excellent agreement, and their nearly sinusoidal form indicates that there are no important shorter periods present. The low

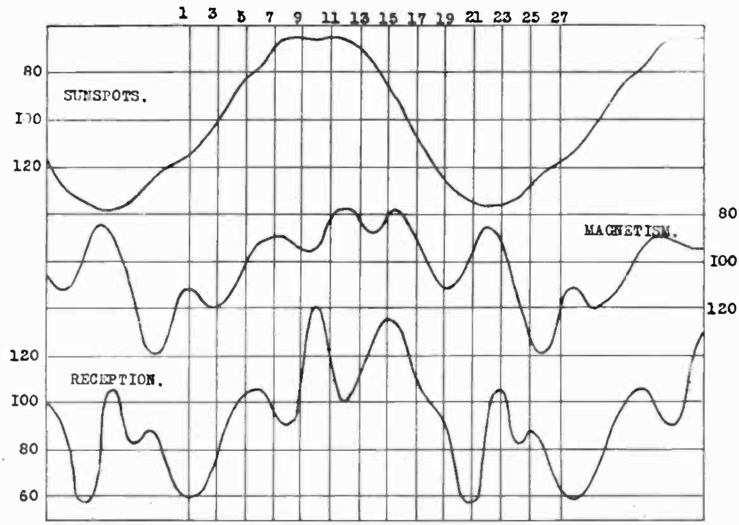


Fig. 7—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Center from WBBM at Chicago. No Smoothing.

amplitude of the sunspot curve is due to the nature of the Wolfer numbers, which are taken over the entire visible solar disk, or over nearly 180 deg. in longitude. This process leaves very little of the shorter periods, although it will be later shown that these are prominent in sunspot numbers taken in restricted zones instead of over nearly a hemisphere.

Figs. 4 and 5 are of Washington day reception for the years 1922 and 1923, the latter year being a sunspot minimum. Again the relation of day reception to solar activity and magnetism is found to be direct, and the presence of shorter and probably harmonic periods is strongly indicated. The irregularity of the 1923 sunspot curve is in part due to the fact that 47 per cent of the days taken had a sunspot number of zero.

In Fig. 6 a comparison is made between sunspots, diurnal range of *H* at Cheltenham, 15-25 kilocycle day reception for a group of 9 stations, and 8-9 megacycle night reception for a group of 7 stations. A close agreement between the two reception curves is evident, but the sunspot and

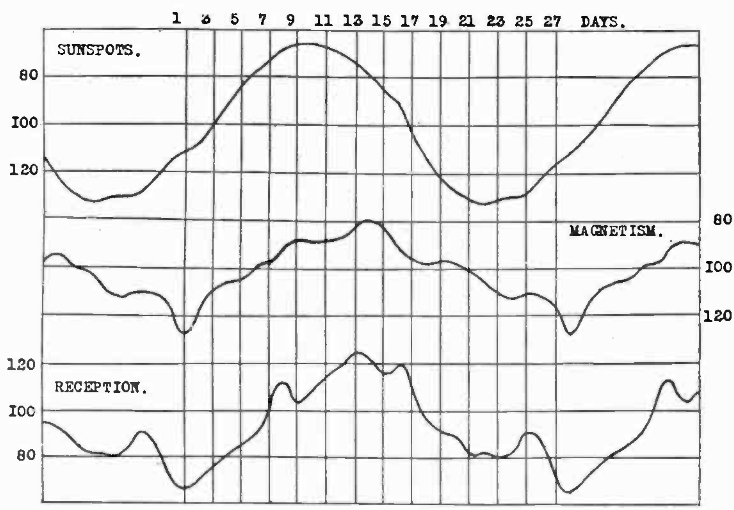


Fig. 8—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Centre from WBBM at Chicago. Smoothed by Moving 7-Day Mean.

magnetic curves lag 4 and 6 days, respectively, behind reception. The relation between sunspots, magnetism and 8-9 megacycle night reception is direct, but is inverted for 15-25 kilocycle day reception. This does not agree with the findings for other years, and apparently 1926 is an exception to a general rule. But the data have been taken from a single station in the preceding years, instead of the group used for 1926, so the apparent inversion may be subject to revision after further study of the data.

As set forth in the previous paper, it appears that reception in the frequency band of 500 to 1500 kilocycles shows higher correlation with solar activity and terrestrial magnetism than does any other investigated portion of the radio spectrum, and for that reason the author's measurements have been largely confined to night broadcast reception. In Fig. 7 are given the means within a 27.3 day period of the Wolfer Provisional Sunspot Numbers, the diurnal range of the earth's horizontal magnetic field as taken by Cheltenham Observatory and mean night field at Newton Centre, Mass., from station WBBM at Chicago, operating at 1330 kilocycles. The sunspot curve is nearly sinusoidal, indicating the absence in the Wolfer numbers of any large amplitude periods shorter than the fundamen-

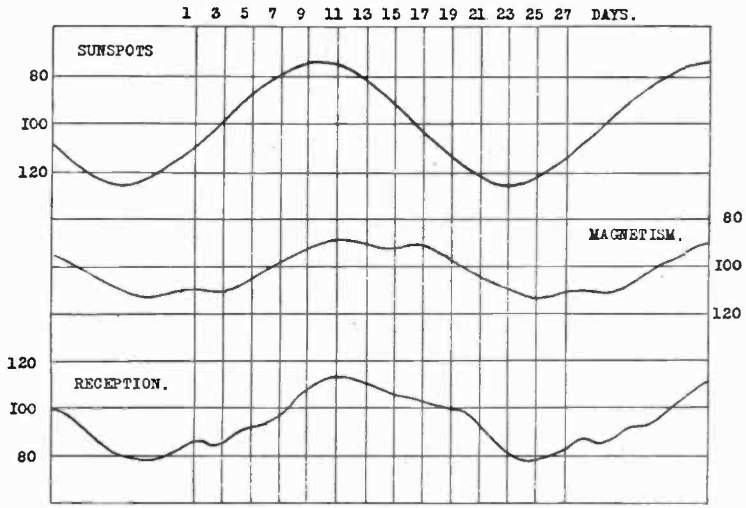


Fig. 9—Mean of 8 Solar Rotations, January 25 to August 31, 1928. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Centre from WBBM at Chicago. Smoothed by 13-day Moving Mean.

tal, but the magnetic and reception curves evidently contain appreciable shorter period components. The fundamental 27.3 day period is best shown by eliminating the shorter periods, which is partially done in Fig. 8 by smoothing with a moving 7-day mean, and more completely in Fig. 9 by a moving 13-day mean, which leaves little but the fundamental.

In Fig. 10 periodic daily means are taken within the second harmonic of the solar rotation period, or 13.6 days. This greatly reduces the amplitude of the sunspot curve, but the reception and magnetic curves still show the presence of shorter periods. In Fig. 11 the third harmonic of 9.1 days is taken, which still further reduces the amplitude of the sunspot curve, leaves the magnetic curve nearly sinusoidal, but the reception curve still shows a shorter period component. Finally in Fig. 12 a 6.8 or fourth harmonic period is used, which reduces the sunspot curve to very small amplitude, while the magnetic and reception curves now become quite similar, with little indication of any shorter period component.

As indicated above, the Wolfer sunspot numbers cannot be effectively used in the investigation of the shorter

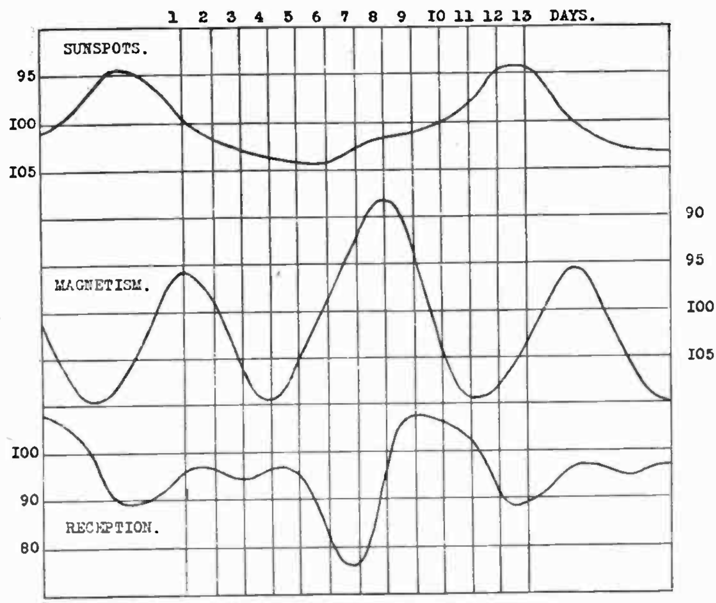


Fig. 10—Mean Within 8 Solar Rotations of Sixteen 13.64 Day Periods, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception at Newton Centre of WBBM.

solar periods. Through the kindness of the U. S. Naval Observatory a list has been given of the times of meridian passage of the larger sunspots and sunspot groups of 1926, and a comparison of this with reception measures shows very clearly that night reception depressions are in general coincident with the meridian transit of the spots, while the appearance of large spots and groups near the sun's limbs had little or no correlation with reception. Through the courtesy of Dr. H. H. Clayton, who has taken daily sunspot observations for the past several years, sunspot numbers for a zone ten degrees on each side of the central solar meridian have been furnished for each day of 1926. Using these sunspot numbers and night reception from WBBM, the relations within the fundamental 27.3 day period, and its second, third, and fourth harmonics are given in Figs. 13 and 14. The correlation between the sunspot numbers for the central twenty degree zone and reception is now found to be very high, and the amplitude of the sunspot curve for even the fourth harmonic of 6.8 days is quite large. Not only are

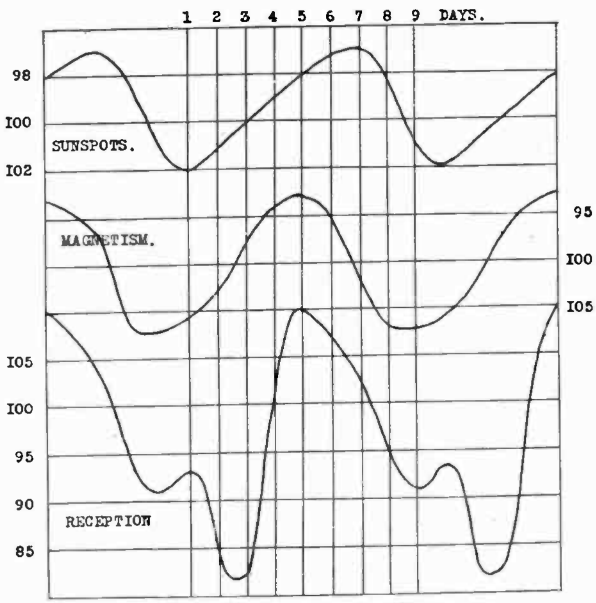


Fig. 11—Mean Within 8 Solar Rotations of Twenty-Four 9.1 Day Periods, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception of WBBM.

the curves closely alike in form, but the maxima and minima agree to the nearest day, which confirms the conclusion reached from the study of the U. S. Naval Observatory data, that reception disturbances generally coincide with the earthward presentation of the sunspots.

Measurements of night static were also taken at 1330 kilocycles, and periodic means made over 14 solar rotations, or from January 25, 1926 to February 11, 1927. The result, as compared with sunspots, is shown in Fig. 15. Unlike the night signal at the same frequency, the relation is direct. An investigation of day static in the 15-25 kilocycle band shows an inverse relation to sunspots, although the agreement is not so good as that shown in Fig. 15. At present there is no satisfactory explanation for this relation, either

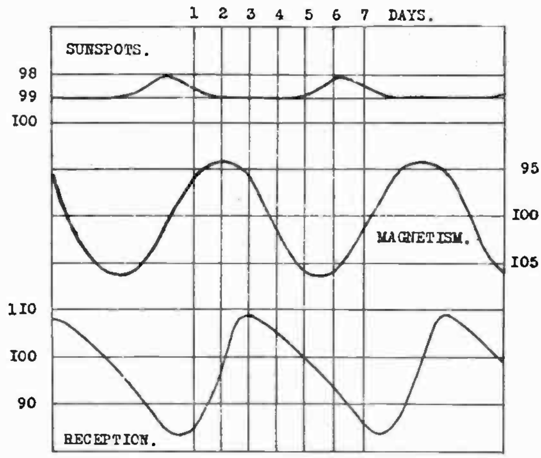


Fig. 12—Mean Within 8 Solar Rotations of Thirty-Two 6.8 Day Periods, January 25 to August 31, 1926. Sunspots, Diurnal Range of H at Cheltenham and Night Reception from WBBM.

on a basis of static wave transmission, or the assumption of electric or other storm areas as static sources. Both in

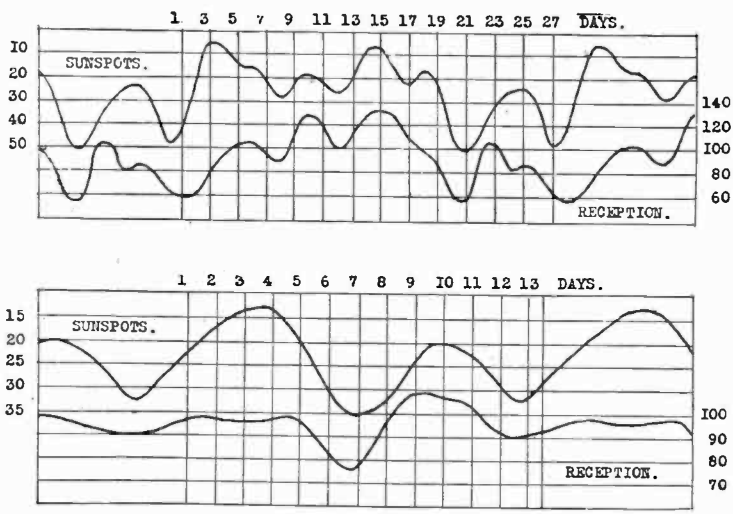
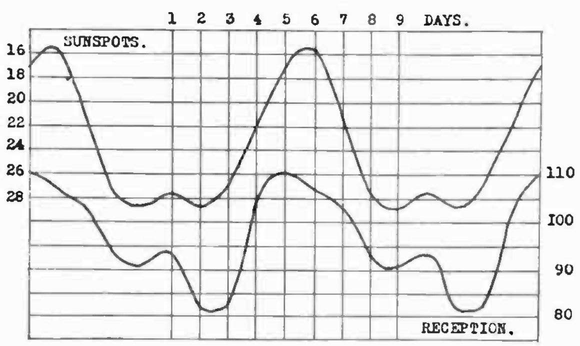


Fig. 13—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots Within Ten Degrees of Solar Meridian and Night Reception of WBBM. Fundamental and Second Harmonic of Solar Reception.

day reception at low frequencies, and in night reception in the broadcast band, the static is inversely related to signal, so it is difficult to frame an hypothesis based on transmission, while if the relation depends upon variability of static sources it must be assumed that the activity of static centers at night is inversely related to that of the day, and at the same time both are linked with solar activity.



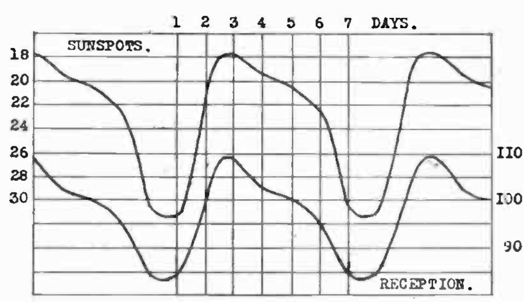


Fig. 14—Mean of 8 Solar Rotations, January 25 to August 31, 1926. Sunspots Within Ten Degrees of Solar Meridian and Night Reception of WBBM. Third and Fourth Harmonics of Solar Rotation.

In conclusion, the existence of a pronounced annual double periodicity reception component, paralleling that already recognized in terrestrial magnetism, is shown in Fig. 16. Here periodic monthly means are made over an interval of 8 years, from 1916 to 1924, and well-defined maxima show at or near the vernal and autumnal equinoxes. The relation between magnetism and reception is direct, for the reception is by day; if such a graph could be made for night reception it would probably show minima at the equinoxes. While it will be several years before really adequate data are accumulated for night reception in the broadcast band, it may be of interest to note that a summation of night spark reception in 1908-1909, scattered observations

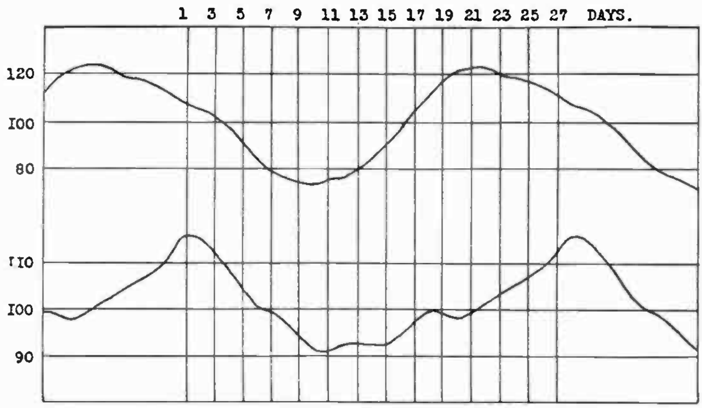


Fig. 15—Mean of 14 Solar Rotations, January 25, 1926 to February 11, 1927. Sunspots and Night Static at 1330 kilocycles.

in 1923-1924 and the WBBM measurements of 1926-1927 shows distinct minima in the spring and fall.

For the convenience of other workers, Dr. Clayton's 1926 sunspot numbers for 20 deg. and 40 deg. central zones are appended. And supplementing the table in the former

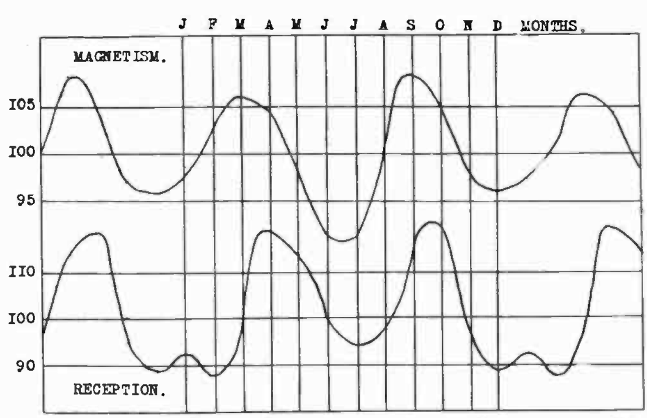


Fig. 16—Mean of 8 years, 1916 to 1924. Monthly Averages of Magnetic Character of Day and Washington Day Reception from POZ.

paper, nightly values of WBBM's field are given from December 15, 1926 to March 31, 1927.

Dr. Clayton has recently found an interesting relation

NUMBER OF SPOTS (UMBRAE) WITHIN 20 DEG. OF CENTRAL MERIDIAN OF THE SUN. ZONE 40 DEG. WIDE.
1926

Day	Jan.	Feb.	Mar.	Apr.	May	June	July	Aug.	Sept.	Oct.	Nov.	Dec.
1	(4)	18	(15)	(0)	0	(8)	19	3	2	1	2	(14)
2	(5)	(10)	40	(0)	7	12	(5)	2	3	(1)	(3)	(12)
3	1	(3)	20	(1)	10	16	7	8	(5)	6	3	10
4	(12)	(0)	27	1	16	9	2	9	3	4	2	5
5	(14)	0	14	1	22	(3)	5	9	(5)	0	0	(7)
6	(14)	(0)	8	1	24	(2)	(7)	4	(10)	(0)	2	4
7	3	(0)	(12)	(0)	6	(4)	(4)	15	10	4	11	5
8	1	0	18	(0)	1	12	(0)	19	2	14	10	0
9	(0)	0	13	0	2	36	(0)	16	0	21	(5)	5
10	7	(0)	13	0	12	(17)	(0)	0	0	(25)	6	(17)
11	16	0	2	2	10	9	0	3	0	17	9	(12)
12	(22)	7	(12)	1	20	3	0	3	3	16	0	12
13	9	(17)	14	2	19	0	0	(1)	5	12	1	1
14	(9)	(22)	7	7	(19)	0	(0)	1	2	11	(3)	16
15	4	14	8	5	0	(7)	(0)	7	3	13	7	(29)
16	5	8	7	8	(0)	14	(0)	(7)	4	10	7	27
17	(4)	9	(8)	8	0	9	0	7	(7)	(11)	8	8
18	(1)	(10)	11	7	0	1	0	2	21	(1)	3	1
19	2	(4)	8	(5)	5	4	0	4	22	(0)	4	1
20	(9)	9	(4)	0	20	10	2	5	24	(0)	(0)	(0)
21	(22)	12	1	0	16	11	0	1	8	(2)	0	0
22	32	(8)	1	0	11	(11)	2	0	1	2	14	12
23	30	0	(1)	0	(0)	(7)	11	0	(3)	3	14	16
24	39	0	1	0	0	1	(15)	3	(9)	(4)	15	23
25	(26)	0	4	0	0	1	20	(8)	4	(7)	2	11
26	(6)	0	(1)	0	(0)	1	1	(8)	(0)	7	(3)	(0)
27	0	(0)	0	0	0	5	5	6	(3)	3	6	0
28	(0)	(0)	3	0	3	8	18	4	(6)	(3)	12	(0)
29	3		6	0	0	25	(29)	1	4	4	(10)	0
30	3		10	(0)	(2)	25	(29)	3	2	9	(9)	2
31	(12)		(7)		(4)		(16)	2	1	(7)		9

Figures in brackets taken from the days preceding or following, allowing for the rotation of the sun.

NUMBER OF SPOTS (UMBRAE) WITHIN 10 DEG. OF THE CENTRAL MERIDIAN OF THE SUN. ZONE 20 DEG. WIDE.
1926

Day	Jan.	Feb.	Mar.	Apr.	May	June	July	Aug.	Sept.	Oct.	Nov.	Dec.
1	(1)	12	(10)	(0)	0	(2)	4	1	2	1	1	(5)
2	(2)	(0)	7	(0)	0	8	(1)	0	0	(1)	(1)	10
3	1	(0)	17	(0)	8	12	0	3	(0)	0	1	0
4	(0)	(0)	15	1	4	1	0	8	3	4	2	0
5	(17)	0	4	0	15	(0)	4	0	(2)	0	0	(7)
6	(5)	(0)	(0)	0	12	(0)	(2)	0	(7)	(0)	1	5
7	0	(0)	1	(0)	2	(1)	(3)	2	8	0	10	0
8	0	0	6	(0)	1	2	(0)	19	0	11	3	(1)
9	0	0	13	0	0	26	(0)	4	0	4	(0)	7
10	0	(0)	6	0	3	(19)	0	0	0	7	6	(5)
11	7	0	0	1	18	2	0	2	0	14	0	(11)
12	(11)	0	(1)	1	3	2	0	2	0	0	0	5
13	6	(8)	13	0	17	0	0	0	4	12	0	(0)
14	(2)	(17)	2	1	3	0	(0)	4	1	4	1	0
15	2	8	5	5	0	(0)	(0)	(7)	2	0	6	(28)
16	3	4	6	2	0	8	(0)	3	4	10	7	15
17	1	8	(3)	3	0	6	0	0	(0)	(3)	6	0
18	(0)	(1)	6	5	0	1	0	0	7	(1)	3	1
19	0	(1)	4	(0)	0	4	0	4	21	0	0	0
20	(2)	5	(1)	0	18	0	0	0	10	(0)	(0)	(0)
21	(16)	11	0	0	15	9	(2)	0	0	(0)	0	(0) (2)
22	21	(3)	1	0	0	(13)	0	0	1	2	3	2
23	12	0	(1)	0	0	(0)	6	0	(1)	0	14	13
24	32	0	0	0 (3)	0	0	(11)	0	(4)	(0)	12	20
25	(3)	(0)	4	0	0	2	0	(2)	1	(4)	0	0
26	(0)	0	(0)	0	0	0	1	(8)	(0)	3	(4)	(0)
27	0	(0)	0	0	0	0	4	0	(0)	(0)	2	0
28	(0)	0	0	0	1	0	0	1	(0)	0	4	(0)
29	2		3	0	0 (3)	18	(13)	0	1	1	(9)	0
30	3		7	0	0	15	(16)	0	1	5	(3)	0
31	(0)		(0)		(0)		0	0		0		5

(1) No observation; number is from preceding or following day for spots which would have been within the central meridian zone on that date.
0 (3) Means that 3 spots developed on that meridian one day later.

Mean field in microvolts per meter at Newton Centre, Mass., from WBBM, Chicago, 9-10 P. M., December 15, 1926, to March 31, 1927.

Day	Dec.	Jan.	Feb.	Mar.
1	...	2.0	1.0	0.1
2	...	3.2	2.4	1.0e
3	...	2.0e	5.2	0.3
4	...	1.5	7.3	3.0
5	...	0.7	5.3	1.5
6	...	8.0	8.0	0.7
7	...	1.3	1.0e	4.0e
8	...	2.2	10.0	5.8
9	...	4.0e	0.3	1.9
10	...	2.0e	0.8	3.3
11	...	4.0e	1.2	1.3
12	...	8.0e	2.6	7.8
13	...	3.0	0.3	4.2
14	...	2.9	5.0m	4.0e
15	3.6	7.9	9.6	2.9
16	2.0	3.3	2.1	0.1
17	3.5	4.0e	3.2	0.1
18	1.5	9.7	1.7	0.7
19	1.4	3.7	1.7	0.1
20	1.0e	10.0	1.6	0.7
21	10.6	10.8	16.0e	0.8
22	12.5	14.7	12.8	0.8
23	0.3	12.3	10.5	3.0
24	2.0	4.0e	0.8	2.0
25	1.9	1.7	1.9	16.0
26	8.2	1.6	6.2	0.5
27	1.0e	4.1	10.7	2.9
28	1.9	4.0e	5.8m	3.2m
29	0.7	5.5		3.5
30	2.7	7.6		0.4
31	6.5	4.0e		0.5
Means	6.0	5.0	4.8	2.5

between night reception of WBBM and the number of sunspots in a central zone of 13.3 deg. (the width of a day's travel in solar rotation). When the number of spots in this zone was under 12, the relation between sunspot number and reception was inverse; when the spot number was over 12, a direct relation was found. The predominant inverse relation which has heretofore been found for night reception is in accord with this, for in two-thirds of the cases examined for 1926, the sunspot number for the zone was under 12. The relation between sunspot number in a central zone, day reception and magnetic measures is now under investigation, and will form the subject matter of another paper.

Summary

The solar, magnetic and reception data from the author's recent paper on this subject are now compared by periodic means, and clearer pictures of their interrelations are given. The solar rotation period and its second, third, and fourth harmonics (27.3, 13.6, 9.1 and 6.8 days) are found for all three elements, the general relation being that sunspots on the solar meridian coincide with disturbances of terrestrial magnetism, lowered night reception and higher day reception. Night static at 1330 kilocycles is found inversely related to night signal reception at the same frequency and therefore directly correlated with sunspots. Day static at 15-25 kilocycles is also, although less definitely, inversely related to the day signal at the same frequency, and therefore in general inversely correlated with sunspots. Periodic means of eight years of day reception show a marked double frequency annual component, with maxima near the vernal and autumnal equinoxes, closely paralleling the well-known annual variation in terrestrial magnetism. Tables of daily sunspot numbers for 20 deg. and 40 deg. central zones are given for 1926, and night reception values for WBBM are continued from the former paper to March 31, 1927.

THE TESTING OF AUDIO-FREQUENCY TRANSFORMER-COUPLED AMPLIFIERS

BY

H. DIAMOND[1] and J. S. WEBB[2]

INTRODUCTION

Any stage of audio-frequency amplification may properly be considered satisfactory if a speech signal impressed upon its input circuit is exactly reproduced, on a larger scale, in its output circuit. For the particular case of transformer coupling, deviation from a faithful reproduction of the applied signal is generally due to one or more of the following four effects:

1. The component frequencies constituting the signal are non-uniformly amplified, causing what may be termed as amplitude distortion.
2. The component frequencies are each shifted in phase while passing through the amplifier, the magnitude of the phase shift angle varying with the frequency of the component considered. The relative phase relationship between the various frequencies making up the reproduced signal is consequently not the same as that between the corresponding components in the original speech signal. This may be considered as causing phase distortion.
3. Harmonics are introduced in the output signal due to the tube characteristics as well as the magnetization characteristics of the iron circuit of the transformer.
4. "Howling" may occur due to the self-generation of audio-frequency oscillations because of circuit instability.

Strictly speaking, the exact performance of a transformer-coupled amplifier is accurately known only when all four of the above effects have been experimentally investigated. For the purpose of the comparison of coupling transformers, however, it is usually sufficient to determine

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quantitatively the degree of amplitude and phase distortion present, making certain qualitatively that the other two effects do not occur in sufficient magnitude to destroy the usefulness of the amplifier. The performance of a given transformer in association with a given tube may therefore be represented graphically by a curve of the type shown in Fig. 1[1], which indicates the variation of amplitude with frequency and, in addition, the dependence of the phase shift

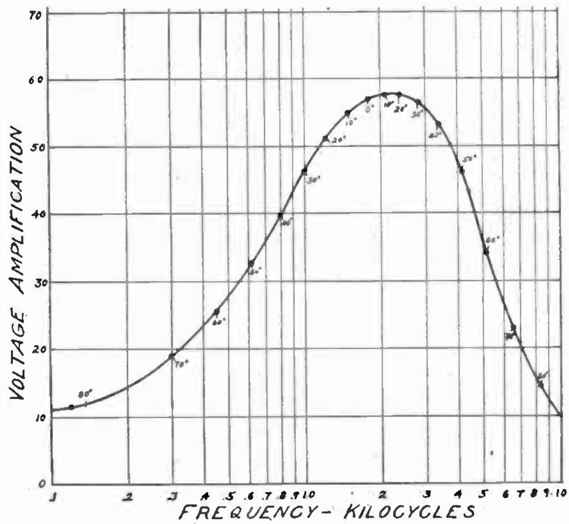

Figure 1—Typical Amplifier Performance Characteristic

angle upon the frequency. Omitting the phase shift angles gives the usual amplitude-frequency characteristic.

The curve of Fig. 1 is of little value, however, unless the conditions of operation under which the amplifier is tested are specified. The dependence of the amplifier performance upon the conditions of its use may be readily understood from a consideration of Figs. 2, 3, and 4. In Fig. 2, an alternating voltage E_{g1} of fixed frequency is applied to the grid of tube A resulting in a fictitious e.m.f. μE_{g1} in its plate circuit. Of this voltage μE_{g1}, a fraction $r_p I_p$ is consumed in the internal plate resistance of the tube and the vector remainder E_{AB} applied to the transformer primary. This is then amplified very approximately in the ratio of turns giving the

[1]See bibliography.

voltage E_{g2} applied to the grid of tube B. The stage amplification may be defined simply as the vector ratio E_{g2}/E_{g1}.

Fig. 3[2] is the approximately equivalent circuit of Fig. 2: —r_p represents the internal plate resistance of tube A; R_1 and X_L, the d.-c. primary resistance and the total leakage reactance, respectively; M the mutual impedance between the transformer windings exclusive of capacity; and C the

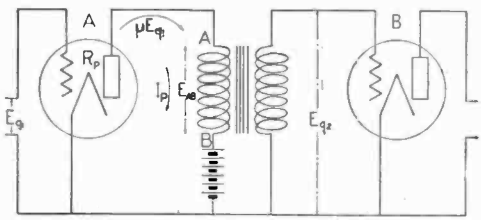

Figure 2—Circuit Diagram for a Transformer-Coupled Amplifier

combined capacitive reactance including the secondary distributed capacity, capacity between windings, etc.; all referred to the primary circuit. The impedance-frequency characteristic of the circuit of Fig. 3, to the right of points A-B, is shown in Fig. 4. This curve was obtained by actual bridge measurements on one of the transformers tested. The

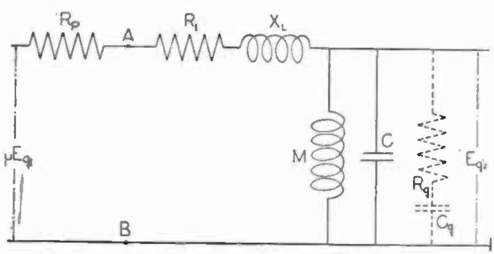

Figure 3—Equivalent Circuit for a Transformer-Coupled Amplifier

impedance Z_{AB} is seen to pass through two resonance points, the first when M and C are in parallel resonance, and the second when X_L is in series resonance with the circuit MC. The ratio E_{g2}/E_{g1} will therefore pass through a maximum value at the first resonance frequency, since Z_{AB} is then very large as compared with r_p; while it may or may not have a second maximum value at the second resonance frequency, depending on whether or not the rise in voltage across C due to

[2]See bibliography.

series resonance is sufficient to compensate for the reduction of the impedance Z_{AB} relative to r_p.

In the above we have neglected the effect of the input impedance of tube B. In everyday use, the tube B will have in its plate circuit a complex impedance, such as another transformer, telephones, loudspeaker, etc. The reaction effect due to this load in the plate circuit may be represented by an input impedance connected between the grid and fila-

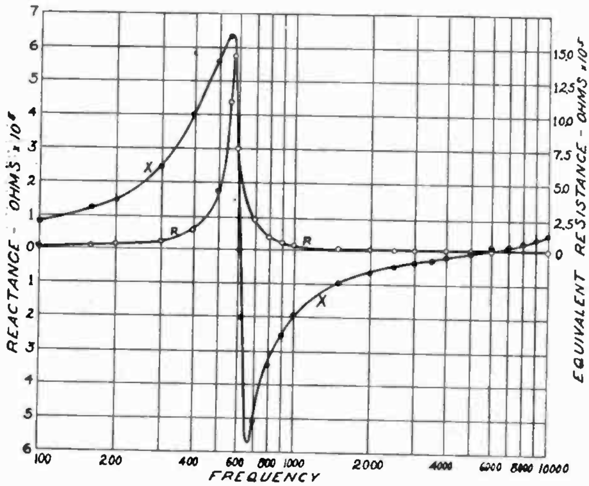

Figure 4—Transformer Impedance-Frequency Characteristic

ment of B and consisting of a resistance R_g and a capacity C_g in series.[3] The sign of this resistance is positive when the load in the plate circuit is resistive or capacitive and negative when the load is inductive. Under normal conditions of use, therefore, there will be a load across the secondary of the test transformer which varies with the frequency. The transformer performance will therefore be altered; C_g causing both resonance points to occur at lower frequencies, and R_g decreasing or increasing the overall amplification depending on its sign.

Since both the internal plate resistance of Tube A and the input impedance of tube B are seen to affect the amplifier performance, the importance of testing under normal operating conditions need hardly be stressed. In the first section of this paper it is proposed to describe several meth-

[3]See bibliography.

ods of test whereby the performance of a given transformer and tube is obtained under actual conditions of use, no matter how varied. Results will be given showing the effect of such variation. In the second section an oscillographic method of test will be described enabling a much more rapid and visual determination of the same results. Being essentially a voltmeter method, however, the degree of phase shift cannot be measured by its means.

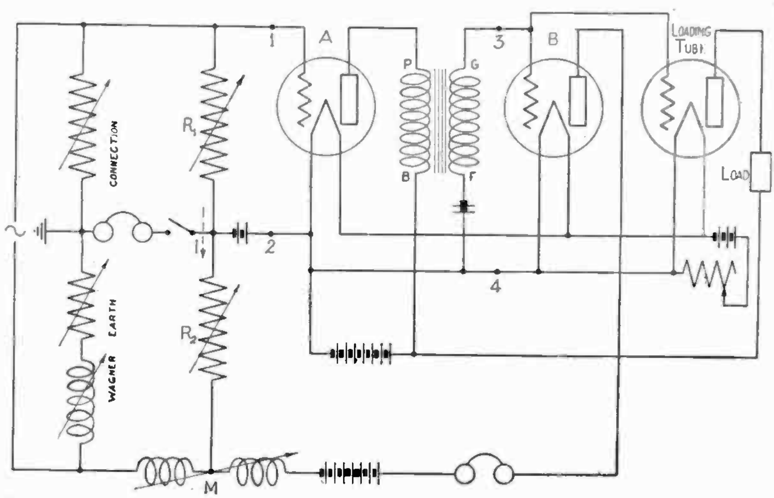

Figure 5—Circuit Arrangement for Measuring Amplification and Phase Shift.

Section 1.—The Measurement of Effective Stage Amplification and Phase Shift Angle

Two alternative circuit arrangements for measuring the stage amplification and angle of phase shift are shown in Figs. 5 and 6. Both are essentially bridge circuits in which the amplifier output e.m.f. is balanced against an auxiliary voltage, the magnitude and phase relationship of which (relative to the amplifier input voltage) is known. In Fig. 5, this balance is effected in the plate circuit of the auxiliary tube *B*, while in Fig. 6 the balancing takes place in its grid circuit. The operation of these circuits may best be explained by means of their vector diagrams. Three such diagrams are given in Fig. 7 illustrating the operation of the circuit of Fig. 5; (a) corresponding to a frequency below the first resonance point, (b) to a frequency between the first and second resonance points, and (c) to a frequency above the second resonance point. Fig. 8 shows a similar set of vector diagrams for the circuit of Fig. 6.

Referring to Fig. 7 (a), $E_{g1} = IR_1$ is the voltage impressed upon the grid circuit of the amplifier tube A, $\mu_A IR_1$ the fictitious e.m.f. in the plate circuit of A and I_p the resultant plate current lagging μ_A IR$_1$ since the equivalent impedance of the coupling transformer (referred to the

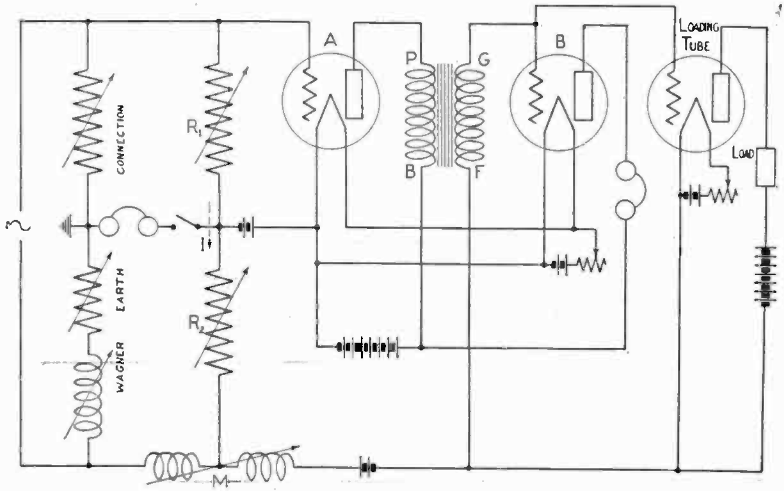

Figure 6—Alternative Method of Test

primary side) is inductive at the frequency considered. The current I_p flowing through the impedance MC results in the voltage drop $E_{g2}{}^1$, which when referred to the secondary side becomes $E_{g2} = nE_{g2}{}^1$ where n is the transformer ratio of turns. This voltage E_{g2} is now impressed between the grid and filament of tube B producing the e.m.f. $\mu_B E_{g2}$ in its plate circuit. Against this e.m.f. is balanced the voltage E_2 made up of IR_2 and IwM. By adjusting R_2 and M a condition of balance may be obtained in the phones, indicating that E_2 is equal and in exact opporition to $\mu_B E_{g2}$. The amplification per stage is then given by the expression

$$K = \frac{\sqrt{(R_2)^2 + (wM)^2}}{\mu_B R_1} \quad (1)$$

and the angle of phase shift by the expression

$$a_0 = tan^{-1} \frac{wM}{R_2} \quad (2)$$

The amplification factor μ_B may readily be determined by disconnecting the apparatus between leads 1-2 and 3-4, and connecting 1 to 3 and 2 to 4, respectively. The circuit is then Miller's dynamic method for measuring amplification constants.

The vector diagram of Fig. 8 (a) differs from that of 7 (a) only in that the voltage E_2 is balanced against E_{g2} rather than $\mu_B E_{g2}$. The expression for the stage amplification is therefore

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{R_1} \tag{3}$$

the phase shift angle being as before, defined by equation (2).

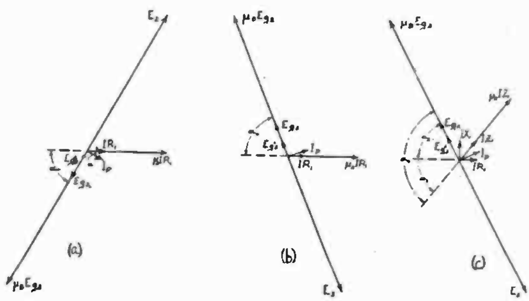

Figure 7—Vector Diagrams for Circuit of Figure 5

For a value of frequency lying between the two resonance points, the vector diagrams must be modified to those of Figs. 7 (b) and 8 (b). The equivalent transformer impedance is now capacitive; the plate current I_p therefore leads the fictitious e.m.f. $\mu_A I R_1$ and the voltage drop E_{g2}^1 leads I_p.

As the frequency is raised above the second resonance point, it will be found impossible to effect a balance with the circuits as shown in Figs. 5 and 6, due to the fact that the phase shift angle becomes greater than 90 deg. A slight modification of these circuits, however, consisting of inserting an inductance L_1 in series with R_1 overcomes this difficulty, as may be seen from the vector diagrams of Figs. 7 (c) and 8 (c). This change in the circuit arrangements results in a corresponding change in the expressions for stage amplification and phase shift. For the circuit of Fig. 5

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{\mu_B \sqrt{R_1^2 + (wL_1)^2}} \tag{4}$$

$$a_0 = tan^{-1}\frac{wL_1}{R_1} + tan^{-1}\frac{wM}{R_2} \qquad (5)$$

while for the circuit of Fig. 6

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{\sqrt{R_1^2 + (wL_1)^2}} \qquad (6)$$

the phase shift angle being given by equation (5).

A consideration of the vector diagrams given above indicates a possible modification of the circuits of Figs. 5 and 6 which may prove quite desirable. For all frequencies above the first resonance point, a variable capacitance standard

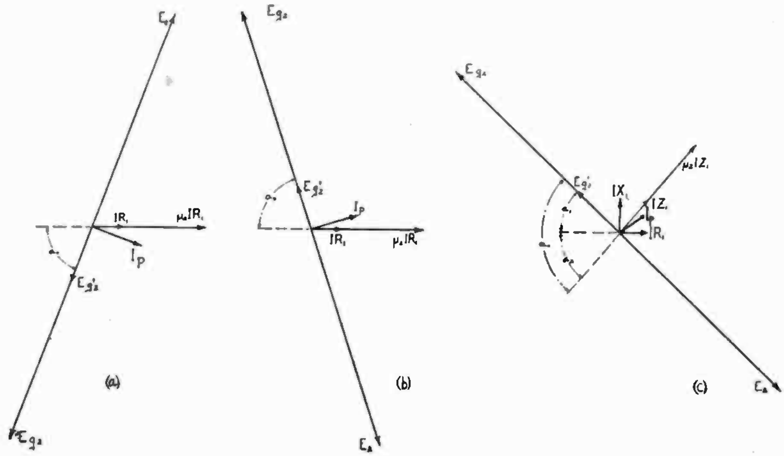

Figure 8—Vector Diagrams for Circuit of Figure 6.

connected in series with R_2 may be used to replace the mutual inductor shown. This is of particular advantage at the higher audio frequencies where errors are possible in the calibration of the mutual inductor due to its distributed capacity between turns, mutual capacity between windings, etc. Where the variable capacitance is used, a modification of the Wagner earth connection is necessary, the inductance shown being replaced by a variable condenser.

The use of the Wagner earth connection[4] is essential if any degree of accuracy is to be expected. This connection insures that the currents in the two arms R_1L_1 and R_2M do not differ from each other because of stray capacity to earth effects, an assumption upon which the expressions derived

[4]See bibliography.

above were based. Even with the Wagner earth, however, considerable care must be taken to prevent direct capacity effects between component parts of the circuit, particularly between the two sets of batteries in the circuit of Fig. 6.

In either of the above two methods, the load on the transformer secondary due to the auxiliary tube *B* is negligible, since at the point of balance no current exists in the plate circuit of this tube. The actual loading is in each case accomplished by connecting the input circuit of a second amplifier across the transformer secondary, as shown, the

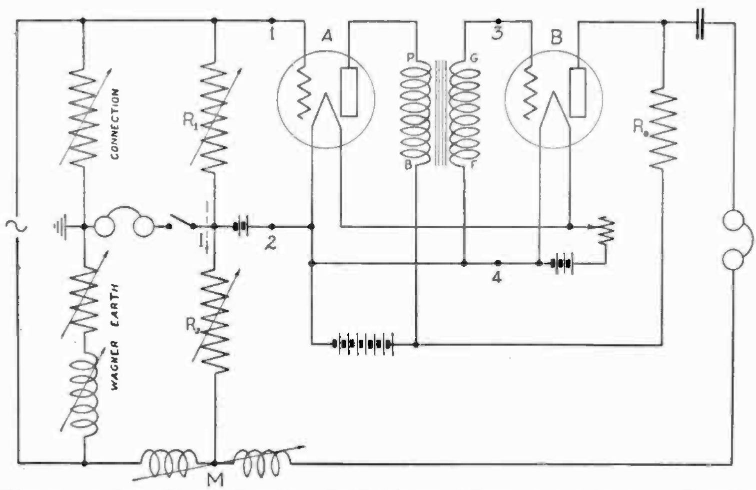

Figure 9—Simplified Test Method When Plate Impedance of Loading Tube Is Resistive.

reaction due to the impedance in the plate circuit of this second amplifier constituting the load. By varying the magnitude and character of this impedance, the effect of varying the reaction upon the amplifier performance may be observed.

For the particular case, where the effect of reaction due to a pure resistance load in the plate circuit of the tube following the transformer is desired, the circuit of Fig. 9 has been found very convenient. This method may be extended to include the reaction effect due to any impedance, whatever, but becomes somewhat involved. In its operation, it is essentially similar to the methods already described. The voltage E_2, due to IR_2 and IwM is here balanced against that part of the driving voltage in the plate circuit of tube

B which is necessary to send the current I_{pB} through the coupling resistance R_0. The expression for the stage amplification and phase shift angle of the test amplifier is respectively

$$K = \frac{\sqrt{R_2^2 + (wM)^2}}{R_1 \left\{ \mu_B \left(\frac{R_0}{R_0 + r_{PB}} \right) \right\}} \tag{7}$$

$$a_0 = tan^{-1} \frac{wM}{R_2} \tag{8}$$

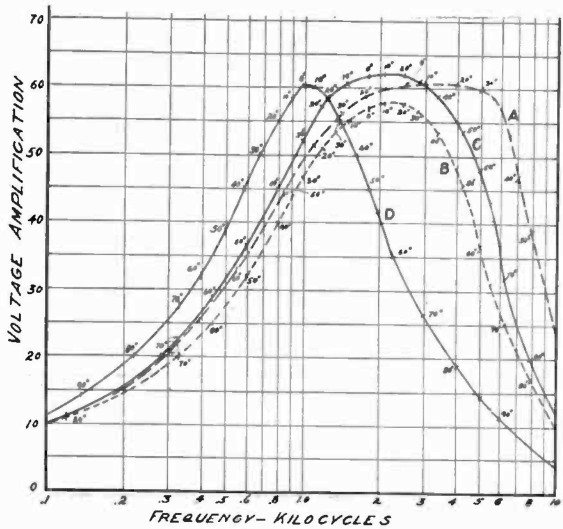

Figure 10—Effect of Load on Amplifier Performance

The value of $\mu_B \dfrac{R_0}{R_0 + r_{PB}}$ does not vary as long as the coupling resistance and plate battery voltage remain unchanged, and may be determined simply by disconnecting the apparatus between leads 1-2 and 3-4, connecting 1 to 3 and 2 to 4, respectively, and with M at zero value adjusting R_2 for a balance in the phones.[5] This quantity is then equal to the ratio of R_2/R_1 when the condition of balance obtains. In all other particulars, the method of Fig. 9 is exactly similar to those already described.

Results

In Fig. 10 are shown the characteristics, under various conditions of load, of a stage of transformer coupling com-

[5]See bibliography.

prising a UX-201A tube and a transformer of rather indifferent design. Curve *A* is for the transformer unloaded, and curve *B* for a reaction load on the transformer secondary produced by a 50,000-ohm resistance connected in the plate circuit of the tube following the transformer. The input capacity C_g of the loading tube, corresponding to 50,000 ohms in its plate circuit, was measured by a resonance method and found to have a value of 85 micro-micro-farads.

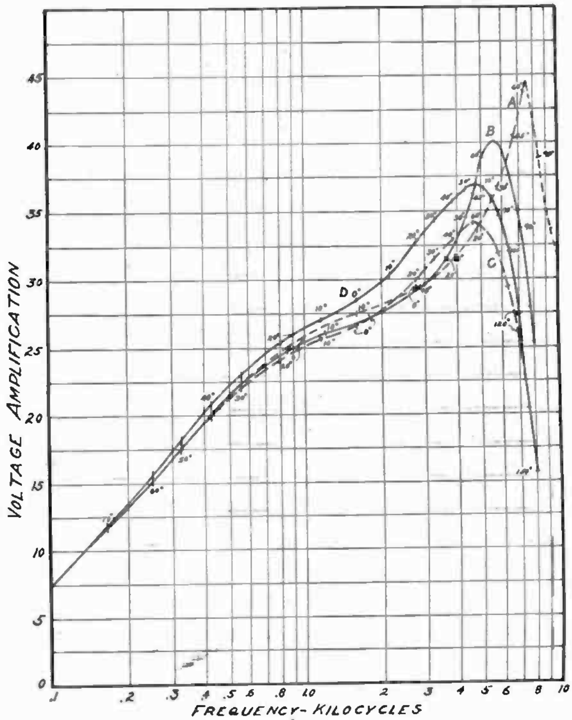

Figure 11—Effect of Load on Amplifier Performance

(This value includes capacities between connecting leads, socket terminals etc., as well as the effect of reaction). Curve *C* was then obtained for an artificial load of 85 $\mu.\mu.$f. across the transformer secondary. A good degree of accordance between curves *B* and *C* will be observed. The lower values of amplification obtained in *B* are no doubt due to the dielectric losses in the tube capacities rather than to the input resistance R_g due to reaction, which is too small to have this effect. Curve D corresponds to an artificial load of 250 u.u.f. across the transformer secondary.

Similar curves are shown in Fig. 11 for a transformer of somewhat different design. Maximum amplification is here seen to occur at the second resonance point. Curve A is for no load on the transformer, and curve B for reaction load due to 50,000 ohms in the plate circuit of the loading tube. For curve C the same resistance was inserted in the plate circuit of the loading tube and in addition a capacity of 8 m.m.f. connected between the grid and plate of this

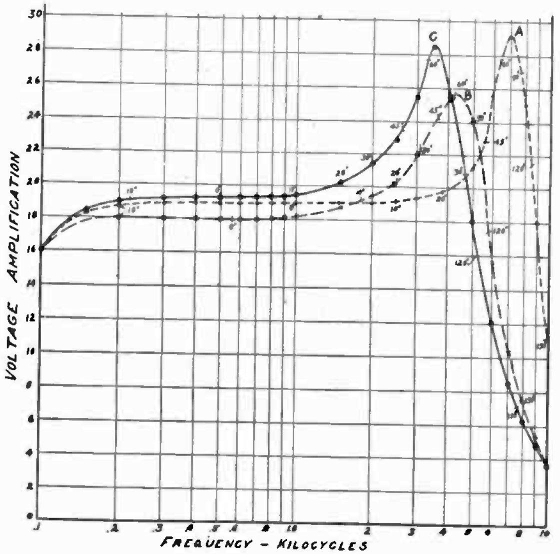

Figure 12—Effect of Load on Amplifier Performance

tube. The input capacity C_g corresponding to this condition was found to be 135 $\mu.\mu.$f. Curve D is for an artificial load of 135 $\mu.\mu.$f. across the transformer secondary.

An additional set of characteristics for an amplifier using one of the better types of transformers is given in Fig. 12. Curve A is the no-load characteristic. To obtain curve B, a resistance of 50,000 ohms was connected in the plate circuit of the loading tube; while for curve C this resistance was replaced by the transformer of Fig. 10 (with its secondary unloaded).

Referring now to Fig. 4, the impedance-frequency characteristic shown was obtained for the transformer of Fig.

12 with an artificial load of 85 $\mu.\mu.$f. across its secondary. The resonance points should therefore occur at the same values of frequency as in curve B of Fig. 12. This is seen to be the case.

The results given above are thought sufficient to illustrate the dependence of the performance of a transformer-coupled amplifier upon the nature of the plate impedance in the tube following the transformer. In general, the no-load characteristic appears to be optimistic, particularly from the point of view of phase distortion. Furthermore, it would

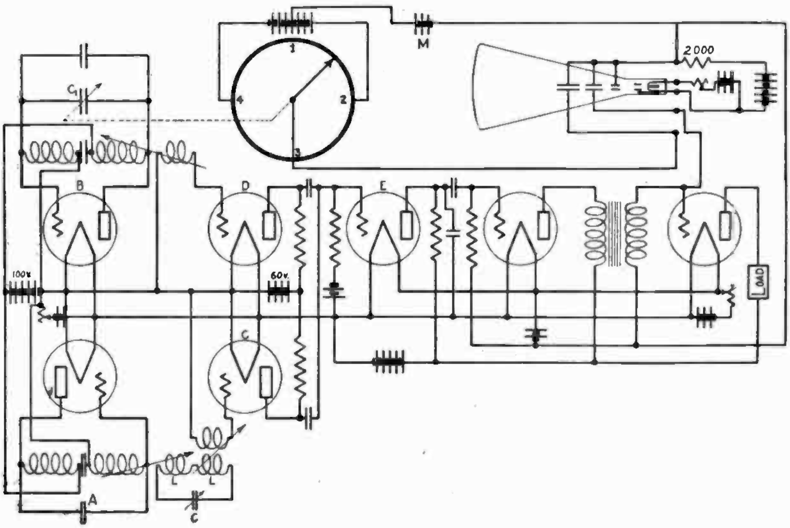

Figure 13—Schematic of Oscillographic Method of Test

seem impossible to predict from this characteristic what the performance under definite loading conditions will be. An accurate specification of the performance of a given coupling transformer in a given circuit can therefore be shown only by its actual operating characteristic as obtained above.

It is realized that there are numerous circuit conditions under which a transformer may be used and that a determination of its performance under all conditions therefore becomes impracticable. This has been used as an argument in favor of testing the amplifier with no load on the transformer secondary. As shown above, however, the no-load characteristic appears of little value.

Merely as a suggestion, the writers point out that the characteristic corresponding to 50,000 ohms resistance in the plate circuit of the tube following the transformer is fairly representative. Moreover, the circuit of Fig. 9 which may be used for obtaining this characteristic is very convenient and quite simple. We therefore prefer this characteristic rather than the no-load characteristic as a standard.

SECTION 2.—OSCILLOGRAPHIC METHOD OF TEST

For a much more rapid determination of the amplitude-frequency characteristic, the circuit of Fig. 13 was devised:—

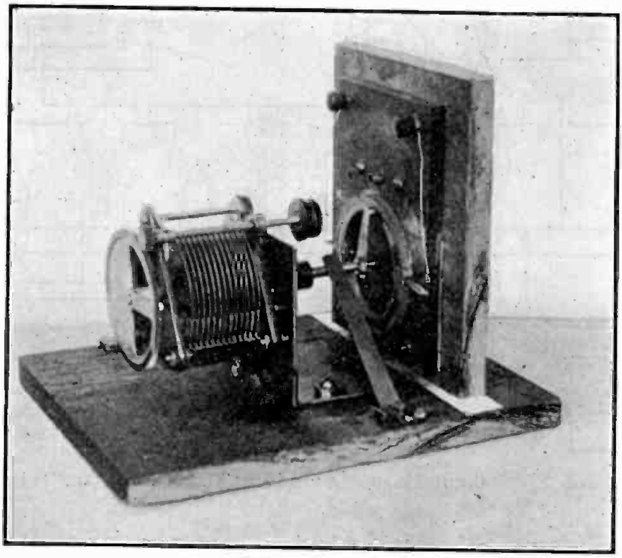

Figure 14—Illustration of Sweep Voltage Arrangement

A heterodyne oscillator is used as the source of a signal note of constant amplitude, but of a frequency varying over the important range of audio-frequencies. This signal is applied to the input circuit of the amplifier under investigation, the output voltage of this amplifier being connected across a pair of deflecting plates of a cathode-ray oscillograph. The electron beam is thus made to vibrate vertically, the amplitude of vibration being a measure of the amplitude of the output signal. Since the input to the amplifier is constant over the complete range of frequencies, the amplitude of vibration will vary with the frequency depending upon the amplifier characteristics.

To the other deflecting plates of the oscillograph is connected a sweep voltage arrangement whereby the electron beam is deflected horizontally, the sweep voltage being so adjusted that the spot is at the extreme left whenever the signal frequency is a minimum and at the extreme right whenever the signal frequency has a maximum value. The pattern appearing on the screen is then the desired amplitude-frequency characteristic and may be photographed or studied visually.

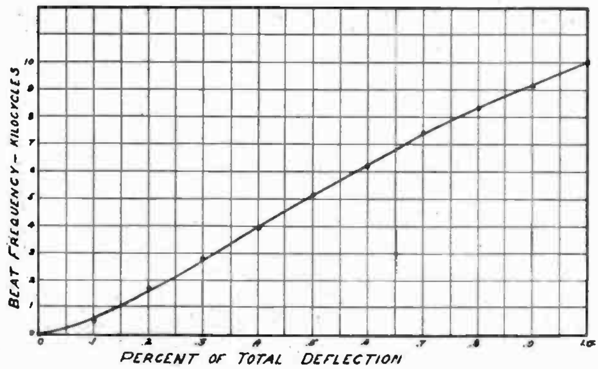

Figure 15—Frequency Calibration Curve for Heterodyne Oscillator

Just as in the test circuits of Section 1, the effect of loading the transformer secondary due to reaction may be obtained by means of an auxiliary amplifier having in its plate circuit any desired impedance.

A detailed description of the circuit of Fig. 13 is given below.

The Heterodyne Oscillator[6,7]. The two Hartley oscillators A and B, their associated amplifiers C and D, and the detector tube E together make up the heterodyne oscillator which is used as the variable frequency source. Oscillator A is adjusted to 100,000 cycles and is coupled to its amplifier C by means of the tuned circuit LCL. Oscillator B has a fixed coil and fixed condenser of such values, that by varying the 500 $\mu.\mu.$f. condenser C_1 from its minimum to its maximum setting, the oscillator frequency is

[6]See bibliography.
[7]See bibliography.

reduced from 100,000 to 90,000 cycles. B is inductively coupled to its amplifier D.

The output of each of the amplifiers C and D is impressed between the grid and filament of the detector tube E by means of resistance-capacity coupling. The frequency of the resultant signal in the detector output circuit is, of course, the "beat" frequency of the two oscillators and should vary from 0 to 10,000 cycles as the dial of the variable condenser C_1 is tuned through an angle of 180 degrees.

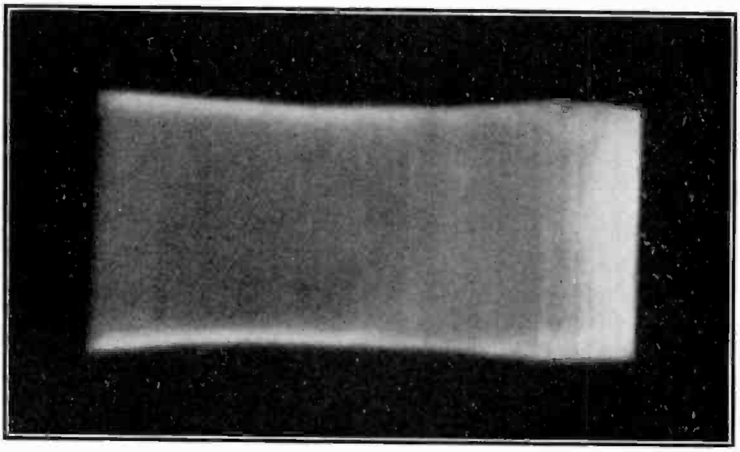

Figure 16—Amplitude-Frequency Characteristic of Heterodyne Oscillator

It is this signal of variable frequency which is impressed upon the amplifier input circuit as shown.

Actually, there is a tendency for the two oscillators to pull into step at the lower frequencies. The function of the amplifiers C and D and of the tuned circuit LCL is to prevent this action, a lower limit of 50 cycles per second being easily obtained.

The tuned circuit has, also, two additional advantages. Kirke has shown that in the case of two carrier waves impressed upon a non-linear detector, if the two waves are of unequal amplitude the stronger wave alone being sufficient to swing over the non-linear portion of the characteristic and reach considerably along the linear portion, the resultant detector current is proportional to the weaker wave, being unaffected by a change in amplitude of the stronger wave. By making the double coupling through the tuned circuit quite loose and applying this loose coupling to the fixed oscillator A, we may make the amplitude of the fixed oscillator weaker than the weakest amplitude of the variable oscillator. A constant output is thus assured which is independent of the frequency.

Kirke has also shown that if one r.f. oscillator contains no harmonics, none will appear in the audio-frequency output except as introduced by the detector tube itself. The tuned circuit here serves to eliminate all harmonics from the fixed oscillator. The harmonics due to the detector itself

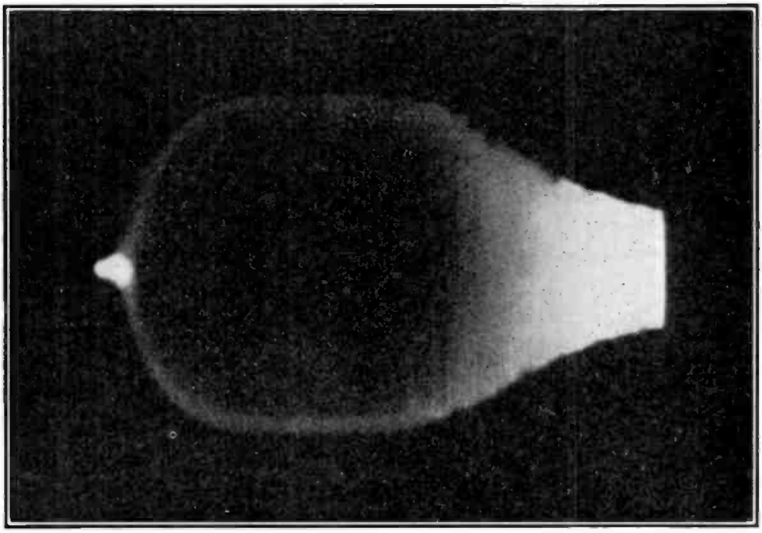

Figure 17—Oscillogram Corresponding to Curve A (Figure 10)

are minimized by using plate rather than grid rectification.

The Sweep Voltage Arrangement. As noted above, the function of the sweep voltage is to deflect the electron beam uniformly from left to right as the frequency is increased from 50 to 10,000 cycles and back to the left as the frequency is decreased to its minimum value. Furthermore it is necessary that both the change in frequency and the rate of deflection of the beam are exactly equal in order that the figure be plotted to a linear scale of abscissas, which, as may be seen from Fig. 15, is also approximately a linear scale of frequencies. Again, the spot must be in exactly the same position for a definite frequency whether the heterodyne oscillator frequency is increasing or decreasing at the instant considered. This is necessary in order that the pattern may exactly retrace itself.

To comply with the above conditions, the arrangement shown in Figs. 13 and 14 was devised. Mounted on the same shaft as the variable condenser C_1, but insulated from it by means of a hard-rubber insulating joint, is a movable contactor arm *A* making a rubbing contact with a circular carbon annulus of rectangular cross-section. This carbon annulus was constructed by placing end-to-end and joining together two circular half-rings each having a resistance between end surfaces of approximately 10,000 ohms. A 70-

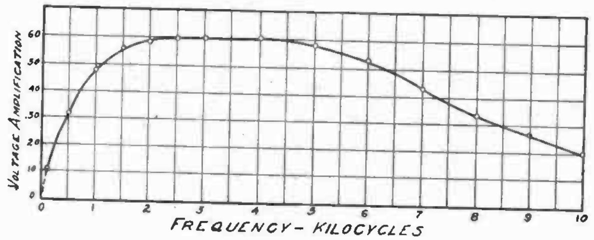

Figure 17a—Curve A Replotted.

volt battery is connected to diametrically opposite points on the circular annulus (to the joints for convenience) and two leads, one from the center in this battery and the other from the movable contactor arm, are brought out to the deflecting plates of the oscillograph. This constitutes the sweep voltage.

With the movable contactor arm at point 1, the voltage across the deflecting plates is zero. The beam should then be in the exact center of the screen, and may be so adjusted, by means of the auxiliary battery *M*. With the contactor arm at 2 the deflecting voltage is +35 volts, at 3 zero, and at 4 —35 volts. The electron stream is thus moved continuously from side to side, simply by revolving the contactor arm at a uniform rate.

The position of the condenser on the shaft is now so adjusted that its minimum setting occurs with the contactor at point 4 and consequently its maximum setting with the contactor at 2. The oscillator frequency thus varies from a minimum to 10,000 cycles and back as the spot travels from left to right and back.

For an exact retracing of the figure, the capacity of the variable condenser should vary linearly over its entire range. Moreover the variable condenser should have as small a zero reading as possible. The condenser used was a General Radio Type 247-H of the geared vernier type, a groove being cut in the vernier knob which served as a driving pulley for the condenser and also the movable contactor arm. A reduction in speed was thus obtained permitting the use of a normal speed d-c. or a-c. motor.

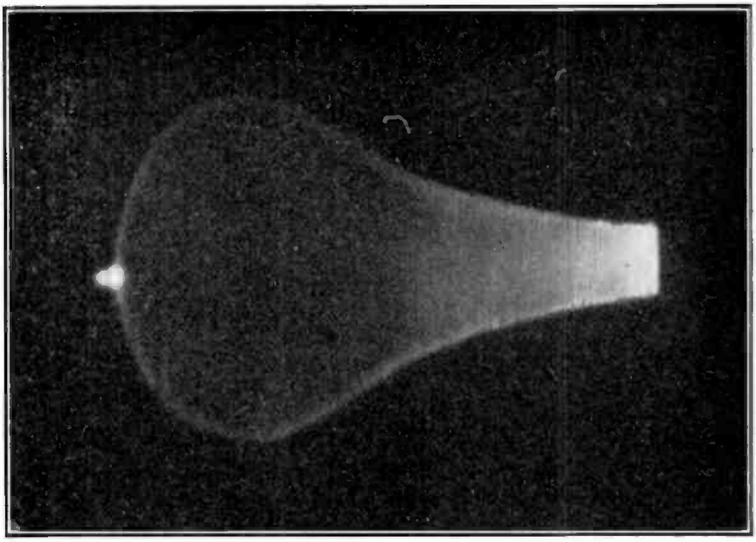

Figure 18—Oscillogram Corresponding to Curve B (Figure 10)

Oscillator Characteristics. Even though an exact retracing of the pattern on the screen is obtainable, does not imply, however, that the curve obtained is plotted to a linear scale of frequencies. The extent of deviation from a linear scale is of course of considerable importance in the analysis of any curve obtained by this method. The calibration curve of Fig. 15 plotting the beat frequency against horizontal deflection on the screen indicates however that the departure from a linear scale is fairly small.

In Fig. 16 is shown an oscillogram of the variation of output amplitude vs. frequency for the heterodyne oscillator. The amplitude is seen to be remarkably constant for the entire frequency range.

Results

The oscillograms chosen for reproduction here are for the transformer of Fig. 10. Figs. 17, 18, and 19, correspond to curves *A*, *B*, and *D*, respectively. For the purpose of comparison these curves are replotted here to the same scale as their corresponding oscillograms. A very good agreement is observed.

APPENDIX

As we have already pointed out, the amplitude-frequency characteristics as obtained by the oscillographic method of test are plotted to an approximately linear scale of frequen-

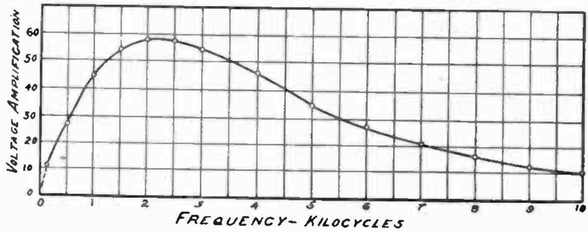

Figure 18a—Curve B Replotted

cies. A slight modification of the sweep voltage arrangement, however, will enable the determination of these curves to a logarithmic scale of frequencies. This modification consists simply of varying one dimension of the circular carbon annulus.

A consideration of the operation of the sweep voltage device indicates that one half-ring is in operation during the time that the electron beam is being swept from left to right, the other half-ring being in use on the return sweep. We may therefore consider each half-ring alone. (see Fig. 20).

Let w be the width of the ring
t its thickness
l its length along a mean radius
and x the distance along its mean radius from the left end to the position of the contactor arm.

On the assumption of a linear characteristic for the variable condenser C_1 and a constant speed of rotation, the following equations may be set down.

$$C_x = 500 \cdot \frac{x}{l} \tag{9}$$

Where C_x is the capacity of the variable condenser when the contactor arm is at the point x.

$$D_x = K_2 R_x \tag{10}$$

D_x being the deflection of the electron beam as measured from its extreme left position,
R_x the resistance of the half-ring from its left end to the point x and

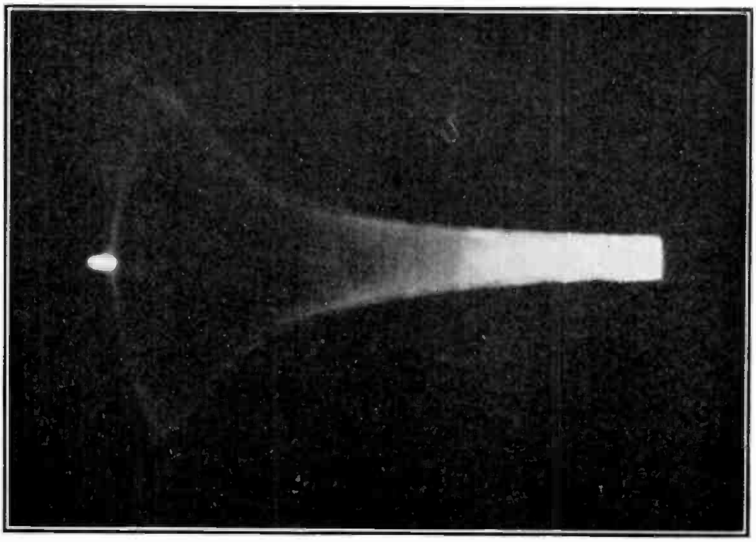

Figure 19—Oscillogram Corresponding to Curve D (Figure 10)

K_2 an arbitrary constant.

$$dR_x = \frac{e \cdot dx}{w \cdot t\,(x)} \tag{11}$$

where dR_x is an element of resistance at point x.
Note that the thickness of the ring is here taken as a function of x. It is this dimension of the ring which is to be varied in order to obtain a logarithmic scale of frequencies.

$$f_x = 100{,}000 - \frac{1}{2\pi\sqrt{K_1 L\left(2500 + 500\,\frac{x}{l}\right)}} \tag{12}$$

Here f_x is the value of the heterodyne frequency when the contactor is at the point x.

In order that a logarithmic scale of frequencies may obtain, the deflection of the beam must vary as the logarithm of the frequency; that is,

$$D_x = K_3 \log f_x \tag{13}$$

Where K_3 is an additional constant.

Substituting equation (10) and (12) in (13) and differentiating, we obtain,

$$\frac{d\,Rx}{dx} = \frac{K}{l} \times \left\{ \frac{1}{\left(2500 + 500 \cdot \frac{x}{l}\right)\left(1 - 2\pi \cdot 10^5 \sqrt{K_1 L} \sqrt{2500 + 500 \cdot \frac{x}{l}}\right)} \right\} \tag{14}$$

Where dR_x is defined by expression (11). After due substitution, equation (15) is obtained.

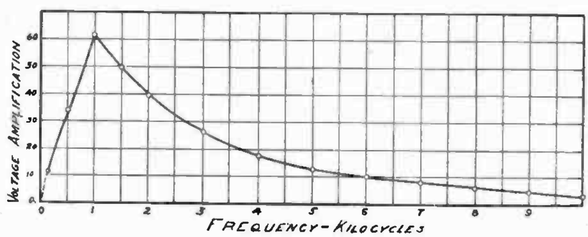

Figure 19a—Curve D Replotted

$$t(x) = \frac{e\,l}{w\,K} \left\{ \left(2500 + 500 \cdot \frac{x}{l}\right) \left(1 - 2\pi \cdot 10^5 \sqrt{K_1 L} \sqrt{2500 + 500 \cdot \frac{x}{l}}\right) \right\} \tag{15}$$

This is the necessary relation for the thickness of the carbon half-ring as a function of the distance x measured along a mean radius from the left end.

The constants of equation (15) may be evaluated by substituting for terminal conditions.

When $x = 0$

$f_x = 100{,}000$

and consequently

$$\sqrt{K_1 L} = \frac{10^{-7}}{\pi}$$

When $x = l$

$$t(l) = \frac{e\,l}{w\,K}\left\{-(3000)\left(1 - \frac{\sqrt{3000}}{50}\right)\right\} = -288$$

$$\frac{e\,l}{w\,K} = -288\,K_4$$

therefore

$$K_4 = -\frac{t(l)}{288}$$

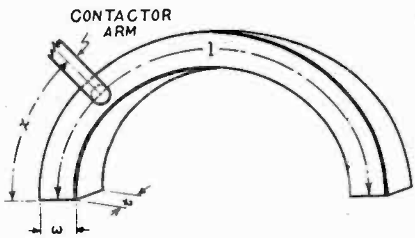

Figure 20—Carbon Half-Ring

Equation (15) may therefore be written

$$\frac{t(x)}{t(l)} = -\frac{1}{288} \times$$

$$\left\{\left(2500 + 500 \cdot \frac{x}{l}\right)\left(1 - .02\sqrt{2500 + 500 \cdot \frac{x}{l}}\right)\right\} \tag{16}$$

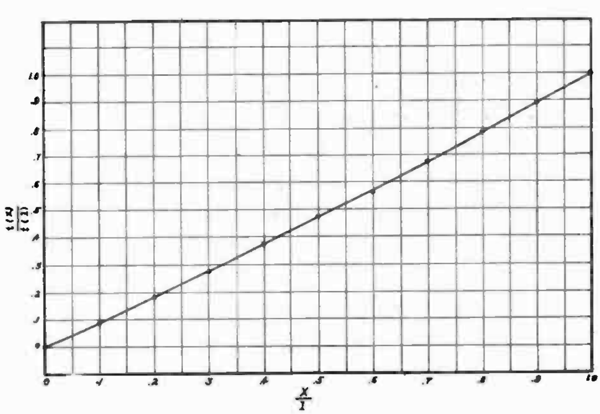

Figure 21—Computed Variation of t (x) to obtain Logarithmic Scale of Abscissas.

This equation is plotted in Fig. 21 and is seen to be linear to a fair degree of approximation. To the same degree of approximation, therefore, we may expect eo obtain a logarithmic scale of frequency merely by tapering the thickness of the annular half-ring from zero at its left end to its normal value at the right end. This obviously applies equally as well to the other half-ring.

SUMMARY

The performance of an audio-frequency transformer-coupled amplifier is considerably affected by the reaction load across the coupling-transformer secondary due to the impedance in the plate circuit of the tube following the transformer. The effect of such reaction is discussed in this paper, and several methods of test described whereby the actual performance of a given amplifier under any condition of loading (due to reaction) may be measured.

The first two methods of test are essentially bridge circuits in which the amplifier output e.m.f. is balanced against an auxiliary voltage, the magnitude and phase relationship of which (relative to the amplifier input voltage) is determinable. To effect this balance an auxiliary tube with phones in its plate circuit is used; the balancing taking place in the plate circuit of this tube in the first method; and in its grid circuit in the second method. The reaction load across the coupling transformer is in each case obtained by connecting across its terminals the input circuit of a second amplifier, having in its plate circuit any desired impedance.

In the third method of test, the second amplifier tube serves both for loading and for balancing, the auxiliary balancing tube being omitted. This circuit is applicable to the particular case where the effect of reaction due to a pure resistance load in the plate circuit of the tube following the transformer is desired and for this case is very convenient and simple.

The fourth test method is oscillographic. A heterodyne oscillator is used as the source of a signal note of constant amplitude, but of a frequency varying over the important range of audio-frequencies. This signal is applied to the input circuit of the amplifier under investigation, the output voltage of this amplifier being connected across a pair of

deflecting plates of a cathode-ray oscillograph. The electron beam is thus made to vibrate vertically. To the other deflecting plates of the oscillograph is connected a sweep voltage arrangement whereby the electron beam is deflected horizontally, the sweep voltage being so adjusted that the spot is at the extreme left whenever the signal frequency is a minimum, and at the extreme right when the signal frequency has a maximum value. The pattern appearing on the screen is then the desired amplitude-frequency characteristic. With a slight modification of the sweep-voltage device, this characteristic may be obtained to a logarithmic scale of frequencies. The loading is obtained in the same manner as in the methods already described.

BIBLIOGRAPHY

1. "Low-Frequency Intervalve Transformers," P. W. Willans, Journal I. E. E., 1926, Vol. 64, p. 1065.

2. "Telephone Transformers," W. L. Casper, Journal A. I. E. E. 1924, Vol. 43, p. 196.

3. "Dependence of the Input Impedance of a Three-electrode Vacuum Tube upon the Load in the Plate Circuit," J. M. Miller, Bureau of Standards, Scientific Paper No. 351, Nov. 1919.

4. "The Measurement of Dielectric Losses by Means of the Wheatstone Bridge," K. W. Wagner, The Electrician 1911, Vol. 68, p. 483.

5. "Notes on the Input Impedance of Vacuum Tubes at Radio Frequency," J. Weinberger, I. R. E. 1920, Vol. 8, p. 334.

6. "The Frequency Characteristics of Telephone Systems and Audio-Frequency Apparatus," B. S. Cohen, A. J. Aldridge, and W. West, Journal I. E. E. 1926, Vol. 64, p. 1023.

7. "Design of a Low-Frequency Generator," H. L. Kirke, Experimental Wireless & Wireless Engineer, Feb. 1927, Vol. 4, No. 41.

NOTE ON DETECTION BY GRID CONDENSER AND LEAK

By

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Introduction

When grid condenser and leak are used for detector the grid is usually kept somewhat positive. This may make the grid-to-filament a.-c. resistance low compared with the external leak resistance. As the two resistances are effectively in parallel, both must be taken into account. A formula is derived showing the variation of detected current with frequency of modulation.

Fig. 1 shows connections commonly used for rectification by means of a grid condenser and leak. The plate circuit

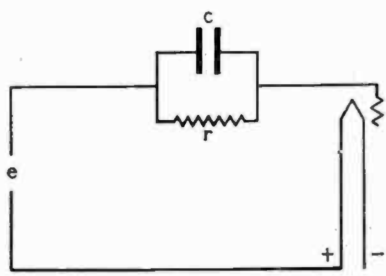

Figure 1

is not shown, as it should be so adjusted that it merely repeats the audio-frequency variation of grid potential.

Fig. 2 shows the circuit in a more general form, a battery being included to indicate that the grid return may be connected to a point of any desired potential. The impedance Z may be of any nature in the following analysis, which presupposes weak signals.

*The method used is essentially that of Carson. (Proc. of the Institute of Radio Engineers, Vol. 7, No. 2) in a paper entitled "A Theoretical Study of the Three Element Vacuum Tube."

Let the applied voltage consist of two components, a voltage $e_0 \cos \omega_0 t$ due to the carrier, and $e_1 \cos \omega_1 t$ due to some typical side frequency. The resulting current through Z will have a number of components, but the only one that interests us is the component whose frequency is the difference between the two applied frequencies. The calculation* of this component is somewhat complicated, and for the present purpose it is sufficient to state that the magnitude of the difference frequency current varies inversely as $\left(1+\frac{Z_0}{R}\right)\left(1+\frac{Z_1}{R}\right)\left(1+\frac{Z_{1-0}}{R}\right)$ where R is the a.-c. resistance between grid and filament, Z_0 is the impedance of Z at carrier frequency, Z_1 is its impedance at the typical side frequency, and Z_{1-0} is its impedance at the difference frequency.

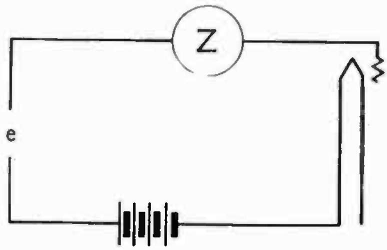

Figure 2

The voltage drop across Z due to this difference frequency current will then vary directly as $\left(\frac{R}{R+Z_0}\right)$ $\left(\frac{R}{R+Z_1}\right)\left(\frac{RZ_{1-0}}{R+Z_{1-0}}\right)$. Now the thing we are trying to find out is how this voltage drop varies with frequency. If it is independent of frequency over the audio range then no frequency distortion results from this method of detection. Let us consider each factor separately. The first factor does not change at all, as the carrier frequency is the same no matter what the audio modulation frequency may be. The second factor is not independent of the beat frequency, but is practically so because the whole audio range may be covered by varying ω_1 about one per cent or less at broadcast frequencies. Hence unless Z contains some element sharply tuned to radio frequency the first two factors are practically constant whatever difference frequency is chosen. Any frequency distortion must then be due to the third factor varying with frequency. An idea that immediately suggests itself is to use for Z some network that has substantially the same impedance to all frequencies in the audio range, and low impedance to radio frequencies. However, before considering this idea further let us find out how much distortion is likely to be introduced by the third factor when Z is composed of the ordinary size grid condenser and leak, as in Fig. 1. From the form of the third factor it is obvious that it represents the impedance of R connected in parallel with Z, and hence of R, r, and C all in parallel. The impedance of this combination varies inversely as

$$\sqrt{1+\left(\frac{rR}{r+R}\;C\,\omega_{1\text{-}c}\right)^2}$$

As an example let us choose $C=0.00025$ mfd. and $r=5$ megohms, and calculate how great $\omega_{1\text{-}c}$ may be without the impedance falling below 90 per cent of the value it has at very low frequencies. If a 201A tube is used with the connections of Fig. 1 and a 5-megohm leak the value of R is about 65,000 ohms. A simple calculation shows that a frequency of about 4,500 cycles must be exceeded before the frequency distortion due to this type of detection exceeds 10 per cent. Hence such distortion is likely to be negligible in comparison with that occasioned by the use of sharply selective radio-frequency circuits. This statement has been verified by experiment, comparing the action of grid condenser and leak detection with that of C battery detection.

It is often stated that the time constant Cr should be smaller than the time of one cycle of the highest audio frequency that is to be received. If that were the case, then the combination assumed above (5 megohms and 0.00025 mfd.) would not allow satisfactory reception of audio frequencies above 800 cycles, which is contrary to experience. However the expression derived above suggests that the time constant of the condenser in parallel with the effective resistance $\frac{rR}{r+R}$ is the proper time constant to consider. Using this value and applying the time constant criterion we predict fairly good detection as high as some 60,000 cycles. This seems very high, but superheterodynes are often built, with similar values of grid condenser and leak for the frequency changing tube, which give a satisfactory output at intermediate frequencies of this order of magnitude.

In the foregoing it has been assumed in all the illustrative examples that the grid was kept positive. If kept negative, R is no longer low compared to r, and the proper time constant to use would be more nearly Cr. It has also been assumed that the source of applied voltage is not affected by tube losses. Actually, if the grid is kept positive, the resonance curve of the tuned circuit which supplies the input voltage will be broadened, so that less frequency distortion is produced by the tuned circuit than if C battery detection were used.

SUMMARY

It has been said that detection by the ordinary grid-condenser and leak gives poor quality (i. e. low audio frequencies favored at the expense of the higher ones) unless an undesirably small condenser or undesirably low leak resistance is used. This paper attempts to show, as simply as possible, that actually, good quality can be obtained even when using as large a grid condenser as desirable, because the resistance which determines the relative loss of high frequencies is not the resistance of the grid leak alone, but the resistance of the grid leak and grid-filament resistance of the tube in parallel. Under usual operating conditions the latter is so low that loss of quality is very slight. The resistance of the grid leak is important chiefly in determining the grid-filament resistance of the tube, and need not be low itself.

THE TORUSOLENOID

An Improved Type of Fieldless Coil Combining the Best Features of the Single Layer Solenoid and the Toroid

By

Ross Gunn, B. S. E. E., M. S., PhD.

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During the last two or three years much good work has been done on the matter of coil design and the prediction of coil performance. This information has to do with the proper design of coil types already existing and to date there appears to have been no fundamental change in the method of winding coils for high frequency. The two general types of coils now available are the solenoids and the toroids, each type offering certain advantages and each having certain inherent disadvantages. The writer has done considerable work with coils and has devised a coil which combines all the good points of each type without incorporating any of the bad features. It is for this reason that the new coils have been called torusolenoids. Because of the novel features of the coils it is believed that information as to the construction and electrical characteristics would be of general interest to those interested in the art.

Before proceeding to the description of the coil it will be found convenient to introduce a factor which the writer has used as a direct measure of coil performance. This factor has been called the "gain" and is, in fact, simply the reciprocal of the power factor of the oscillating circuit or $\frac{2\pi fL}{R}$ where f is the frequency of the oscillations, L is the inductance and R is the equivalent series resistance. Most modern radio circuits are worked in conjunction with vacuum tubes and since vacuum tubes are controlled by potential solely, one is in general interested in the potential applied to the grid of the associated tube in terms of the e.m.f. applied to the circuit. It can easily be shown* that

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the "gain" is simply the ratio of the generated potential (i. e. the potential appearing across the coil, hence the potential applied to the grid) to the applied e.m.f. These potentials are obviously 90 deg. out of phase and their ratio as given above is always greater than unity and corresponds roughly to a circuit potential amplifying factor. Thus if a circuit has a high gain it may be considered very good whereas a low gain corresponds to a poor circuit. This idea may be easily extended to a coil if by the gain of a coil we mean the gain of a circuit made up of the coil and a condenser of negligible resistance. This factor, the gain, as defined above is also equal to the circuit selectivity or sharpness of resonance and it is therefore obvious that for broadcast reception the gain should not be excessive, not over 250 perhaps for the ordinary ear.

It is readily shown that

$$\text{Selectivity} = \text{Gain} = \frac{f_r}{f_2 - f_1}$$

where f_r is the natural frequency of the oscillating circuit when in resonance with the applied frequency, f_2 is the natural frequency of the circuit *above* resonance such that the oscillating current is 70.7 per cent of the current *at* resonance and f_1 is the natural frequency of the tuned circuit at a point *below* resonance such that the oscillating current is 70.7 per cent of the current *at* resonance. Choosing therefore a mean broadcast frequency of 750 kc. and making the assumption that $G = 250$; one finds that the width of the resonance curve between points where the oscillating current (and hence the generated potential) falls to 70.7 per cent of its maximum at resonance is 3,000 cycles. With $G = 250$ modulation frequencies of 1500 cycles will then be amplified in potential but 70.7 per cent that of low modulation frequencies. This difference in amplification would be noticeable to a good ear but there is little doubt that the audio amplifier will partially make up this loss due to audio frequency regeneration which will accentuate the high frequencies and the arbitrary selection of 250 for the maximum value of G seems satisfactory for present day needs at least.

Returning now to the consideration of the coil, let us write down the essentials of the supreme coil. Perhaps not

*Some Facts About Coil Design, R. Gunn, Radio Broadcast, May, 1927.

all of these can be obtained immediately but the coil to be described seems to fulfill the requirements to a surprising degree. The essentials of the ideal coil are somewhat as follows:

(1) The gain factor and hence the selectivity of the circuit as defined in the preceding paragraphs should be a maximum but should not exceed a mean value of 250 if the coil is to be used for broadcast reception and faithful reproduction desired.

(2) The external field of the coil should be zero or certainly nearly so in order to reduce to a minimum stray energy exchanges within the radio set.

(3) The distributed capacity of the coil should be very low in order that the tuning range should be a maximum and in order that the dielectric losses within the coil may be a minimum. A nearly uniform potential gradient should exist throughout the coil and the terminals should be well separated so that connecting wires will not introduce excessive capacity.

(4) The physical structure of the coils should not be excessively large taking into account the fact that a small coil necessarily means a low value for the gain under similar types of construction.

(5) Mechanically it should be strong and capable of withstanding a certain amount of abuse.

The first essential contains much implied meaning, since for the gain and the selectivity to be large, great care must be exercised in the selection of materials and so placing them as to produce no unnecessary losses. It would appear that essentials (2) and (3) are entirely incompatible and it is on this point that the great advantage of the new coils depends for it has been found possible to satisfy both conditions simultaneously by a special method of winding the coils.

The single layer solenoid which is the coil in general use today has a relatively large value for the gain when properly constructed. It has a sufficiently low distributed capacity and satisfies all requirements well enough except that relating to its external magnetic field. On this point the simple solenoid is very poor for it is well known that if a solenoid is not carefully placed at some distance from other parts of the radio set its exterior field will play havoc by causing undesired interaction. The external field is a double disadvantage in that the coil will easily affect other parts of the circuit and conversely other parts of the circuit easily affect it. In addition to this difficulty the field will set up eddy currents in any metal parts and dielectric losses in any insulator which will react unfavorably on the gain factor.

The toroid on the other hand has a very small exterior field but the selectivity and gain are very low. The distributed capacity is relatively large so that on the whole one is worse off by using a toroid in place of the single layer solenoid.

The points in favor and against each type of coil may be summarized as follows. The good features have been marked *plus* and the bad features *minus.*

Single Layer Solenoid.

+(a) High value for the gain and selectivity.

—(b) Large exterior field.

+(c) Moderate values for the distributed capacity.

Simple Toroid.

—(a) Very low to moderate values for the gain and selectivity.

+(b) Very small exterior field.

—(c) Large distributed capacity.

The points not mentioned are on the whole satisfactory for both types. It will be observed that one type of coil has one set of advantages and the other coil has the complementary ones, so if one could combine the advantages of each into one coil the resultant coil should be most satisfactory. The new Torusolenoids do just this; whence the name.

During the preliminary work on this subject the author saw that there was little chance of modifying the solenoid so that its exterior field would be small but there did seem to be lots of room for improvement in the toroid! The toroid was chosen as the coil that offered the best all around possibilities and much work was done in an attempt to improve it in its simple form but with scant success.

Method of Construction

The author's coil in external appearance is not unlike the usual toroid except it will be observed that the terminals of the new coil are on opposite ends of a diameter while in the conventional type the coil ends are adjacent. By the use of the new method the potential gradient is made nearly uniform along the coil and the high and low potential ends are well separated. This fact helps explain how the distributed capacities are kept to low values. It will be observed on closer examination that the coil is wound in two sections, the sections are wound in opposite directions, that is to say one half of the coil is a right hand spiral and

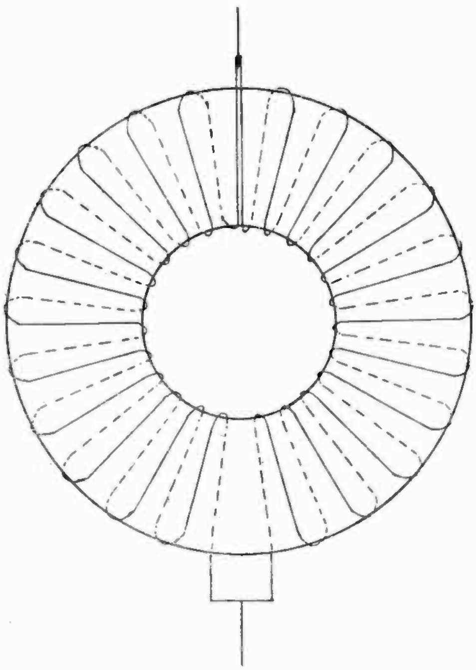

Fig. 1—Torusolenoid—The fieldless low-capacity high-gain coil.

the other half is a left hand spiral. This arrangement means that the two wires leading to either terminal *start* around the coil in the *same* direction. This particular method of winding secures a uniform magnetic field intensity within the coil so that there is no tendency whatever for the magnetic flux to stray out of the coil and cause interference with other parts of the set. The diagram of Fig. 1 shows the general method of winding much better than it is possible to describe it in words. It will be observed that each section or half is in itself a complete coil and the purpose of its associated coil is simply to provide a return path for the magnetic flux of the first coil. Thus each coil is mutually dependent on its mate for its efficiency. Now since the two coils are connected in parallel it is obvious that the inductance of each half must be twice the resultant inductance so that in the calculation of the number of turns necessary for a given inductance a factor of four must be introduced into the usual equation. This brings us to a third fundamental difference between the new coils and the old type. The high value of inductance for each *half* of the coil allows the use of a large number of turns of

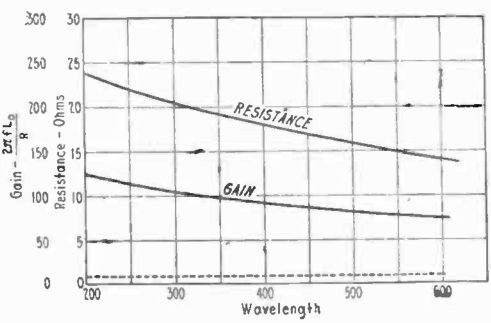

Fig. 2—Commercial "low loss" toroid
110 Turns, No. 20 B. & S.
$L_0 = 330\mu h. — \lambda_0 = 93.4m. — C_0 = 7.5\mu\mu f.$

relatively small wire. This is of considerable value in increasing the gain when taking into account other factors such as the fact that there are effectively two coils in parallel, for it is well known that at high frequencies the smaller wires use the copper more effectively. The use of small wire aids greatly in cutting down eddy current losses and the distributed capacity of the coil. One other feature should be mentioned and that is to point out that while the simple toroid has an exterior field corresponding to a single turn of wire whose mean area is that of the coil, the new torusolenoid has practically no field since the fields in each half tend to cancel each other outside the coil proper.

Some of the new coils have been mounted on pyrex rings and others on hard rubber forms upon which the coil and binding posts could be mounted. This method reduces to a minimum all losses since the coils are almost self supporting and the individual turns are properly spaced. The size of the coils have been limited to 4¼ inches since it was recognized that coils much larger than this would be unsuitable for ordinary use. By increasing the size of the coil the gain may be made much larger than the coils described in this paper and conversely if a coil is made smaller, assuming the same general type of construction the gain factor will necessarily be decreased.

The inductance may be computed for this type of coil by aid of the following relation which is similar to the usual equation save for a factor of four.

$$L_0 = 0.00316\, N^2 \left[R - \sqrt{R^2 - A^2} \right]$$

where L_0 is the inductance in microhenrys, R the distance from the axis of symmetry to the center of cross section of the winding, A is the radius of the turns of the winding and N is the total number of turns of the winding. All dimensions are to be inserted in centimeters. There is no known way of computing the other constants of the coil and these must be determined by experiment.

Results

Turning our attention to the actual results of tests on the improved coils we shall discuss in order the important characteristics of the coils as follows:

(a) The gain and selectivity.
(b) The exterior field.
(c) The distributed capacity.

As a basis of comparison the writer purchased a so-called low loss toroid and obtained all necessary data to rate the coil properly. It was wound with 110 turns of No. 20 B.S. wire, was self supported and had no dielectric whatever except the cotton insulation. Its direct current resistance was but 0.9. ohm while its high frequency resistance at 300 meters was 20.5. ohms. The curves of Fig. 2 show the resistance and gain for wavelengths within the broadcast band. The fundamental wavelength of the coil was 93.4 meters and its distributed capacity was 7.5 micro-microfarads.

GAIN

It will be remembered that the gain is a function of the circuit constants only and is equal to

$$G = \frac{2\pi f L_o}{R} \quad \text{or} \quad G = \frac{1}{R}\sqrt{\frac{L}{C}}$$

where the symbols have the usual meaning. This factor may be determined directly by means of a vacuum-tube voltmeter or it may be computed from known data on the coil. The value of the gain will obviously depend on the coil and its numerical value will range from perhaps 60 for a poor coil to 300 for a high-class coil. A value much over 250 is undesirable for broadcast reception since the quality of reproduction would be impaired due to the suppression

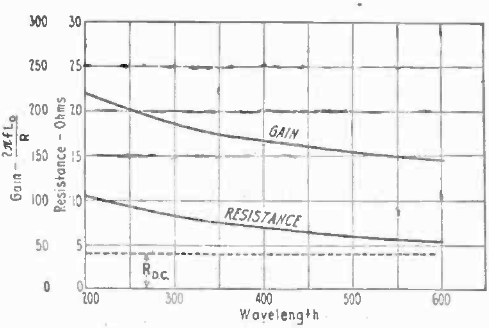

Fig. 3—Torusolenoid

480 Turns, No. 30 D. S. C.
$L_o = 242$ μh.
$\lambda_o = 63$m.
$C_o = 4.6$ $\mu\mu$f.

of the side bands. The new torusolenoids have values for the gain from 135 to 250 for solid wire-wound coils and as high as 300 for litzendraht coils of moderate size. The curves of Fig. 3 show how the gain and effective series resistance of a typical solid wire-wound torusolenoid changes with wavelength. It is interesting to note that the average high frequency resistance of the torusolenoids is only about twice its direct current resistance whereas the sample of the simple toroid has a high frequency resistance about 18 times the d.-c. resistance. The data for a litzendraht coil which is typical of the class are shown in Fig. 4 and are plotted in solid lines. It will be seen that the gain over the broadcast band averages well over 200 and rises to well over 250 at 550 meters. The dotted curves of Fig. 4 show plainly the effect of enclosing the coil in a metal can and is remarkable in that the gain is decreased by only about 18 per cent and rather definitely showing that the exterior magnetic field is zero.

An examination of the curves, which have all been plotted to the same scale, shows that the solid wire-wound torusolenoids have the highest gain at the short wavelengths and decrease as the wavelength is increased. On the other hand the litzendraht coils show a relatively lower value at the short wavelengths and increase uniformly with increasing wavelength. This type of characteristic seems to offer

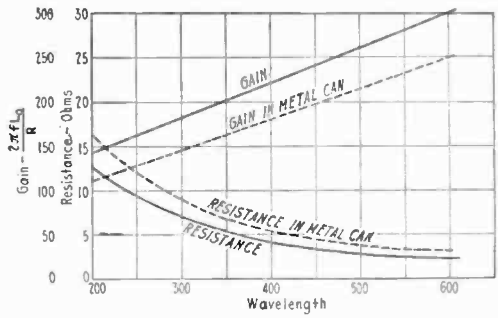

Fig. 4—Torusolenoid
300 Turns — "Litz."
In Air $L_0 = 194 \mu h$.
Solid Curves $\left\} \begin{array}{l} \lambda_0 = 52m. \\ C_0 = 4 \mu\mu f. \end{array} \right.$
In Metal Can $L_0 = 194 \mu h$.
Dotted Curves $\left\} \begin{array}{l} \lambda_0 = 81.3m. \\ C_0 = 9.6. \end{array} \right.$

certain advantages since in general the energy transferred to a coil is greater at high frequencies with the usual methods of coupling, so a lower value for the gain can be tolerated. By proper design it would appear that uniform amplification could be obtained over the entire broadcast range since the gain seems to be nearly a linear function of the wavelength and the induced e.m.f. a linear function of the frequency. More work is being done on this matter in order to verify this prediction since a straightforward electrical solution of this problem is more desirable than mechanical means of varying the coupling.

The peculiar characteristic just described together with gain factors at least equal to the simple solenoid seems to indicate that the new coils are at least on the par with any other type when considering their gain.

External Field

The external field of the new coil is even less than that of the ordinary toroid since the ordinary toroid has a field associated with it which is equivalent to a single turn whose area is equal to the mean area of the coil. In the torusolenoid the current divides and the external field of each half tends to neutralize the other.

In order to test experimentally whether the torusolenoid had an external field, a test coil was properly connected to a small fixed condenser and a sensitive thermogalvanometer. This combination was tuned to resonance with a powerful oscillator. The coil was then placed directly in the field of an oscillator plate coil which was carrying 10 amperes of a frequency corresponding to the tuned circuit of the coil. The coil under test was oriented in all directions but no current could be detected in the coil circuit. This test appears to show that the external field of the torusolenoid may be said to be effectively zero and hence stray energy exchanges due to magnetic effects are nonexistent.

Distributed Capacity

It has been recognized for a long time that one of the essentials of a good coil, regardless of its type, was a low distributed capacity. A low distributed capacity guarantees that the electric gradient within the coil has been kept small and the size and distribution of the wire used in winding has been well chosen. A low distributed capacity is particularly valuable these days when many broadcast stations are at the extremes of the band and the band none too narrow. The torusolenoids have shown an exceptionally low value for the distributed capacity, so low in fact that the usual method of plotting λ^2 against the capacity and extrapolating could not be used. The procedure used was to determine the fundamental wavelength of the coil by a simple means and then compute the distributed capacity from the relation

$$C_d = \frac{\lambda_o^2}{3.55\,L}$$

where C_d is the distributed capacity in micro-micro-farads, λ_o is the fundamental wavelength of the coil and L is the inductance in microhenrys.

The average torusolenoid will show distributed capacities of from 4 to 5 micro-micro-farads when properly constructed. The data on the commercial coil are included in Fig. 2 for comparison.

There are many variations which may make use of the new type winding. One method which has been found fairly good is to make use of two slender solenoids which were wound in opposite directions and connected in parallel. This type of mounting yields a coil whose characteristics are

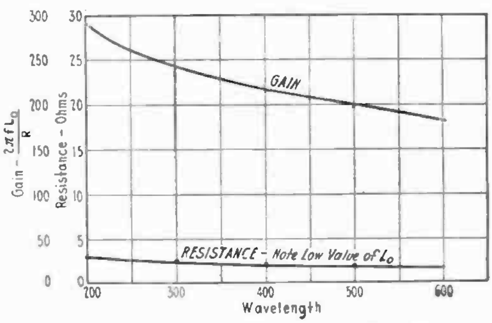

Fig. 5—Binocular—New Type Winding
No. 26 D.C.C. on 2 inch form—4 inches long.
$L_o = 93 \mu h$.

shown in Fig. 5. It will be observed that the gain may be made quite high but here again there is an appreciable exterior field and a choice made as to what characteristics are most desirable.

Consideration of all the factors involved in these coils and their properties show that the torusolenoid fulfills to a remarkable degree the requirements of the ideal coil. The features possessed by the torusolenoids may be summarized as follows:

(a) They have a high value for the gain and selectivity.

(b) Their external magnetic field is vanishingly small.

(c) The distributed capacity is remarkably low.

(d) They are moderate in size taking into account item (a).

(e) They are mechanically strong enough to withstand a reasonable amount of abuse.

SUMMARY

An improved high frequency inductance of radical design is described which represents the successful attempt to combine into one inductance all the desirable features of both the toroid and the single layer solenoid.

The new inductance which has been called the "Torusolenoid" has been found to incorporate the following features:

1. It has substantially zero external magnetic field.
2. It has a high value for the "gain".
3. It has very low distributed capacity.
4. It has no major disadvantages when used as an inductance.

The term "gain" is defined and shown to be a measure of coil performance. Supporting data on representative inductances incorporated in the paper show that the new inductances offer many advantages, and for all-around performance are unsurpassed.

BOOK REVIEW

Robison's Manual of Radio Telegraphy and Telephony

BY

COMMANDER (now REAR ADMIRAL) S. S. ROBISON,
U. S. Navy.

Revised by COMMANDER S. C. HOOPER, U. S. Navy and
LIEUTENANT COMMANDER T. A. M. CRAVEN, U. S. Navy.

Seventh Revised Edition; 737 pp.; 424 fig.; The United States Naval Institute, Annapolis, Md.; Price $5.50.

This book is divided into four sections headed in order, (1) theory of radio communication, 462 pp.; (2) application, 125 pp.; (3) radio measurements and precision instruments, 92 pp.; (4) useful information, 46 pp.

Section 1 dealing with the theory of radio communication includes the elementary theory of electricity, alternating current theory, application of alternating current theory to radio circuits, theory of damped oscillations, theory of vacuum tubes, theory of wave propagation, and radio accessories such as inductance coils, condensers and quartz crystals. A chapter on quartz crystals and one on the Taylor-Hulburt theory of transmission appear as additions in this revision. This section presents the principles underlying radio communication in a clear and concise manner.

Section II, presenting the applications, includes radio transmission, radio reception, and radio compass. This section deals with descriptions of apparatus for transmission, reception, and radio compass. The work on spark and arc transmitters is abbreviated in this edition. The work on receivers is largely rewritten.

Section III on radio measurements and precision instruments consists of twenty-six experiments of practical and precise measurements. These are suitable for teachers in need of such material as well as for laboratory workers. This section is almost identical with the corresponding work in the 6th edition.

Section IV, useful information, includes tables and formulas which are needed by the radio student. In this edition this section has been shortened by the omission of the work on elementary mathematics, and on laws and regulations.

Taking the book as a whole it is very readable. The explanations are remarkably clear and concise. Figures are good, well placed, and used to advantage in connection with the text. The many worked examples are helpful. Mathematics is not avoided but is used in such a way that the inexperienced reader can understand it easily. Important statements are emphasized by bold-faced type. The binding is sturdy. The Manual may be highly recommended not only to the elementary student to whom it is addressed chiefly, but also to the more mature worker and teacher.

The following errors in the statements of formulas and mathematical work should be noted:— p. 35 $R = lSP$ should be $R = \frac{lP}{S}$; p. 36 $g = \frac{\gamma}{lS}$ should be $g = \frac{\gamma S}{l}$, and $g = \frac{1}{lSP}$ should be $g = \frac{S}{lP}$; p. 93 $blB_g \cos \omega (t + dt) = blB_g \cos \omega t \sin \omega dt - \sin \omega t \cos \omega dt$ should be $blB_g \cos \omega (t + dt) = blB_g (\cos \omega t \cos \omega dt - \sin \omega t \sin \omega dt)$; p. 167 line beginning "and substituting", 10^2 should be 10^4 p. 176 $I = .701\ I_0$ should be $I = .707\ I_0$; p. 225 $\frac{R}{2L} < \frac{1}{LC}$

should be $\frac{R}{2L} < \frac{1}{\sqrt{LC}}$; p. 642 $\lambda_1 = 1885 \sqrt{(L_1 - L)C}$ should be $\lambda_1 = 1885 \sqrt{(L_1 + L)C}$; and $\lambda_2 = 1885 \sqrt{(L_2 - L)C}$ should be $\lambda_2 = 1885 \sqrt{(L_2 + L)C}$.

S. S. Kirby.

DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued August 2, 1927—August 23, 1927

By

JOHN B. BRADY

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

1,637,445—A. W. FRANKLIN, New York, N. Y. VARIABLE MICA CONDENSER. Filed Oct. 11, 1923, issued Aug. 2, 1927. Assigned to Chas. Freshman Co., Inc.

1,637,613—F. A. KOLSTER, Palo Alto, Calif. RADIO COMPASS. Filed Nov. 22, 1920, issued Aug. 2, 1927. Assigned to Federal Telegraph Co.

1,637,646—H. KIMURA, Los Angeles, Calif. Filed Jan. 7, 1925, issued Aug. 2, 1927.

1,637,864—F. A. KOLSTER, Palo Alto, Calif. ELECTRON TUBE APPARATUS. Filed May 25, 1926, issued Aug. 2, 1927. Assigned to Federal Telegraph Co.

1,637,615—F. A. KOLSTER, Palo Alto, Calif. RADIO COMPASS. Filed Nov. 22, 1920, issued Aug. 2, 1927. Assigned to Federal Telegraph Co.

1,638,158—ELECTRIC APPARATUS. P. C. HEWITT, Ringwood Manor, N. J. Filed July 30, 1920, issued Aug. 9, 1927. Assigned to Cooper Hewitt Electric Co.

1,638,278—CONTACT RECTIFYING DEVICE. W. O. SNELLING, Allentown, Pa. Filed Feb. 8, 1923, issued Aug. 9, 1927.

1,638,417—SHIP'S SIGNALING OR BROADCASTING DEVICE. E. A. SPERRY, Brooklyn, N. Y. Filed Nov. 6, 1923, issued Aug. 9, 1927. Assigned to Sperry Gyroscope Company.

1,638,608—WAVE FILTER RECEIVING CIRCUIT. PETER WELCH and AUDLEY MURPHY, of Olean, N. Y. Filed Feb. 6, 1925, issued Aug. 9, 1927.

1,638,320—SOUND REPRODUCING APPARATUS. CYRIL A. BRIGHAM, of East Orange, N. J. Filed Nov. 2, 1926, issued Aug. 9, 1927. Assigned to Brandes Laboratories, Inc.

1,638,499—ELECTRON DISCHARGE DEVICE. A. MAVROGENIS, Milwaukee, Wis. Filed Jan. 2, 1926, issued Aug. 9, 1927.

1,638,699—A. L. MOBURG, Palms, Calif. UNDERGROUND ANTENNA. Filed Mar. 30, 1926, issued Aug. 9, 1927.

1,638,551—ELECTRON DISCHARGE DEVICE. V. L. RONCI, Brooklyn, N. Y. Filed July 30, 1924, issued Aug. 9, 1927. Assigned to Western Electric Co. Inc.

1,638,925—HIGH FREQUENCY SIGNALING SYSTEM. L. ESPENSCHIED, Hollis, N. Y. Filed Aug. 16, 1922, issued Aug. 16, 1927. Assigned to American Telephone and Telegraph Co.

1,638,930—VOLTAGE REGULATOR FOR VACUUM TUBES. J. HERMAN, New York City. Filed Nov. 26, 1923, issued Aug. 16, 1927. Assigned to American Telephone and Telegraph Co.

1,639,414—RADIO RECEIVING SYSTEM. J. M. MILLER, Philadelphia, Pa. Filed June 23, 1926, issued Aug. 16, 1927. Assigned to Atwater Kent Mfg. Co.

1,639,597—CONDENSER. WILLIAM DUBILIER, New York, N. Y. Filed May 23, 1922, issued Aug. 16, 1927. Assigned to Dubilier Condenser Corp.

1,639,650—HIGH POTENTIAL CONDENSER. WILLIAM DUBILIER, of New York, N. Y. Filed March 7, 1919, issued Aug. 23, 1927. Assigned to Dubilier Condenser Corp.

1,639,667—METHOD FOR RADIO POSITION FINDING. R. H. RANGER, of Brooklyn, N. Y. Filed March 8, 1924, issued Aug. 23, 1927. Assigned to Radio Corp. of America.

1,639,686—WIRELESS TELEGRAPH STATION. H. J. J. M. DeBELLESCIZE, of Toulon, France. Filed August 29, 1921, issued Aug. 23, 1927.

1,639,695—ARRANGEMENT FOR THE REGULATION OF THE FREQUENCY IN FREQUENCY CHANGERS. K. HEEGNER, of Berlin, Germany. Filed Oct. 23, 1925, issued Aug. 23, 1927. Assigned to Gesellschaft Fur Drahtlose Telegraphie M. B. H.

1,639,696—MEANS FOR CHANGING WAVELENGTHS. G. HILL, of Brooklyn, N. Y. and GEORGE H. CLARK, of Washington, D. C. Filed May 3, 1915, issued Aug. 23, 1927. Assigned to Radio Corporation of America.

1,639,698—ELECTRON EMITTING CATHODE AND PROCESS OF PREPARING SAME. F. HOLBORN, of Hoboken, N. J. Filed Jan. 14, 1926, issued Aug. 23, 1927. Assigned to Hazeltine Corporation.

1,639,727—AERIAL. M. E. FRASER and LOUIS SCHULTZ, of Detroit, Mich. Filed Feb. 23, 1926, issued Aug. 23, 1927.

1,639,773—TWO-WAY TELEPHONE TRANSMISSION. H. S. HAMILTON, of New York, N. Y. Filed Nov. 24, 1926, issued Aug. 23, 1927. Assigned to American Telephone and Telegraph Co.

1,639,805—RADIO APPARATUS. F. S. McCULLOUGH, of Cleveland, Ohio. Filed Jan. 9, 1920, issued Aug. 23, 1927. Assigned to Glenn L. Martin.

1,639,816—RADIO SIGNALING CIRCUIT. A. H. TAYLOR and LEO C. YOUNG, of Washington, D. C. Filed Jan. 7, 1925, issued Aug. 23, 1927. Assigned to Wired Radio, Inc.

1,639,817—PIEZO ELECTRIC CRYSTAL SYSTEM. A. H. TAYLOR, of Washington, D. C. Filed July 29, 1925, issued Aug. 23, 1927. Assigned to Wired Radio, Inc.

1,639,913—ANTISTATIC AERIAL. R. A. WEAGANT, of New York, N. Y. Filed Oct. 6, 1920, issued Aug. 23, 1927. Assigned to Radio Corporation of America.

1,640,200—MOUNTING MEANS FOR RADIO RECEIVING SETS. F. L. LORD, of Newark, N. J. Filed Aug. 28, 1925, issued Aug. 23, 1927. Assigned to The Lord Laboratories, Inc.

1,640,234—ANTENNA FOR USE WITH RADIO SETS. M. M. CONYERS, of Custer, South Dakota. Filed June 14, 1926, issued Aug. 23, 1927.

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	Catalina Island, Avalon, Station KFWO	Paggi, M. M.
	Long Beach, 4300 East 5th St.	Randall, A. G.
	Los Angeles, 5356 5th Ave.	Kuchenbaur, Edw.
	Los Angeles, 621 So. Hope St.	Nolan, A. B.
	Manhattan Beach, Box 74-R1	Tischer, Frank A.
	Oxnard, P. O. Box 523	Newcomb, C. A.
	Piedmont, 1661 Grand Ave.	Myers, W. M.
	Sacramento, 1340 33rd St.	Koch, Wm. F.
	Sacramento, 2181 Weller Way S.	Semoni, A. J.
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	San Francisco, Care U. S. Engineer Dept. Dist. 1.	Senior, David B.
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Dist. Columbia,	Washington, 3401 16th St., N. W.	Israel, J. O.
	Washington, Navy Dept., Rm 2624	McCandlish, B. V.
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	Indianapolis, 1863 Dexter St.	Chrisney, R. G.
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	Boston, 528 Beacon St.	Olmstead, N. C.
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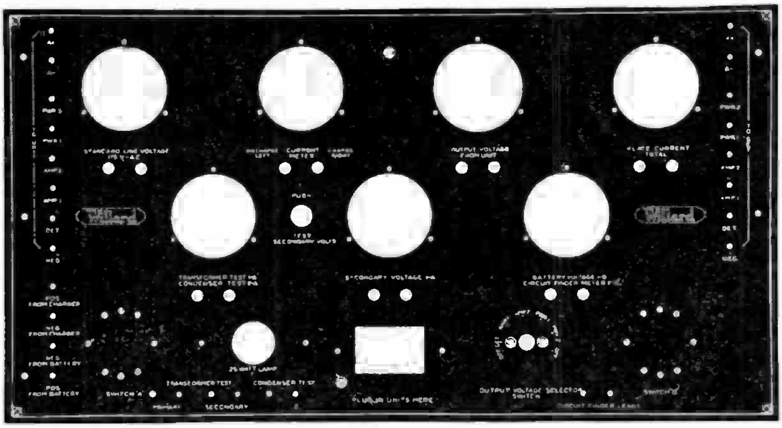

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