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ALFRED N. GOLDSMITH, Ph.D.

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A METHOD OF MEASURING RADIO FIELD INTENSITIES AND ATMOSPHERIC DISTURBANCES*

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(Communication from the International Union of Scientific Radio Telegraphy)

THE TELEPHONE COMPARATOR

The following method of measuring telephone currents has been developed to take the place of the older telephone shunt method.

The apparatus is shown in Figures 1 and 2. Here 1 represents the 1,000-cycle tuning fork generator, manufactured by the General Radio Company, Cambridge, Massachusetts, Type 213. This generator operates on 6 volts and delivers a current of about 15 milliamperes thru 100 ohms. This has been found considerably superior to any audio-frequency electron tube generator which has been constructed, on account of relative freedom from harmonics, and ease in suppressing residual effects. To prevent hearing the fork directly, the oscillator is generally placed outside the receiving room.

Figure 2 represents the circuits making up the telephone comparator. The voltage divider has a resistance of 100 ohms, with two dials having steps of one and ten ohms. The milliammeter used is of the thermo-element type, having a range of 0 to 25 milliamperes, the current used being generally 10 milliamperes, giving a fall of potential across the voltage divider of one volt. The current to the telephones passes thru resistances R_1 and R_2, which are large enough to make variations in the telephone impedance

nearly negligible. Generally 200,000 ohms are used. In this way it is easy to calculate the current in the telephones without accurate knowledge of their impedance. The telephone terminals are connected to a double-pole, double-throw switch which permits them to be connected either to the comparator, or to the radio receiving circuit.
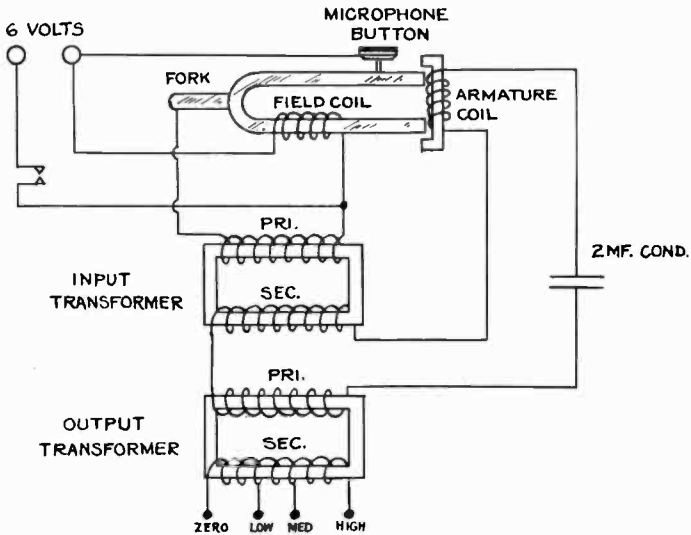
FIGURE 1—General Radio Company Type 213 1000 ∽ Audio Oscillator
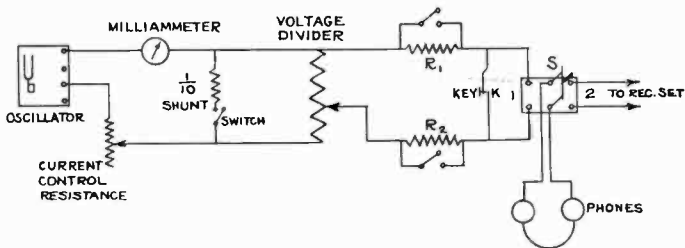
FIGURE 2—Telephone Current Comparator

MEASUREMENT OF SIGNALS

The process of measuring signals in the case of beat reception is as follows: the telephones are connected to the receiving set, and the radio frequency circuits are tuned to give beat reception, and the heterodyne or autodyne condenser adjusted to give the same beat note as the audio oscillator. If the signals are at all strong, the telephones are shifted forward on the head until the

received signals are just audible since the comparison of sounds is much more accurate if they are weak. The telephones are then thrown to the telephone comparator side and the voltage divider adjusted until the telephone sound is of the same intensity in the two positions of the switch, the comparator currents being broken up to simulate signals by means of the key K. In this way the telephone currents can be measured with an accuracy of from 3 to 5 percent.

For the measurement of spark signals and atmospheric disturbances, no matching of the note of the comparator is possible, which, of course, materially reduces the accuracy of the measurements.

For the measurement of very weak signals it is desirable to reduce the current thru the voltage divider without correspondingly reducing the current in the milliammeter. This may be done by placing a one-tenth shunt across the voltage divider at S.

Electric Field Intensity

In order to make absolute measurements of the field from the transmitting stations, it is necessary to know (1) the resistance of the receiving antenna at the given wave length, (2) the radiation height (effective height) of the receiving antenna, and (3) to calibrate the telephone current as measured on the telephone comparator in terms of antenna current.

(1) The resistance (R) of the antenna is best obtained by the resistance-variation method, using a tube oscillator as the source of power and measuring the current in the antenna with a thermo-milliammeter or thermo-element and galvanometer (see Bureau of Standards, Circular Number 74).

(2) The radiation height (h_r) of the receiving antenna can be measured by measuring the current received from a station one to ten wave lengths away, of known transmitting current (I_s), radiation height (h_s) and distance (d), using thermo-element and galvanometer, from the formula, expressed in wave length λ and also in frequency f,

$$I_r = 120\,\pi\,\frac{I_s\,h_r\,h_s}{\lambda\,d\,R} = 40 \times 10^{-5}\,\pi\,\frac{I_s\,h_r\,h_s\,f}{d\,R}$$

in meters, ohms, kilocycles, amperes,

$$h_r = \frac{I_r\,\lambda\,d\,R}{120\,\pi\,I_s\,h_s} = \frac{I_r\,d\,R}{40 \times 10^{-5}\,\pi\,I_s\,h_s\,f}$$

If this measurement cannot be made, the approximate height h_r may be taken as $\frac{1}{2}$ the average height of the flat top (equivalent

to the proposal by Dellinger.[1] If a coil antenna (loop) is used instead of an elevated antenna, its effective height can be calculated, disregarding the effect due to capacity to earth (antenna effect) from the formula.

$$h_r = 2\,\pi \frac{area \;\times\; number\;af\;turns}{\lambda} = 6.67 \times 10^{-6} f \cdot area \times turns$$

(3) The telephone current as measured in the telephone comparator can be calibrated in terms of antenna current from a signal of known intensity or better from a local oscillator as described later. The constancy of this calibration over long periods is very satisfactory, provided the receiving circuits are tuned as follows: Both antenna and secondary circuits are first tuned with very loose coupling so that there is no reaction between them. The coupling is then closed until the best signal is obtained, returning the secondary as may be required, but leaving the *antenna tuning unchanged*. This method, while not giving the loudest possible signals, can be depended on to give constant and perfectly reproducible results. The detuning of the secondary at least at the longer wave lengths in autodyne reception prevents the introduction of appreciable coupling resistance in the antenna.

Experiment shows that the telephone current, I_t, in beat reception is proportional to the square root of the watts, W, in the receiving antenna, and if E be the total emf. on the antenna, h the radiation height, e the electric field intensity, and R the resistance

$$W = \frac{E^2}{R} = \frac{h^2 e^2}{R} \quad \text{and}$$
$$e = \frac{\sqrt{WR}}{h}$$

since $\sqrt{W} = B\,I_t$ where B is the sensitivity constant of the apparatus.

The field intensity $e = \dfrac{B\,I_t \sqrt{R}}{h}$

where B is to be determined from observations on signals of known strength or from the local oscillator.

If it is desired to construct the apparatus for the measurement of signals, the following may be of service:

A suitable two-dial voltage divider can be made from any

[1]"Principles of Radio Transmission and Reception with Antenna and Coil Aerials," J. H. Dellinger, Bureau of Standards Scientific Paper, Number 354; 1919.

dial resistance box by reconnecting the units and tens as shown in Figure 3 and cutting out the other dials or keeping them at zero.
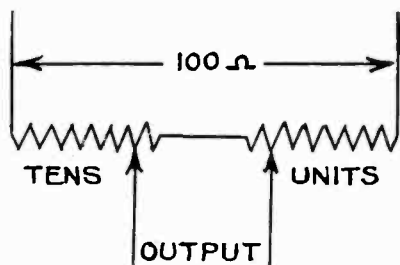
FIGURE 3—Connection for Voltage Divider

The thermo-milliammeter in the comparator may be replaced with some considerable reduction in cost by a Western Electric Company's vacuum thermo-element and General Radio Company's 10-ohm pointer galvanometer. A 35-ohm thermo-element shunted by six or seven ohms is convenient as this can be used without shunt in the local oscillator to be described later. The thermo-element galvanometer combination is calibrated by direct current which must be reversed thru the thermo-element heater and the mean deflection taken if a difference in reading is shown on reversal. Another possible inexpensive combination for measuring the comparator current is a Weston thermo-galvanometer (115 milliamperes full scale) and proper current transformer, tho the deflection will be small.

Two Lavite units of 100,000 ohms each are convenient for the resistances in the telephone leads.

A simple and compact tube oscillator can be constructed, using a low-current tube (UV-199) with dry cell "A" battery and "B" battery and the circuit shown in Figure 4. It is important that the whole oscillator be enclosed in a copper-lined box with tight-fitting cover. The copper box is grounded and also the lower end of the 1-ohm shunt, preferably a straight piece of Number 36 "Advance" wire. It is also important that the coil L_4 (Figure 5), to which the secondary of the receiving set is coupled, should be next the ground, for even a small amount of capacity coupling will produce large errors in the measurements. If a closed coil antenna (loop) is used instead of an elevated antenna the 1-ohm shunt from the oscillator is best placed in the middle of the loop. The thermo-element used in our laboratory is a Western Electric Company's type of 35 ohms heater resistance and the galvanometer is a General Radio Company's 10-ohm pointer instru-
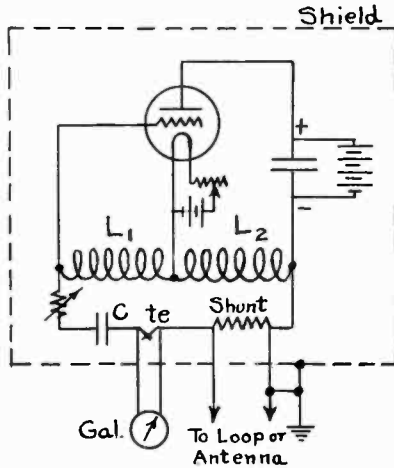
FIGURE 4—Shielded Oscillator

ment. The galvanometer can in most cases be kept outside the shielded box. As calibration on one wave length in the general range of observation is all that is required, if the oscillator is not to be used for other purposes, the condenser C for the sake of compactness can be of the fixed type. For work with trans-Atlantic stations, $L_1 = 25$ mh., $L_2 = 5$ mh., and $c = 0.002$ μf. This will give a wave length of about 15,000 m. The radio frequency current in the thermo-element should be between one-half and three milliamperes. This may be regulated by the coupling between the coils and by the filament current, or by resistance.

FIGURE 5

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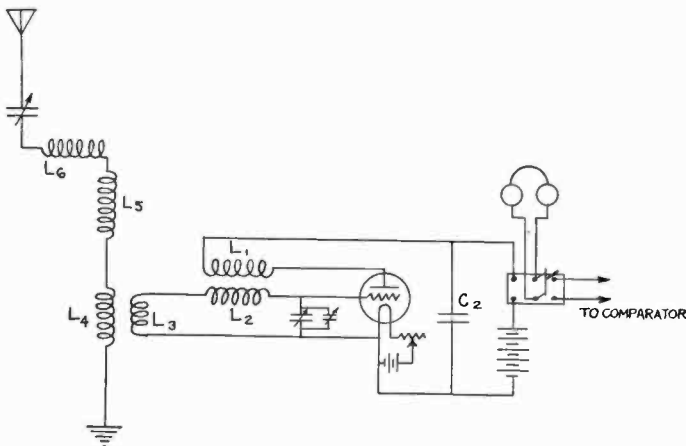
Example of Calibration

Suppose a current of one milliampere is sent thru the 1-ohm shunt of the tube oscillator. This impresses 1 millivolt on the antenna, so if the antenna resistance including coils is 75 ohms, the power in the antenna

$$W = \frac{E^2}{R} = \frac{1 \times 10^{-6}}{75} = 1.33 \times 10^{-8} \; watts$$

$$\sqrt{W} = 1.15 \times 10^{-4}$$

Experiments in the laboratory show that the strength of the signal current in the telephones with this connection, for a certain potentiometer setting, is, for our apparatus, approximately five times the strength of the current with the comparator used normally with 200,000 ohms in series with the telephones.

Now if at the same time the telephone comparator with its voltage divider set at 13 ohms gives the same intensity of sound in the telephones as the tube oscillator with the secondary properly coupled to the antenna, there will be a voltage on the telephone circuit from the comparator of 0.13 volts. If the telephones have an impedance of approximately 30,000 ohms and the added resistance in the telephone circuit is 200,000 ohms, the telephone current

$$I_t = \frac{1.30 \times 10^{-1}}{2.3 \times 10^5} = 5.65 \times 10^{-7} \; amp.$$

Since the telephone current is proportional to the square root of the antenna watts in beat reception

$$\sqrt{W} = B \, I_t$$

where B is the sensitivity constant of the receiving set and

$$B = \frac{\sqrt{W}}{I_t} = \frac{1.15 \times 10^{-4}}{5.65 \times 10^{-7}} = 204$$

The Receiving Set

The autodyne receiving set used in the measurements is shown in Figure 5. The inductances L_1, L_2, L_4, L_5, and L_6 are coils of 30 mh., inductance and L_3 is 5 mh. The 30 mh. coils are wound with litz wire (48 strands, Number 38) on flat wooden spools. The wire space is 21 mm. long with an inside diameter of 72 mm. and outside diameter of 150 mm. The 5 mh. coils are similar but with an outside diameter of winding of about 100 mm. These have a resistance at the trans-Atlantic wave lengths of a little less than one ohm per millihenry. While it is desirable to have the coils in the antenna of low resistance, this is of no con-

sequence in the secondary circuit, and solid wire can be used if desired, since resistance can be compensated by increasing the back coupling. Flat type coils are convenient for long-wave work but are not suitable below 2,000 meters and may be replaced thruout the set by honeycomb or other good coils which may be used at all wave lengths. The coil L_3 is attached to flexible leads for convenience in varying the coupling to the antenna coil L_4, which should be nearest the ground. The secondary variable condenser has a maximum capacity of 0.003 μf, in parallel with a 0.0005 μf. condenser used for exact tuning. The variable antenna condenser has a capacity of 0.005 μf., while the capacity of the fixed stopping condenser C_2 is 0.002 μf. To guard against variations in tube sensitivity with use, two tubes of the same sensitivity are chosen and connected so that either one can be thrown into the set. In this way a check can be kept on the constancy of the regular observing tube by occasional comparison with the spare.
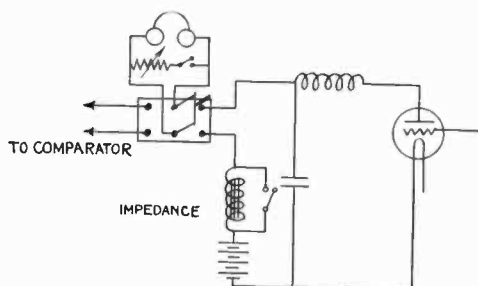
FIGURE 6—Circuit for Measuring Strong Signals

SPECIAL METHODS OF USING THE TELEPHONE COMPARATOR

1. MEASUREMENT OF VERY STRONG SIGNALS

Some difficulty has been experienced in measuring the strength of strong American stations with the telephone comparator, because it is generally necessary to keep the telephones at a considerable distance from the ears, in order to obtain a perfect comparison. This condition often allows outside noises to distract the observer's attention.

The following arrangement permits keeping the telephones on the ears, as in ordinary reception, even in the case of very strong signals.

An impedance equivalent to that of the telephones is connected in series with the telephones in the plate circuit of the receiving

set with a switch provided to short circuit it when necessary. The telephones are shunted with a variable non-inductive resistance which can be switched out of the circuit when not in use. With the equivalent impedance in the circuit the shunt across the telephones is varied until the received signal becomes as weak as desired. The shunted telephones may then be switched alternately from the receiving set to the comparator and measurements made in the usual manner with the telephones directly on the ears.

Comparatively weak signals can be measured thru heavy atmospheric disturbances with this method, provided the disturbances are not too continuous, as it eliminates the deafening effect of the strong disturbances.

For ordinary signal measurements the impedance coil is shorted out by a single pole single throw switch and the shunt disconnected from the telephones. Figure 6 shows the schematic diagram.

2. Measurement of Atmospheric Disturbances

The early method of determining the strength of atmospherics was unsatisfactory. In this the disturbances were compared with the 1,000-cycle generator output in the same manner that radio signals are compared. The difference in the sound of the two sources and the variations in the strength of the disturbances made measurements difficult and inaccurate. Often, disturbances were too strong to measure.

The method now in use applies the voltage from the output of the comparator directly to the plate circuit of the receiving set, the high resistances being cut out. Atmospheric disturbances and the 1,000-cycle signal from the comparator can then be heard simultaneously, and the strength of the comparator signals is adjusted until when sending at about twenty words per minute they become just readable thru the disturbances. Hand sending or an automatic transmitter placed in the tuning fork oscillator line can be used to produce the signals. Experiments in the laboratory show that the strength of the signal current in the telephones with this connection, for a certain voltabe divider setting, is for our apparatus as nearly as can be measured five times the strength of the current with the comparator used normally with the 200,000-ohm resistance in series with the telephones. Therefore, the field intensity corresponding to the voltage divider setsing necessary to read thru the atmospheric disturbances, when multiplied by five, is taken as the field strength of the disturbances. Figure 7 shows the connections used.

3. MEASUREMENT OF AMPLIFIED SIGNALS

It is nearly impossible to measure the strength of signals with the telephone comparator in the usual way when audio-frequency amplification is used. When the 1,000-cycle tuning fork is running, the amplifier generally amplifies the residual current induced in the receiver from the leads of the comparator, sufficiently to drown a radio signal of moderate strength. It has been found necessary therefore to open the "zero" side of the tuning fork oscillator leads when the comparison switch is thrown to the radio signal and close it when thrown to the comparator. This can be done by means of an ordinary telegraph relay placed close to the tuning fork oscillator and controlled by a switch which can be thrown simultaneously with the comparison switch. The signals can then be measured in the usual manner.
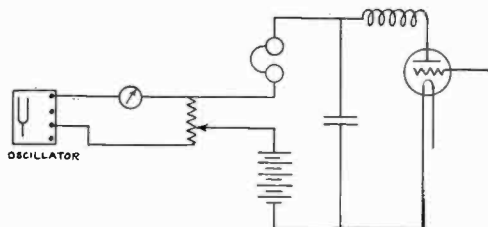
FIGURE 7—Circuit for Measuring Atmospherics

4. APPARATUS FOR MEASURING THE RADIATION HEIGHT OF TRANSMITTING STATIONS

The apparatus described was intended for measurements on stations having antenna currents of 40 amperes to 250 amperes, masts more than 60 m. high, and wave lengths of 6,000 m. and above. The receiving loop is 24.4 m. long, 9.15 m. high, and wound with 15 turns of Number 16 (diameter = 1.29 mm.) hard drawn copper wire, the turns being spaced approximately 15 cm. apart. The frame is swung between two portable masts 14 m. high and 30 m. apart at a distance of about one wave length from the transmitting station to be measured, preferably with level ground between, and no trees in the immediate vicinity of the loop. The direct current resistance of such a loop is approximarely 12 ohms.

The indicating apparatus consists of a Leeds and Northrup portable galvanometer having a coil resistance of approximately 20 ohms, and requiring about 20 microvolts for a deflection of one division; and a Western Electric Company's vacuum thermo-

element, Type 5a, having a heater resistance of about 35 ohms
and a thermocouple resistance of about 10 ohms. The thermo-
element can be calibrated by direct current provided the current
is reversed and the mean of the two galvanometer readings taken,
if any difference is shown. A typical calibration shows 10 mm.
galvanometer deflection for 0.71 milliamperes in the heater of
the thermo-element, and 140 mm. for 2.58 milliamperes.

The transmitting station usually makes one-minute dashes,
with intervals of one minute between. This enables the resist-
ance of the loop to be measured by inserting known additional
resistance and noting the reduction in deflection. A station with
180 m. masts transmitting with 100 amperes at 10,000 m. will
give 1.5 milliamperes in the loop, or 48 mm. galvanometer deflec-
tion assuming a 63-ohm total loop resistance; and indicates a
radiation height of the transmitting station of 120 m. A station
with 90 m. masts, transmitting at 6,000 m. with 50 amperes will
give about 30 mm. on the galvanometer scale.

The calculations expressed both in wave length and frequency
are made from the formula

For Two Antenna:

$$I_r = 120\,\pi\,\frac{I_s\,h_s\,h_r}{\lambda\,d\,R} = \frac{40\times10^{-5}\,\pi\,I_s\,h_s\,h_r\,f}{d\,R}$$

$$\text{then}\quad h_s = \frac{I_r\,\lambda\,d\,R}{120\,\pi\,h_r\,I_s} = \frac{I_r\,d\,R}{40\times10^{-5}\,\pi\,h_r\,I_s\,f}$$

For Loop Receiving:

If the radiation height of the receiving loop is

$$h_r = 2\,\pi\,\frac{area\times turns}{\lambda} = 2\,\pi\,\frac{NHL}{\lambda} = 6.67\times10^{-6}\,\pi\,NHL\,f$$

then the radiation height of the transmitting antenna will be

$$h_s = \frac{I_r\,\lambda^2\,d\,R}{240\,\pi^2\,I_s\,NHL} = \frac{I_r\,d\,R}{26.7\times10^{-10}\,\pi^2\,I_s\,NHL\,f'}$$

when I_r is received current in loop,

  I_s the transmitting current,

  h_s the transmitting radiation height.

  N, H, L are respectively the number of turns, height, and
      length of the loop,

  λ the wave length,

  f the frequency,

  d the distance from transmitting station, and

  R the total resistance in the loop circuit.

  All quantities are expressed in meters, kilocycles, ohms,
      and amperes.

SUMMARY: The method of measuring radio signals here described depends on the determination of the equality of the sound intensities in the telephones from the signal and from a known alternating current preferably giving the same tone. The measured local current is taken from a thousand-cycle tuning fork oscillator and the electromotive force impressed on the telephones is controlled by a voltage divider. The errors of measurement may be under good conditions less than five percent, tho in case of spark signals and atmospheric disturbances when the two tones cannot be matched it will be very much greater. Special arrangements are described for measuring very strong signals and atmospheric disturbances.

DISCOVERY OF A PLACE WHERE NO STATIC COULD BE HEARD IN AUGUST*

By

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(NAVY YARD, BREMERTON, WASHINGTON)

In August, 1921, we found a place at Pop Point on Thorne Arm in Southeastern Alaska, sixteen miles (25.6 km.) east-northeast of Ketchikan, where no static was heard for six days, altho numerous radio stations on the Pacific Ocean and on the Pacific Coast could be heard complaining to each other and asking for repetition of messages because of interference from static.

Tests made with directive antenna, at Astoria, Oregon, for several days in July, 1920, and at odd times later, indicated that the chief static centers affecting receiving at Astoria were probably Mt. Jefferson, Mt. Wilson, Mt. Hood, Mt. St. Helens, Mt. Adams, and Mt. Rainier; Mt. Rainer apparently being the chief center, altho farther from Astoria than Mts. St. Helens and Adams.

Tests made at Bremerton, Washington, two or more times per day from July 26, 1920, to September, 1921, gave Mt. Rainier as the chief center about 90 percent of the time. Mt. Rainier is about fifty miles (80 km.) from Bremerton. Some of the tests were made simultaneously at Astoria and Bremerton, giving cross bearings on Mt. Rainier. In addition to the static maxima received from the direction of Mt. Rainier, occasional storm static centers became apparent first in the southwest and traveled over Bremerton, producing an all-direction effect when near by, followed by a dying-out directional effect in the northeast. From November 15th to April 15th the static was usually too indefinite to judge direction.

Mt. Rainier (another name for which is Mt. Tacoma) rises abruptly from a nearly sea level elevation. To the west of the mountain is a comparatively low land area over which cool air from the Pacific can move toward the mountain and be heated by the sun's rays and warm earth and obtain some dust and considerable moisture. This more or less saturated and dust-bear

ing air may then come quite abruptly in contact with Mt. Rainier, which is 14,471 feet (4,410 m.) high and capped by perpetual glaciers. If the temperature of the air west of Rainier reached as high a value as air reaches in the eastern part of the United States, Mt. Rainier might be the most violent static center in the United States. However, our air averages probably 15° F. (8° C.) cooler than eastern air, and we are bothered less by static. Probably it should also be mentioned that Mt. Rainier is classed as an extinct volcano and that visitors to the top occasionally report that some gases are still given off.
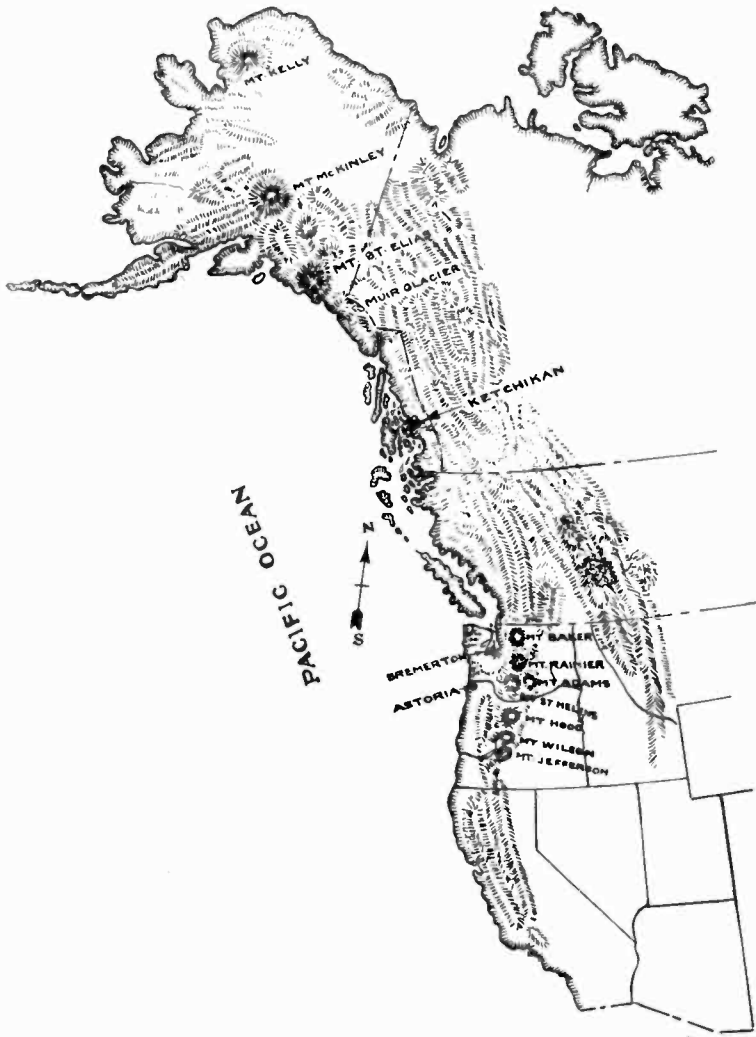
The other peaks previously mentioned are high and cold on top but are not so high as Rainier or do not bear the same relation to warming-up areas for air. Numerous other mountains in the territory to be considered are not snow-covered in summer or else the warm, moisture-bearing air probably does not come in contact with them, and they probably screen off the effect of static and radio signals instead of producing static. The static-producing peaks also apparently have a screening effect for radio signals and probably as well for static disturbances coming from points beyond them.

Passing northward along the Pacific Coast at sea, the effect of static apparently decreases. Passing northward from Tacoma, thru Puget Sound, which is separated from the Pacific by the Olympic peninsula, and then north back of the islands along the coast of Canada and the coast of southeastern Alaska, the static effect apparently decreases on the average until a point is reached in Icy Straits southeast of Mt. St. Elias, where the ocean-going vessels leave the inside passages and go out to sea across the Gulf of Alaska towards Cordova. From that point, just inside of Icy Straits, the static interference apparently gets stronger. Directional measurements from Cordova indicated maxima in the direction of Mt. St. Elias. Mt. St. Elias rises abruptly from the sea to 18,024 feet (5,500 m.) This mountain and the vicinity are covered by many glaciers. However, air from the Gulf of Alaska is probably never very warm because the Pacific Ocean is comparatively cool. Also there have been reports of quite violent static by radio operators located at Yakutat, which is near Mt. St. Elias.

Returning from Cordova we selected a location sixteen miles (25.6 km.) east-northeast of Ketchikan.

In choosing this site we gave particular attention to the fact that the study of natural static is commonly interfered with by man-made static. Not only is it sometimes difficult to distinguish man's static from nature's static, but man-made static sometimes drowns out the natural static. For example: our readings on Mt. Rainier static were somewhat confused by a "Cottrell" precipitator of dust from the smoke of the Tacoma, Washington smelter. That device uses many kilowatts in producing a multitude of sparks. Mt. Rainier's cold top, east of Tacoma, apparently serves to condense warm moisture and dust-bearing air and to precipitate static discharges, thus making little and big sparks. The man-made precipitator west of Tacoma charges the plates of air condensers to high voltages by rectifying high

voltage alternating current thru synchronous rotary spark gap rectifiers. Each air condenser consists of an insulated negative wire in a grounded positive concentric vertical tube one foot (0.305 m.) in diameter. The smoke passes upward thru the tube in the static field between the wire and tube where the dust is precipitated to the wire and wall of the tube, the gases passing on and out the stack. The charges on the tube and wire do not confine themselves to remaining static electricity or even kinetic charges in the form of minute sparks to minute particles of dust, but they jump across five or six inches (12 to 15 cm.) quite frequently and irregularly, between the wire and tube. Also the high voltage alternating current occasionally flashes over. Altogether this is an extreme case of man-made static, some of which is quite like natural static, and all of which is capable of drowning out natural static, radio messages, and broadcasts over a considerable area. Tacoma radio listeners are between the beautiful natural precipitator (which they call Mt. Tacoma) and the economical man-made precipitator on Puget Sound.

The site we chose at Pop Point was sixteen miles (25.6 km.) from the nearest high tension line or other likely source of man-made electrical interference. Having arrived at that site, a directional receiving loop and antenna similar electrically to those we had been using for a year was set up and used. No static was heard for the six days from August 9th to 15th. Nothing could be heard but radio stations and the faint sounds produced by the tubes and batteries. We had avoided both man-made and nature-made static. Ships and shore stations as far north as Cordova, as far west as Honolulu and as far south as southern California were heard. We used the regenerative detector and two stages of audio frequency. However, we did not hear any radio stations east of us or south of us as far inland as Seattle or Mt. Rainier. We were apparently screened from that section. The testing party consisted of Mr. R. L. Barbur and I. Mr. Barbur, in addition to being able to make the direction tests, is an experienced radio operator. On the seventh day static was heard just as we were leaving and after the loop part had been dismantled. This static was apparently from a moving storm which blew in the following day from the southwest.

Leaving this point we went to Ketchikan, where we heard static on a small antenna. From there we went to Sitka and, mounting our directional apparatus on a forty-foot (12.2 m.) gas boat, we cruised around among the islands. The static in the vicinity of Sitka was weak and no definite directions were noted.

The ships' radio operators traversing the inside passages in the island fringe along the coast of Canada and Southeastern Alaska have reported for years their inability to receive radio stations in certain spots. These reports have shown numerous combinations of ability to hear certain stations and not others, which sometimes reverse as the vessel proceeds. Broadcasting has brought out similar reports relative to the screening effect of inland mountains, and our receiving about Bremerton has always been handicapped by such mountains as the Olympics west of us. The screening effect of the high rugged land of this country is apparently much greater than the diminishing effect of the lower land on the east coast.

Those islands of Western Canada and Southeastern Alaska are numerous. The larger islands are mountain high. Some islands carry peaks over three thousand feet (910 m.) in height. The channels between the mountains as a rule are comparatively narrow. In some cases ocean-going vessels pass thru waterways that are not over four hundred feet (120 m.) wide. In many cases standing on the deck of a vessel and endeavoring to see the tops of the mountain, it is necessary to look up as you would look up from the street to see the tops of the buildings in an American city. Also the mainland of Alaska and Canada is a similarly rough and rugged arrangement of mountains and valleys. Tall trees are characteristic of the entire section.

My assumption is that the no-interference point sixteen miles (25.6 km.) east-northeast of Ketchikan was separated from permanent static centers by just enough rough and rugged topography to be screened from those static centers while that point was not similarly screened from the Pacific Ocean and Pacific Coast.

SUMMARY: The origin of natural static near mountain peaks is described and compared, in its mode of production and character, with that produced by smoke precipitators.

Some mountains are believed to be soucres of static, while others act as absorbing shields to signal and static. On these bases, a certain point near Ketchikan, Alaska, was found where static was not heard in midsummer, while signals were received from some directions but not from others.

THE SUPER-HETERODYNE—ITS ORIGIN, DEVELOPMENT, AND SOME RECENT IMPROVEMENTS*

By

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The purpose of this paper is to describe the development of the super-heterodyne receiver from a wartime invention, primarily intended for the exceedingly important radio telegraphic direction-finding service in the Signal Corps of the American Expeditionary Force, into a type of household broadcasting receiver, which, with our present vision, appears likely to become standard.

The invention of the super-heterodyne dates back to the early part of 1918. The full technical details of this system were made public in the Fall of 1919.[1] Since that time it has been widely used in experimental work and is responsible for many of the recent accomplishments in long distance reception from broadcasting stations. While the superiority of its performance over all other forms of receivers was unquestioned, very many difficulties rendered it unsuitable for use by the general public and confined it to the hands of engineers and skilled amateurs. Years of concentrated effort from many different sources have produced improvements in vacuum tubes, in transformer construction, and in the circuits of the super-heterodyne itself, with the result that at the beginning of the present month there has been made available for the general public a super-heterodyne receiver which meets the requirements of household use.

It is a peculiar circumstance that this invention was a direct outgrowth to meet a very important problem confronting the American Expeditionary Force. This problem was the reception of extremely weak spark signals of frequencies varying from about 500,000 cycles to 3,000,000 cycles, with an absolute minimum of adjustments to enable rapid change of wave length. The

technical difficulties of this problem are now so well known that it was not necessary to consider them. Round in England and Latour in France, by some of the most brilliant technical radio work of the war, succeeded in producing radio frequency amplifiers covering the band from 500,000 to 1,000,000 cycles and tho covering a much more limited band, amplifiers operating on 2,000,000 cycles had been constructed. These results had been accomplished by the use of vacuum tubes and transformers of a minimum capacity. As this apparatus was used in the highly important intelligence services, all information was carefully guarded. When the United States entered the war, the facts that it was necessary to produce sensitive receivers for short wave lengths and that tube capacity would prove the bar to a straightforward solution of the problem were not known in this country. As a result, no attention was paid to the capacity in the type of vacuum tube which was adopted, and while the tube met the requirements of the lower frequencies admirably, it was impossible to use it effectively for the frequencies of importance in the direction-finding service.

During the early part of 1918, thru the courtesy and energy of General Ferrié and his staff, the American Expeditionary Force was supplied with apparatus of French manufacture. It was quite apparent, however, that this source of supply could not be a permanent one, and a solution of the problem became essential. During the early part of 1917, I had made a careful study of the heterodyne phenomena and their effect on the efficiency of amplification. With this work freshly in mind, the idea occurred to me to solve the problem by selecting some frequency which could be handled by the tubes available, building an effective amplifier for that frequency, and then transforming the incoming high frequency to this readily amplifiable value by some converting means which had no low limit; preferably the heterodyne and rectification. The principles and advantage of this method were explained in a paper presented before this INSTITUTE[3] and are now so well known that no further explanation is required here.

After much experimental work, an eight-tube set was constructed consisting of a rectifier tube, a separate heterodyne oscillator, three intermediate frequency amplifiers, a second

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[2] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, April, 1917. Presented October 4, 1916.

[3] This amplification is based on the ratio of the voltage applied to the second detector to the voltage at the loop terminals. The intermediate frequency amplification is unknown.

rectifier or detector, and two audio frequency stages. The intermediate frequency stages were coupled by tuned air-core transformers set for a frequency of about 100,000 cycles, with an adjustment for controlling the regeneration. The amplification of voltage measured at the input of the second detector with the amplifier just below the oscillating point, was about equivalent to a radio frequency amplification of 500.[4] This is illustrated in Figure 1 and the arrangement of its circuits in Figure 2. It gave satisfactory results except that the inclusion of a regenerative control on the intermediate frequency amplifier made skilled handling necessary, as the adjustment of the frequency of the oscillator changed the plate current of the detector tube and this, in turn, varied the resistance which that tube introduced into the amplifier system and upset the regenerative adjustment.
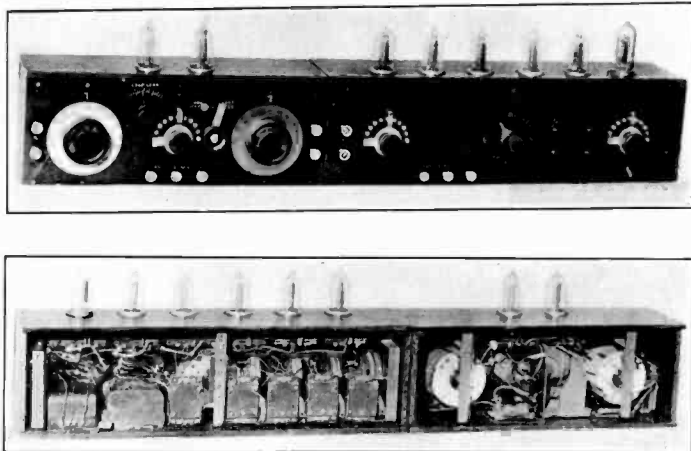
FIGURE 1

The Armistice ended development at this point, but in the fall of 1919, for the purpose of determining the results which could be obtained by pushing the super-heterodyne method of reception to the limit, a resistance-coupled intermediate frequency amplifier consisting of five high-μ (amplification factor) tubes was constructed. The voltage amplification of these five stages was probably between 5,000- and 10,000-fold. While greater amplification could have been obtained, the sensitiveness of a set composed of a two-tube frequency converter, a five-

[4] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, February, 1921. Presented December 3, 1919.

tube intermediate frequency amplifier, a detector, and one-stage of audio, was such that on a three-foot (one-meter) loop, the sole criterion of reception was simply whether the signal was stronger than the atmospheric disturbances.

The sensitiveness of the super-heterodyne was demonstrated during the winter of 1919-1920, when the spark signals from amateur stations on the West coast and telephone signals from destroyers in Southern waters were received in the vicinity of New York on a three-foot (one-meter) loop. Probably the most striking demonstration of the capabilities of the method occurred in December, 1920, when Paul F. Godley, at Ardrosson, Scotland, received the signals of a large number of amateur stations located in the United States, many of them being spark stations. The super-heterodyne used by Godley consisted of a regenerative tube for the rectifier, a separate oscillator, four stages of re-

3 Stages of Amplification
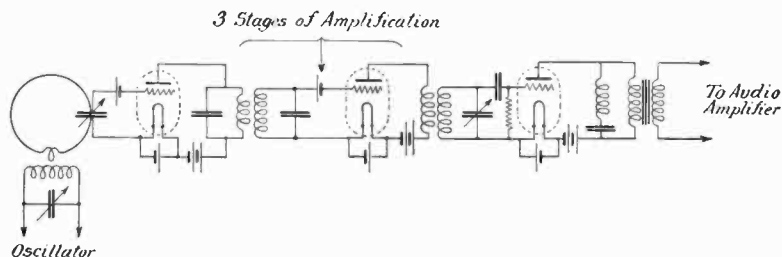
FIGURE 2

sistance-coupled intermediate frequency amplification, a second rectifier, and two stages of audio. While it is difficult to state definitely the actual voltage amplification obtained, it appears to have been between 3,000- and 5,000-fold.[5]

With the coming of the broadcasting art, and with the great increase in the number of stations and the consequent interference, the super-heterodyne began to take on a new importance —an importance which was based not on its superior sensitiveness nor on its selectivity, but on the great promise which the method offered in simplicity of operation. It was, and still is, the standard practice to furnish the public with receivers equipped with a variety of tuning adjustments for the purpose of amplifying the desired band of radio frequencies and excluding all others. As a matter of fact, many more adjustments than are on receivers should be used—more than could be placed in the hands of the average user. It would obviously be of the

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[5] Based on standard previously described. This is without the second heterodyne which was used in receiving continuous waves.

greatest importance if in some way these tuning adjustments could be made in the laboratory by skilled engineers and sealed, leaving some relatively simple adjustment for the hands of the operator. The super-heterodyne offered the ideal solution. This solution lay in the construction of an intermediate frequency amplifier which would amplify a given frequency and a band 5,000 cycles above and below it and which would cut off sharply on either side of this desired band. The adjustments necessary to accomplish this could all be made by skilled men, and the only operations left for the user would be the two adjustments necessary to change the incoming frequency down to the band of the amplifier—adjustments which are not dependent on each other, which are of extreme simplicity, and which can be made equally well by the novice or the engineer. To determine just what could be accomplished along these lines, the writer, working in conjunction with Mr. Harry Houck, constructed during the spring of 1922, a set designed for the maximum usable sensitiveness and selectivity. The set-up consisted of one radio frequency stage (non-tuned transformer) a rectifier tube, and oscillator tube (used as a separate heterodyne), a three-stage iron-core transformer-coupled intermediate frequency amplifier designed to cover a a band of 20,000 to 30,000 cycles, a second detector tube, and two-stage of audio frequency amplification. UV-201 tubes were used. The set without the audio frequency amplifier is illustrated in Figure 3 and Figure 4. To prevent the intermediate frequency amplifier from oscillating, each stage was shielded separately. The use of a radio frequency stage ahead of the first detector possesses a number of advantages, but the chief one is in eliminating the reaction between the loop circuit and the oscillator circuit. Experience with the original type had shown that when an oscillator of ordinary power was used, it was necessary to couple it rather closely with the loop circuit in order to insure a sufficiently strong heterodyne current. This close coupling affected the tuning of both circuits, an adjustment of one changing the setting of the other. To avoid this trouble and to produce a system wherein a station could always be tuned in on exactly the same settings, a single stage of radio frequency amplification (non-tuned transformer) was used, and the oscillator was coupled into this transformer. This arrangement eliminated the reaction, reduced the radiation to a minimum and, in addition, removed the damping of the first rectifier from the loop circuit and improved its selectivity.

The results obtained with this set were about as expected.
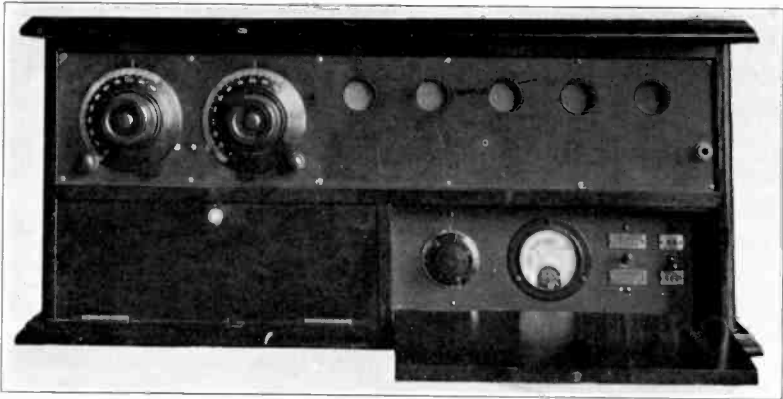
FIGURE 3
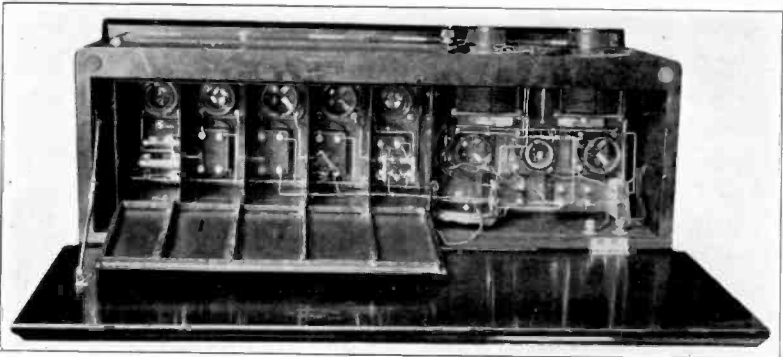
FIGURE 4

On a three-foot (one-meter) loop, the factor determining the reception of a station was solely whether the signal strength was above the level of the atmospherics. The selectivity was such that stations which had never been heard before on account of blanketing by local stations, were received without a trace of interference. While the performance of the set was much superior to any other receiver, it was apparant that the cost of construction and maintenance was prohibitive. The single item of a ten-ampere filament current will give some idea of the size of the storage battery and auxiliary apparatus required.

With the coming of the low filament consumption, or dry battery type of tube, the possibilities of producing a super-heterodyne for household use were tremendously improved. The set of Figure 3 was remodelled for the WD-11 tube, and its sensitiveness was brought to about the same value as obtained with the

storage battery tubes. This was a long step forward, but still the cost was prohibitive.

It has been apparent ever since the question of the application of the super-heterodyne to broadcasting had been considered, that there were too many tubes performing a single function which were quite capable of performing a double one. The most outstanding case is that of the separate heterodyne oscillator. In view of our knowledge of the self-heterodyne, it appears quite obvious to perform the first rectification by means of a self-heterodyne oscillator and thereby save a tube. As a matter of fact, this was one of the very first things tried in France, but, except for very short wave lengths, it was never very successful when a high intermediate frequency was necessary. The reason was this: If a single tuned oscillating circuit was used, the mistuning to produce the proper beat caused a loss of signal strength which offset the gain of a tube. If two tuned circuits were used on the oscillator, one tuned to the signaling frequency and the other arranged to oscillate at the heterodyne frequency, then on account of the relatively small percentage difference in frequency a change in the tuning of one circuit changed the tuning of the other. The solution of this problem was made by Houck, who proposed an arrangement so simple and so effective that it completely solved the problem. Houck proposed to connect two tuned circuits to the oscillator, a simple circuit to the frequency of the incoming signal and a regenerative circuit adjusted to oscillate at such a frequency that the second harmonic of this frequency beating with the incoming frequency produced the desired intermediate frequency. The general arrangement is illustrated by Figure 5.

Incoming Signal   A   B   C   D   To Amplifier
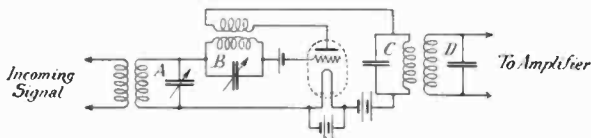
FIGURE 5

In the diagrammatic illustration, circuit A is tuned to the incoming signal, circuit B is tuned to one-half the incoming frequency plus or minus one-half the intermediate frequency, and the circuits C and D are both tuned to the intermediate frequency. The operation of the system is in line with ordinary self-heterodyne action. By reason of the asymmetrical action of the tube, there are created in the circuits a variety of har-

monics. The second harmonic combines to produce beats with the incoming signals of the desired intermediate frequency, the tube rectifies them to produce the desired intermediate frequency and, thru C and D, the new frequency is supplied to the amplifier. On account of the fact that circuits A and B are tuned to frequencies differing by approximately 100 percent, a change in the tuning of one has no appreciable effect on the tuning of the other. This arrangement solved the oscillator problem and, in addition, practically eliminated radiation.

The next step in the reduction of the number of tubes was to make the radio frequency amplifier perform the function of amplifying intermediate frequency as well. This can be done with none of the difficulties inherent in audio frequency amplification, as the very small amplitudes of voltage handled by the first tube preclude the possibility of the grid becoming positive with respect to the filament. The general arrangement of circuits for carrying this out is illustrated by Figure 6. In this

To 2nd Stage
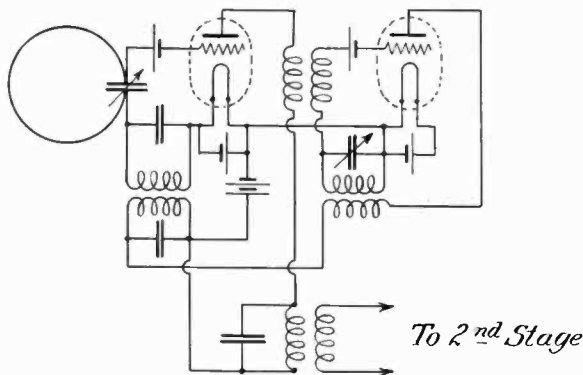
FIGURE 6

arrangement the signals received by the loop are amplified at radio frequency by the first tube and applied to the grid of a second harmonic oscillator by means of an untuned radio frequency transformer. The combined signaling and heterodyning currents are then rectified by the second tube, producing a current of the intermediate frequency which is applied to the grid of the first tube, amplified therein, and passed on to the second stage of the intermediate amplifier. A more practical method of carrying out this idea is illustrated in Figure 7. In this arrangement, a secondary of the first intermediate frequency transformer is connected to the grid of the first tube and in parallel

with the loop circuit. Otherwise, the arrangements of Figures 6 and 7 are identical. The parallel type of circuit arrangement eliminates a variety of reactions which would give rise to oscillations of various frequencies and in addition, prevents the reception of long wave signals by the intermediate frequency amplifier. When this development had been completed, improvements in the design of the intermediate frequency transformers made it possible to obtain with two stages all the amplification which could be used.

To 2 nd Stage
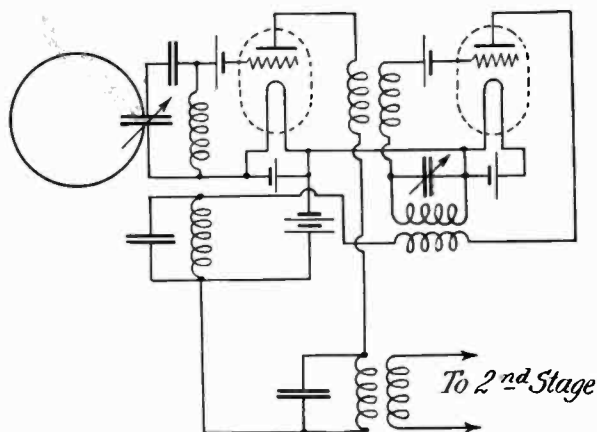
FIGURE 7

On account of the high amplification, signals from local stations overload the second rectifier and introduce distortion. Control of the amount of intermediate frequency amplification is essential. While there are numerous methods equally effective, the simplest one appears to be the control by means of the filament temperature of the second intermediate frequency amplifier.[6]

The features just described were all incorporated in the receiver, which is illustrated in Figures 8 and 9. The set measured 18 by 10 by 10 inches (45.6 by 25.4 by 25.4 cm.) and was completely self-contained—the batteries, loop antenna, and speaker mechanism being enclosed in the box. The results were highly satisfactory, and loud speaker signals (at night) in the vicinity of New York were obtained from stations in Chicago and Atlanta. It demonstrated that not only could a household receiver

[6]Although some form of potentiometer type of control of the voltage (a. c.) applied to the grid of one of the amplifier tubes would obviously be better, the simplicity of the filament control has many advantages in manufacture.

of the super-heterodyne type be built, but that the first practical solution of the portable set was at hand.
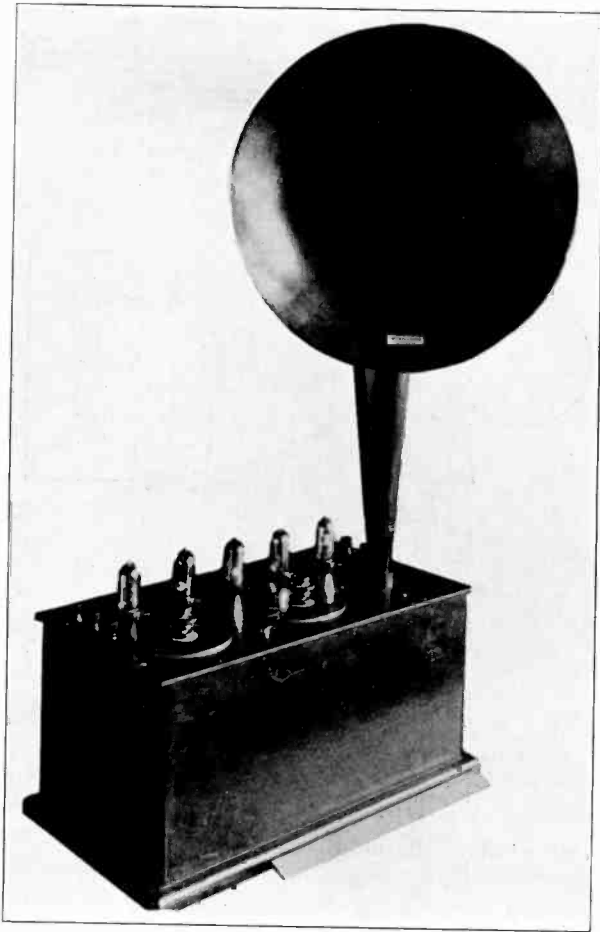
FIGURE 8

In this form, the capabilities of the set were brought to the attention of the Westinghouse Electric and Manufacturing Company and the Radio Corporation of America a little over a year ago. Its possibilities were instantly visualized by Mr. David Sarnoff, who immediately took steps to concentrate the resources of the research laboratories of the Radio Corporation of America, the Westinghouse Electric and Manufacturing Company, and the General Electric Company on this new development. From that point on it passed into a new phase—that of placing an in-
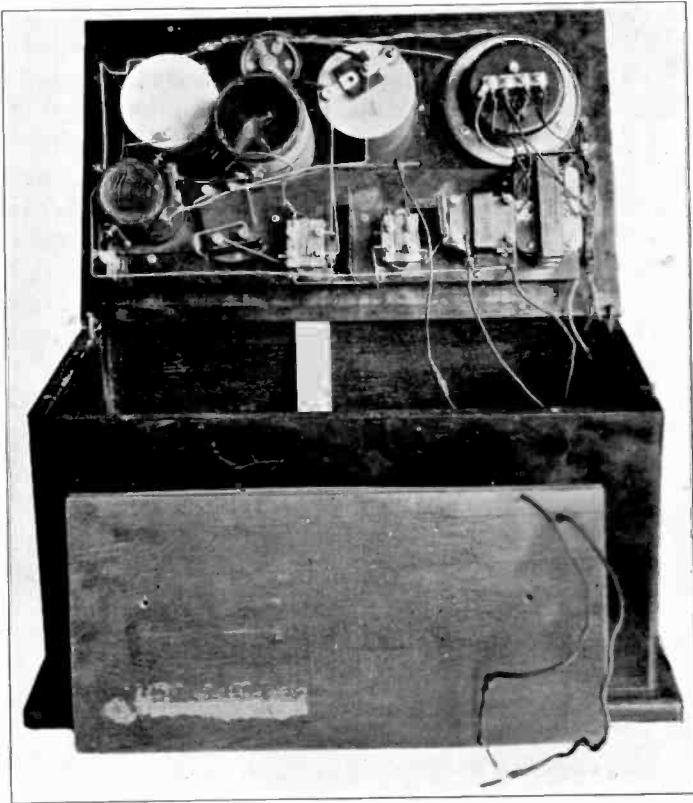
FIGURE 9

vention in a commercial form. In the limited time available, this was a most extraordinarily difficult proposition, and credit for its accomplishment is due to the untiring efforts on the part of the engineers of the above organizations. Many improvements and some radically new ideas of designs have been introduced, but it is the privilege of those responsible for them to present these. In the final development, an additional stage of audio frequency amplification was added in order to insure operation within steel buildings, particularly those within the city limits where signals are relatively very weak compared to suburban locations. This makes a six-tube set, but six tubes can be readily operated on dry batteries and the increase in sensitiveness is well worth the extra tube.

Some idea of the sensitiveness and the ease of operation of the set illustrated in Figures 10 and 11 may be gathered from an incident during the trans-Atlantic broadcasting tests of November and

December, 1923. On December 1st, two ladies, neither having any technical radio knowledge, received loud speaker signals from station 2LO, London, England. This was accomplished at Merrimac, Massachusetts, with the set and loop illustrated in Figures 10 and 11 and probably constitutes a record for the first radiophone reception from Europe with a portable receiver. With the same set and a three-foot (one-meter) loop, loud speaker signals from broadcast stations on the Pacific Coast were received in the vicinity of New York on an average of three or four times a week. The factor determining reception was simply whether the signal strength was above the level of the atmospheric disturbances.
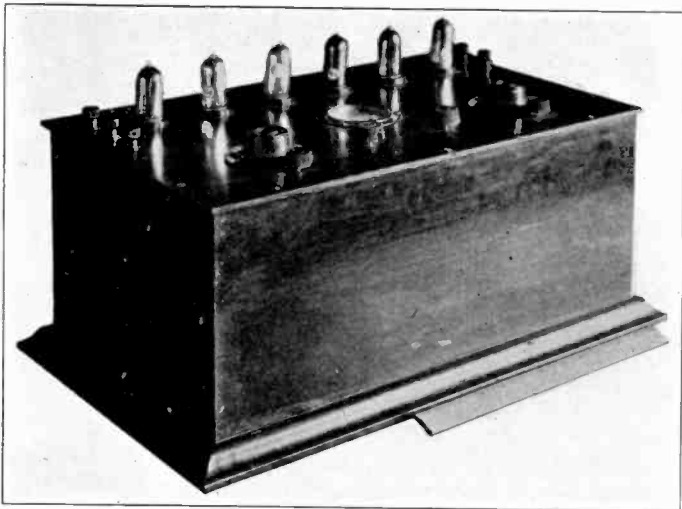
FIGURE 10

The type of super-heterodyne described is now available to the public in the two forms illustrated in Figures 12 and 13. Each of these sets incorporate the arrangements herein described. Their sensitiveness is such that, with a two-foot (61-cm.) loop and an unshielded location, the atmospheric disturbances are the criterion of reception. Here we reach a milestone in the development of broadcast receivers, for no increase in the distance of reception can now be obtained by increase in the sensitiveness of the receiver. Unless the power of transmitting stations is increased we are about at the limit of the distance which can be covered. Future improvement of this receiver will lie along the line of selectivity and simplifying the construction.

SUMMARY: This paper describes the development of the super-heterodyne receiver from a wartime invention into a commercial form of broadcast receiver apparatus now available to the general public. The success of the development is due to the low filament consumption vacuum tube and to the reduction in the number of tubes required by self-heterodyning, reflexing, and improvement in transformer design.

Instances are cited of trans-Atlantic and trans-continental reception of broadcast stations by completely portable sets constructed in accordance with the methods described.
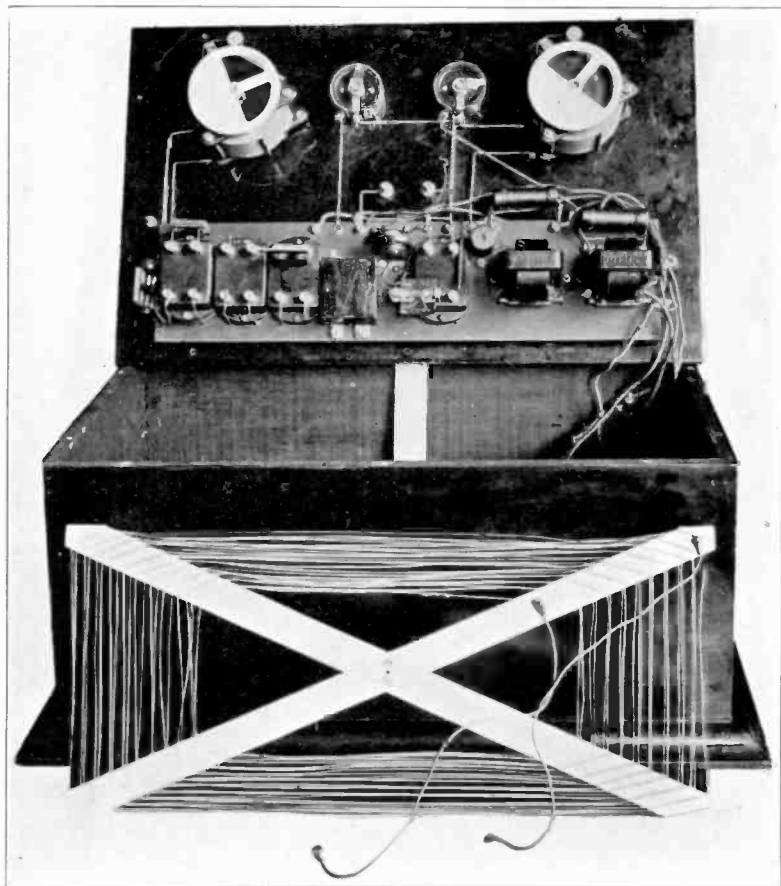
FIGURE 11
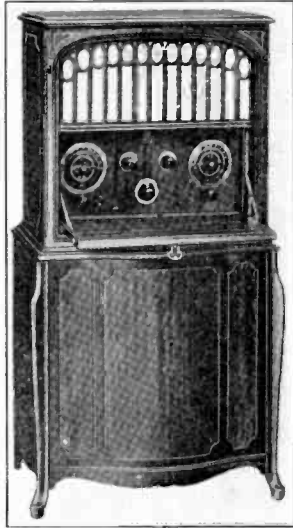
FIGURE 12
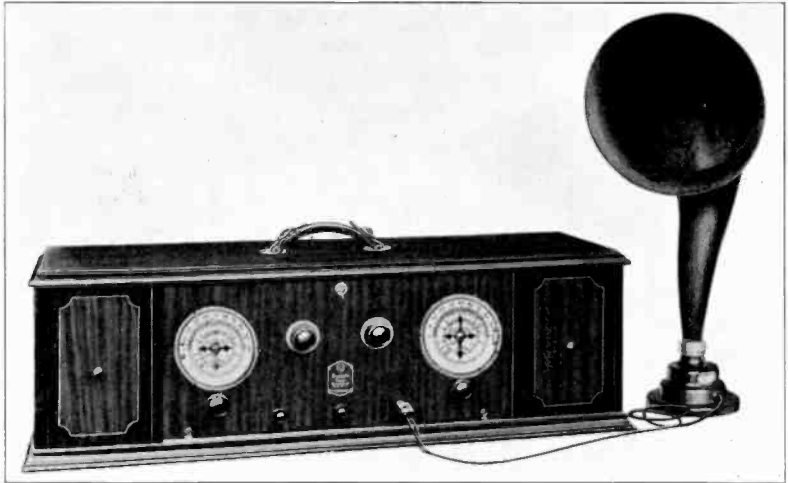
FIGURE 13

TRANSMITTING EQUIPMENT FOR RADIO TELEPHONE BROADCASTING*

By

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(Research Laboratories of The American Telephone and Telegraph
Company and the Western Electric Company, Incorporated)

One of the most remarkable industrial developments of recent times is the advent of radio telephone broadcasting, which within the past three years has become a matter of national importance supporting a rapidly expanding industry. That a tremendous popular enthusiasm for this new art has been aroused cannot be denied. No better evidence is available than the fact that, in the face of an economic outlook which is indefinite, to say the least, some five hundred stations are now in operation, many of them representing investments of tens of thousands of dollars.

Several factors have apparently contributed toward this rapid growth. In the early stages, no doubt, the novelty of the experience was of considerable importance. The discovery of the possibilities of this latest communication development exerts a powerful stimulative influence upon the imagination which is not easily forgotten. The fact that radio presents qualities which render it ideal as a hobby has also undoubtedly aided. There are few fields in which equally satisfying results can be obtained at the expenditure of so little effort and expense while, on the other hand, consistently good performance demands a high degree of skill and more or less elaborate equipment so that interest is likely to be sustained. A more important element than either of the foregoing, however, is believed to be a keen appreciation on the part of an increasing proportion of the public of what radio is and may become as a medium of education and entertainment. The already demonstrated possibility of establishing direct personal contact with the world's most gifted minds, of hearing the greatest artists and the most renowned musical organizations, of sitting at the banquet table with the world's leaders in every field of human activity, of following athletic

events of national interest play by play in distant cities, all of these things without leaving one's own fireside, constitute an appeal which even the imperfections attending the establishing of a new industry cannot destroy.

On the basis of this analysis the fundamental requirement to be met by a broadcasting equipment becomes clearly defined. It is obvious that the public in listening to the voice of the great artist desire,s above all, a natural and faithful reproduction. The slightest change in tone, any embellishment incidental to the transmission, represents a condition against which the industry must contend. Perfect imagery is the goal and future success depends upon how nearly it can be attained.

General Transmission Considerations

In general, the material to be transmitted may be divided into two classes, speech and music. There is no reason to believe, however, that the general transmission requirements for these two classes are essentially different and, in view of the important part which musical entertainment is playing and will probably continue to play in all broadcast programs, it is thought to be advantageous to define these requirements in terms of music.

Experience indicates that for the successful transmission of music, consideration must, in general, be given to frequencies up to 5,000 cycles per second and down to at least 50 cycles per second. The upper limit mentioned (5,000 cycles) lies somewhat above the range of most musical instruments, but it is nevertheless essential that the transmission band of the equipment extend to some such value. Tone quality, that characteristic by which the ear distinguishes between the note of the piano and the banjo, or between that of the violin and the flute, is a matter of overtones in varying amplitude. Naturalness in transmission, therefore, requires that these harmonics be reproduced. Further, it has been shown that the higher harmonics contribute materially to the artistic effect. Musical "brilliance" is apparently merely a matter of their presence in proper proportion.

The lower end of the frequency range is of even greater importance or perhaps it may more properly be said that it appears to be, since difficulty is usually encountered in obtaining satisfactory transmission for the lower registers. One of the most common criticisms of present day radio equipment, and probably of the phonograph as well, is that it discriminates markedly against the bass. Good reproduction is obtained for most kinds of music by utilizing a frequency range extending down to 50

cycles. In certain types of music, however, in particular that of the pipe organ, considerably lower frequencies occur. The fundamental of certain of the percussion instruments, notably the bass drums, is also quite low.

The fact that some of the notes of certain instruments lie outside of the frequency band for which the apparatus is designed does not necessarily preclude satisfactory transmission. It is a remarkable circumstance that in certain cases the fundamental and lower overtones may be entirely suppressed without appreciable change in pitch and with very little change in quality.[1] This phenomenon is attributed to the production of subjective tones due to the non-linear characteristic of the ear. To what extent it can be depended upon to facilitate transmission of the lower frequencies is problematical. It would appear to promise little relief in so far as the percussion instruments are concerned, since in that case many of the overtones are not true harmonics, a fact which may explain the difficulty usually experienced in reproducing the drums. On the other hand, this action probably accounts for the astonishingly good performance of certain apparatus and renders it difficult to name with complete assurance the lower frequency limit required for satisfactory transmission. It will be interesting to watch the development of practices with respect to this point as the art progresses.

A second important characteristic of musical transmission is the range in volume, which for certain classes of music may represent a power change in the ratio of as great as 100,000:1. Orchestral and pipe organ numbers are particularly noteworthy in this respect. Volume is of especial importance from an engineering point of view in that it determines the load conditions under which the equipment will be operated. As broadcasting transmitters are now designed, the factor which determines their useful output range is distortion. The point at which distortion becomes objectionable under increasing load is generally quite clearly defined, being followed by a more or less abrupt change in quality. The phenomenon is usually the result of the peak alternating grid voltage in some one of the amplifier stages exceeding the polarizing voltage in the positive portion of the cycle or during the negative swing momentarily interrupting the plate current. In either case, the wave form of the transmission is modified by the peaks being flattened on one or both sides, and to the listener the music appears to break into a jumble of harsh

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[1] "Physical Criteria for Determining the Pitch of a Musical Tone," H. Fletcher, "Physical Review," September, 1923.

noise. On account of its nature, this effect has come to be known as "overloading."

It is evident that if the volume is to vary in a ratio approaching that mentioned for many musical numbers, the average will have to be set at a relatively low level in order to prevent overloading during fortissimo passages. In general, this will result in weaker signals at receiving stations and bring about a decrease in the effective range of the equipment. The practice has grown up, therefore, of manually controlling the volume by adjusting the amount of amplification employed so that while consideration is given to the demands of musical interpretation, the volume is held within considerably narrower limits, perhaps not greater than 1,000:1. While such practice admittedly precludes a really natural reproduction, experience has indicated that this discrepancy is more than compensated for by the resulting overall improvement in general transmission conditions and, as a matter of fact, the volume variation experienced with most studio music falls very nearly within this narrower range.

From another point of view, the volume range is of further importance in that it determines the permissible noise levels in the various apparatus units. At numerous points in the system slight variations in the electromotive forces of batteries or generators, mechanical vibration, induction from adjacent circuits, variable leakage paths, and similar phenomena introduce disturbances which appear as noise accompanying the transmission. It is obvious that if this noise is not to be objectionable it must be kept well in the background. In other words, its level must be low with respect to the minimum value attained by the transmission.

Apparatus Requirements

Thus far the discussion has been concerned with fundamental considerations which are in general applicable to any system for the transmission of speech and music. In proceeding to the specific requirements for a radio broadcasting equipment, it seems desirable to examine the various aspects of the problem with direct reference to the apparatus units to which they apply. For this purpose, it is convenient to divide the equipment into five groups along the following lines:

> (1) Input Equipment.
> (2) Radio Transmitting Equipment.
> (3) Power Equipment.
> (4) Control Equipment.
> (5) Radio Receiving Equipment.

The input equipment comprises the telephone transmitter or microphone, and the associated audio-frequency amplifiers.

The design of a suitable telephone transmitter for broadcasting service is admittedly one of the difficult phases of the problem. In a paper presented to the American Institute of Electrical Engineers and published in the Journal,[2] Messrs. I. W. Green and J. P. Maxfield have discussed in some detail the requirements to be met by a transmitter in a high quality transmission system and have described the condenser and double button carbon types of instruments. Both of these types have now been developed to the point where they are practically distortionless and both have found extensive application in the broadcasting field.

At the present time, the carbon transmitter is more generally employed on account of its considerably better efficiency. For broadcasting applications, however, a three-stage amplifier is usually provided for use with the carbon transmitter and an additional two stages, making a total of five, for the condenser transmitter. A second advantageous feature of the carbon transmitter is its relatively low impedance, which is of the order of 200 ohms. This characteristic makes it possible to separate the instrument from the amplifier by distances of several hundred feet without experiencing serious transmission losses or trouble due to inductive disturbances.

The advantages of the condenser transmitter are its somewhat better frequency characteristic and its freedom from carbon noise. For laboratory purposes and for the high quality transmission of the most difficult types of music it has, therefore, found some application. As has been indicated above, however, approximately five stages of amplification are subsequently required to attain ordinary telephone volume levels. The rather small capacity of the instrument (of the order of 400 micro-microfarads) also represents a more or less serious limitation to the extent of requiring cables between the transmitter and the amplifier having a very low electrostatic capacity.

It is interesting to note that in popular discussion the "microphone" is frequently regarded as the prime offender in case of poor transmission, and as the secret of success if satisfactory results are obtained. It is significant, however, that before modulation actually occurs, the output of the telephone transmitter or microphone is often passed thru as many as seven stages

[2] "Public Address Systems," I. W. Green and J. P. Maxfield. "Journal of the American Institute of Electrical Engineers," volume XLII, number 4, April, 1923.

of amplification in which the power is increased several million times. That such a feat is possible at all is a glowing tribute to present day amplifier technique, and this fact should be given due weight when passing judgment on the "microphone."

The principal requirements to be met by the amplifier associated with the telephone transmitter in the input equipment are those relating to frequency range, volume range, and noise level, which have already been discussed. A further matter of some importance is that concerned with amplification control. The necessity of providing for manual variation in amplification to reduce the range in volume in the higher powered stages has already been mentioned.

The radio transmitting equipment must afford means of generating alternating currents of a frequency corresponding to the operating wave length, means for modulating these currents faithfully in accordance with the desired transmission, and suitable radiating means, in the form of an antenna.

The vacuum tube oscillator in its present form is a very satisfactory solution to the problem of generation. Since a broadcasting station usually operates on a single assigned wave length, a relatively simple circuit arrangement can be employed. Under the present regulations, however, all coastal stations must be capable of transmitting for ship emergency service on wave lengths of 300 and 600 meters. The equipment must, therefore, admit of adjustment on occasion to these wave lengths.

Other requirements arising from the provisions of the radio regulations relate to frequency stability and to freedom from harmonics. Frequency stability is an almost essential pre-requisite to the success of any scheme looking to the intensive utilization of the ether such as has been put into effect within the past year. If stations are to be successfully operated with frequency intervals between adjacent wave lengths as small as 10,000 cycles, a high degree of constancy in this respect must obviously be maintained, not only during the course of any individual program, but also from day to day and from month to month. For those who have not had the privilege of listening thruout an entire evening to the beat-note from a distant transmitter superimposed upon the program of their favorite broadcasting station, it may be stated that if proper precautions are taken no difficulty is ordinarily experienced in holding the frequency constant within considerably narrower limits than those specified by the present regulations, that is ±2,000 cycles. Since common use is made, however, of transmitters in which the antenna forms part of the

oscillatory circuit or is coupled thereto, it is evident that such performance requires the antenna to be so secured that even the effect of a high wind will not bring about a material change in its constants.

With respect to freedom from harmonics, the present regulations covering broadcasting stations merely contain the provision that the transmitted wave shall be free from harmonics. Unfortunately, this requirement is very nearly an exact statement of the case. The sensitivity of modern receiving equipment and the extremely small amount of power necessary to produce an audible beat-note require the suppression of harmonics to a most unusual degree and render the problem exceedingly difficult. Apparently satisfactory performance is now being attained, however, thru the use of properly designed coupled circuits.

Primarily, the system of modulation employed in a broadcasting transmitter must be one which will enable it to meet the rigorous transmission requirements. On account of its excellence in this respect, as well as its simplicity, the arrangement commonly known as "constant current" or "plate current" modulation has found extensive application. In this system, the amplified transmission is superimposed upon the direct voltage impressed upon the plate of the oscillator tubes. Under these conditions, the output voltage of the latter varies proportionally, producing the desired modulated radio-frequency current in the antenna circuit. The operation of such a system has been discussed at length in a paper presented to THE INSTITUTE OF RADIO ENGINEERS in 1921 by Mr. R. A. Heising,[3] and need not be reviewed here.

The power equipment for a radio broadcasting station is subject to no requirements of an unusual nature with the possible exception of freedom from electrical noise. This factor is of no little importance, however, and exercises considerable influence on the choice of power sources for use in the various circuits.

The effect of slight voltage variations in producing disturbances which appear to the listener as noise has already been mentioned. In general, it is essential that all power sources be substantially constant in voltage. The absolute value of the allowable voltage fluctuation is variable, depending upon the position of the battery or generator in the circuit. If associated with one of the early stages where the transmission level is low and where any fluctuation will be subsequently amplified to a considerable extent, relatively rigid requirements are imposed and batteries

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[3] "Modulation in Radio Telephony," R. A. Heising, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 9, number 4, August, 1921.

may be required. If, on the other hand, the source supplies power to one of the later stages, relatively large fluctuations in voltage may occur without producing effects comparable in magnitude with the desired transmission, and rotating machines or rectifiers may prove to be entirely satisfactory. The power supply to the plates must, in general, meet more rigorous requirements than that for the filaments. If, however, grid biasing voltages are obtained by the use of resistances in the filament circuit, a rather common practice, the current in that circuit must, for obvious reasons, be maintained substantially constant.

When the power sources themselves are not inherently free from noise to the required degree, the most direct solution lies in the use of suitable electrical filters. Considerable improvement can generally be effected in this manner, but such combinations are not always practical possibilities on account of the cost and bulk of large condensers to withstand high voltages and of large inductances to carry relatively heavy direct current.

Another requirement of some importance applying particularly to power sources in the filament circuits, is that relating to voltage regulation. The life of a vacuum tube is very definitely limited by that of the filament. The latter slowly evaporates and in time burns out. This process is necessarily a function of the filament current and it is an experimental fact that an increase in the current of a few percent above the normal value will reduce the life to one-half or even less, depending upon the type of filament, the design of tube, and the service which it is called upon to give. Close voltage regulation is, therefore, essential and is one of the factors to which particular attention must be given in preparing specifications for filament machines.

A third requirement of considerable commercial importance is that of meeting the power supply available at the point of installation. For a number of reasons the majority of broadcasting stations now in service are building top stations. Many of them are operated by newspapers and stores having isolated power plants, or are located in business districts where alternating current power is not readily obtainable. Under these conditions, the application of rectifiers as a source of high voltage power is at an obvious disadvantage, and direct current generators appear to be the most satisfactory solution.

In broadcasting, to an even greater extent perhaps than in other amusement enterprises, the program must go forward smoothly and briskly to obtain the most favorable reaction from the audience. Delays of even a relatively few seconds often be-

come painfully evident since the sense of hearing alone is involved, and as soon as the transmission is interrupted the listener is beyond the influence of the personality of the entertainer. It is of primary importance, therefore, that such facilities as are required to expedite the program to the utmost should be provided.

The control equipment required depends largely upon the class of service in which the station operates. The smaller stations necessarily serve a limited audience and often employ considerable mechanically reproduced music. Under these conditions, the simplest possible arrangement will generally suffice. On the other hand, many of the larger stations now number their audiences in tens of thousands, and the talent available often includes figures of national or international prominence. Two or more studios are frequently employed, and the use of wire line connections to outlying points is finding increasing application. Under such circumstances, a well-worked-out control system is essential.

In general, the facilities required may be classified into three separate groups. In a broadcasting station, it is just as important to close and open the circuit promptly at the proper time as it is to raise and lower the curtain in the theatre. There is also the problem of insuring that nothing but the desired transmission is broadcast, since failure in this respect undoubtedly detracts materially from the artistic effect produced and often results in considerable embarrassment to the station personnel or to the talent. These conditions require that control over the outgoing transmission be vested in the studio director or announcer and that suitable provision be made in the studio for this purpose. The apparatus employed, in addition to the necessary keys and relays for circuit control, must include suitable indicating means such as a series of signal lamps, so that circuit conditions at any particular instant will be evident to all persons involved.

Another requirement arises from the close cooperation which must be maintained between the studio and operating personnel and the fact that in many stations the studio and operating room are widely separated. A suitable intercommunicating telephone and signal system is almost an absolute essential. Both audible and visible signals are desirable, and suitable means must be provided to prevent such signals being broadcast.

A third requirement relates to monitoring facilities. The operator of a radio broadcasting station has considerably more

important functions than those of telegraphist, "trouble shooter," and wireman. Not only is he the critic upon whom the studio director in most cases must depend for information concerning the effect which the audience is actually observing, but also thru his control of the input volume he himself contributes materially to the excellence of the program or tends to spoil it entirely. It is quite necessary, therefore, that he be provided with suitable apparatus facilities to enable him to obtain the desired results. Preferably, these should take the form of a suitable loud speaker equipment arranged for monitoring the output of the speech amplifier and, by means of a suitable rectifier, the radio frequency output of the transmitter.

In so far as receiving apparatus is concerned the requirements to be met are, at best, somewhat indefinite. Since a large number of broadcasting stations may be classified as "coastal stations," however, equipment suitable for maintaining the receiver watch required by law is essential. In this connection, the original practice, it will be remembered, involved an interruption of the program every fifteen minutes for three minutes to "listen in." Such a procedure was necessarily of short duration and in almost all stations facilities are now provided to permit of reception at 500 kilocycles (600 meters) while the local transmitter is in operation. This requires a separate receiving antenna and imposes certain restrictions upon the design of the receiving apparatus. In addition to its use in the reception of telegraph signals, it is also desirable that the receiver be suitable for employment in the broadcasting range and be capable of the high quality reproduction of speech and music, since, on occasion it may be employed to considerable advantage to supplement the regular monitoring facilities and to observe the programs from other stations.

The Western Electric 101-B Radio Broadcasting Equipment

Having discussed various aspects of the general design problem, a brief description of a commercial system of the type which has found application in some forty of the larger broadcasting stations in the country may be of interest.

The transmitter in its housing is, no doubt, already a familiar object since it has appeared with great frequency in published photographs of broadcasting activities. An idea of the arrangement in more detail, however, may be obtained from Figure 1. On the left is shown the housing with the cover removed. The

transmitter is supported by means of a series of coiled springs to reduce noise effects due to mechanical vibration. A 12-foot (3.7 m.) flexible cord is ordinarily provided which terminates in a standard three-wire attachment cap. When in service, the latter is inserted in a suitable receptacle usually mounted on the baseboard in the studio.
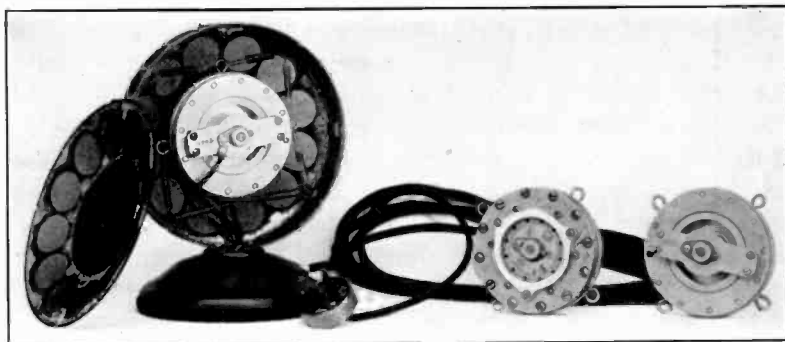
FIGURE 1—Carbon Transmitter and Housing

The appearance of the transmitter itself from the rear, and from the front, respectively, is shown by the two views on the right. A light, thin, metallic diaphragm is employed which is securely clamped between the two heavy exterior rings. An interior ring stresses the diaphragm like a drumhead so that its natural frequency is brought above the range which it is desired to transmit. In this manner "blasting" effects due to diaphragm resonance are practically eliminated. Its performance in this respect is still further improved by the application of air damping. A perforated metal plate is mounted close to the diaphragm in the rear, providing a thin film of air which serves to increase the damping and bring about a much flatter frequency characteristic. Two carbon buttons are employed, one on each side of the diaphragm. One of these is mounted in the center of the damping plate; the other is supported by means of a heavy bridge secured to the clamping rings. It will be noted that, with this arrangement, as the diaphragm vibrates, the pressure on one button is released when that on the other is increased. For this reason it is sometimes referred to as the "push-pull" type. These various improvements have resulted in a transmitter which permits of a reproduction of speech and music not greatly different from the original. As has been indicated, however, two or three stages of amplification are required to attain ordinary telephone volume levels.

The vibratory system of the condenser transmitter is similar in construction to that of the carbon unit. The buttons and bridge are omitted, however, and the rear damping plate is carefully insulated to serve as the second plate of the condenser. The capacity of the condenser thus formed is approximately 400 micro-microfarads and the variation brought about by the vibration of the diaphragm is of the order of one-hundredth of one percent. As has been stated, this device in its present form is less efficient than the carbon transmitter and requires an additional two-stage amplifier if equal volume is to be obtained. Also, it requires special handling in certain respects. On account of its low capacity and correspondingly high impedance, a relatively short cable of low electrostatic capacity must be employed to connect with the initial stage of the amplifier if serious transmission losses are not to be experienced. In addition, its high impedance results in the amplifier being operated with the input of the first amplifier tube practically on open circuit and the latter becomes sensitive to inductive disturbances which introduce noise into the circuit. The fact that four or five stages of amplification is required to attain ordinary telephone volume levels is, in itself, a disadvantage from a noise point of view, since under these conditions, the problem of adequately controlling tube noises, variable leakage paths, battery voltage irregularities, and the like, becomes difficult. On the other hand, the condenser transmitter is the most highly developed device of its kind now available; when properly maintained it is free from the background of noise which is sometimes so objectionable with the carbon instrument, and it affords somewhat better transmission than the latter. Where the very best results are desired, therefore, irrespective of expense, it finds application.

A type of condenser transmitter equipment designed primarily for studio use is shown in Figure 2. The transmitter is mounted in a housing of the usual type and fitted with a special 6-foot (1.8 m.) low capacity cord. The amplifier unit takes the form of a cabinet of the phonograph type with the amplifier proper mounted in the upper portion on a horizontal panel. Adequate shielding is afforded by a metal backing on the panel and a removable metal cover attached thereto. Space is provided in the lower portion for a 6-volt storage battery for heating the filaments and a 200-volt dry battery for the transmitter and plate circuits. The amplifier circuit is shown schematically in Figure 3, which probably requires little explanation. Provision s made for controlling the device from the radio room by means

FIGURE 2—Condenser Transmitter Equipment

of a relay in the filament circuit operated by the current which would normally flow thru the buttons of the carbon transmitter. Since the amplifier is fitted with an output cord similar to that used with the carbon transmitter, it may be plugged into the circuit interchangeably with the latter and employed without change in the operating procedure.

The studio signaling and control unit is shown in Figure 4. As is evident, it takes the form of a wooden cabinet similar in outline to the ordinary subscriber's set, having the necessary
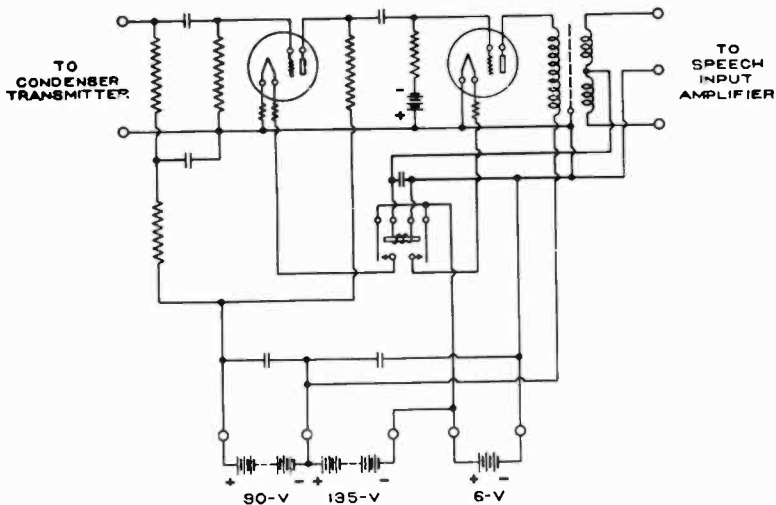
FIGURE 3—Schematic of Amplifier for Condenser Transmitter

switches and signal lamps mounted on the front panel. The telephone handset permits of direct communication with the operating room. The white lamp, key and push button to be seen on the left and at the bottom of the panel are associated with a signaling buzzer circuit. The lamp at the right, which is green, is controlled from the operating room and serves to indicate to the studio director when the radio equipment is in operation and ready for the program to begin. The key immediately
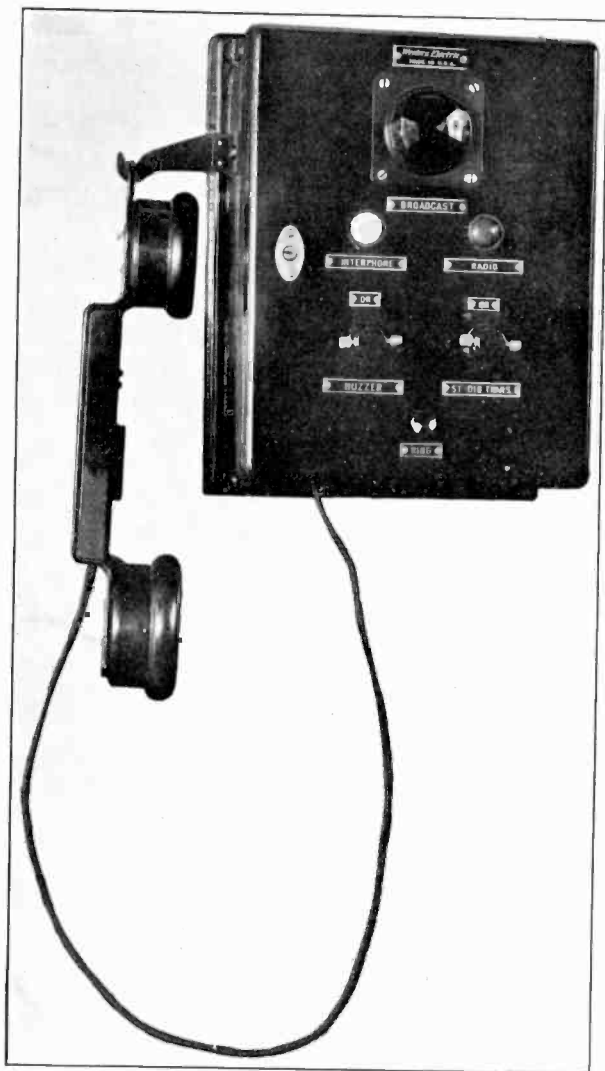
FIGURE 4—Studio Signaling and Control Unit

below it serves to control the studio transmitter circuit and, in addition, when thrown to the broadcast position, opens the circuit of the signaling buzzer and lights a lamp behind the large red bull's-eye at the top of the unit.

The speech input equipment is shown in Figure 5. As will be
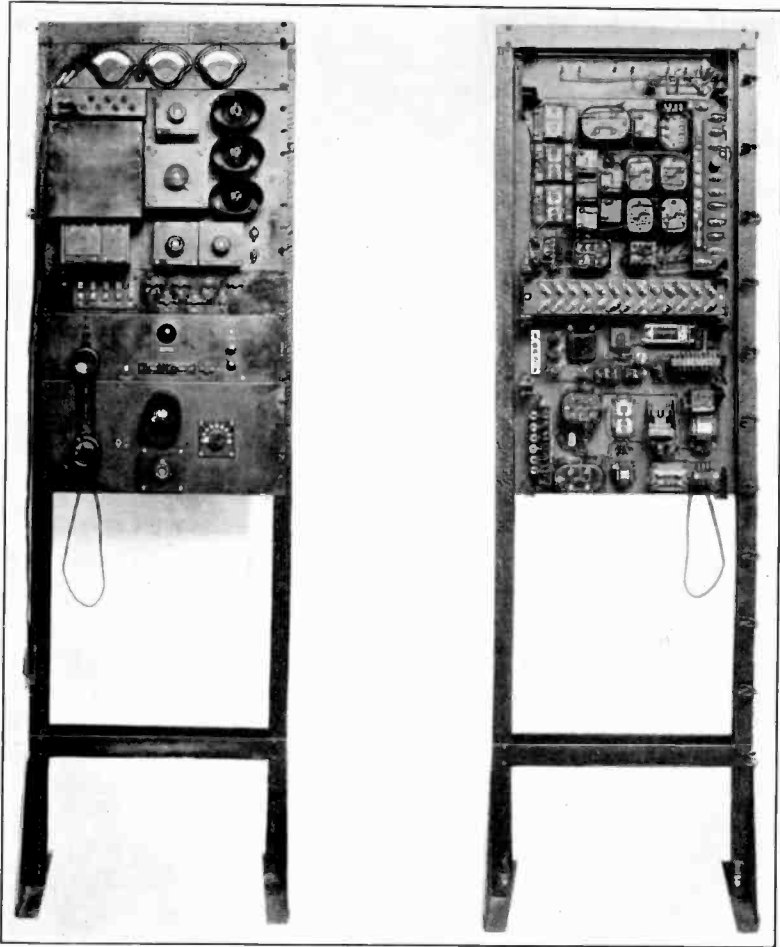
FIGURE 5—1-B Speech Input Equipment

noted, this apparatus consists of five separate units: a meter panel, a three-stage amplifier, a switch and fuse panel, a signal and control panel, and a single stage monitoring amplifier. In the rear view, the metal covers, which normally enclose the individual units, have been removed to enable the apparatus to be seen in more detail.

The circuit of the three-stage amplifier is shown schematically in Figure 6. Vacuum tubes having an amplification constant of
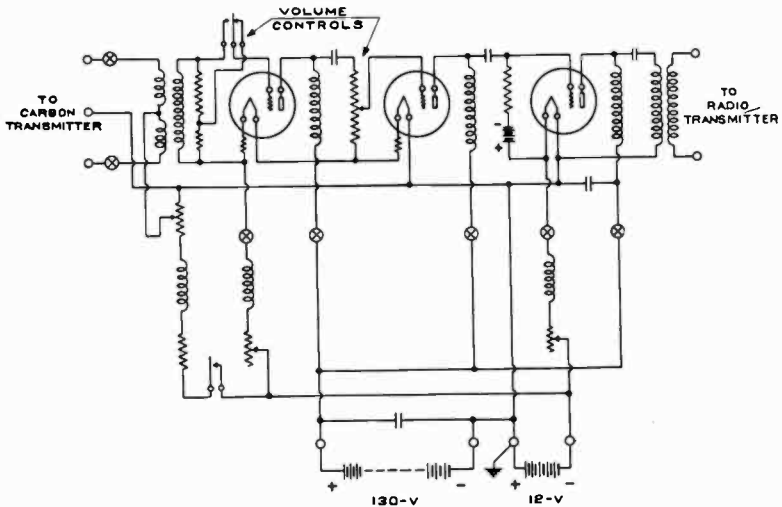
FIGURE 6—Schematic Diagram of Speech Input Amplifier

approximately thirty are employed in the first two stages. The last stage employs a tube having a considerably lower amplification constant but a greater power capacity. Potentiometers in both the first and second stages afford a wide range in amplification control. Power for the filaments is obtained from a suitable 12-volt storage battery. A 130-volt battery consisting of standard Number 6 dry cells supplies power to the plate circuits. Grid voltages for the first two tubes are obtained by utilizing the voltage drop over suitable resistances in the filament circuit. A grid battery of the flashlight type is employed with the third tube. The maximum amplification to be had with this amplifier is 85 miles[4] representing a power ratio of 1.2×10^8. By means of the potentiometers, the amplification can be varied by 3-mile steps from the maximum down to approximately 25 miles.

The signal and control panel is employed in connection with the studio unit, which has already been described. On this panel, in addition to the intercommunicating telephone circuit and the buzzer and lamp signaling facilities, arrangements are made for the control of the radio transmitter, the microphone circuits, and the monitoring circuits. The radio transmitter control feature

[4]Miles gain $= \dfrac{21.13}{2} log_{10} \dfrac{P_2}{P_1}$ where P_1 is the input power, and P_2 the output power.

takes the form of a key which operates a relay in the field circuit of the plate generator. In the microphone circuit, a suitable three-position, multiple-contact key is provided enabling the studio transmitter, the operator's transmitter, or a third instrument (or a wire line) to be connected to the input terminals of the amplifier, as the occasion may require. These two keys are electrically interlocked with the green "RADIO" lamp on the studio unit, so that both keys must be thrown in the proper direction before a signal is given to the announcer. For monitoring purposes, a suitable key is provided to connect the monitoring loud speaker to either a special vacuum tube rectifier which is built into the radio transmitter and coupled to the antenna circuit, or the monitoring amplifier which is connected permanently to the output terminals of the three-stage amplifier. By the operation of this key it is possible, therefore, for the operator to listen either to the outgoing transmission or to the input to the radio transmitter and to keep an accurate check on the operation of the equipment. When the microphone control key is thrown to the operating room position, a red lamp on the panel is lighted and the monitoring loud speaker is disconnected, irrespective of the position of the monitoring key, to prevent "singing" between the operator's transmitter and the loud speaker. A white guard lamp is also provided on the panel to indicate to the operator when the studio control key is thrown to the broadcast position.

The monitoring amplifier is a single stage unit of the usual type but with transformers designed for high quality transmission. A potentiometer is provided for volume control. A suitable loud speaker mounted on the operator's desk is associated with this amplifier for monitoring purposes.

Front and rear views of the radio transmitter are shown in Figure 7 and the circuit employed is represented schematically in Figure 8. This unit is of the oscillator-modulator type employing two 250-watt tubes as oscillators and two similar tubes as modulators. One 50-watt tube is also utilized as a speech amplifier working into the modulator tubes. The oscillator circuit is of the Meissner type, that is, both the grid and plate circuits are inductively coupled to the oscillatory circuit, but it departs from the usual arrangement in that an adjustable condenser is employed in parallel with the plate coil. This variation has been shown to afford a simple and convenient means of adjusting impedance conditions in the circuit so as to bring about efficient operation. As will be noted, the oscillator does not work directly into the antenna, but a closed oscillatory circuit
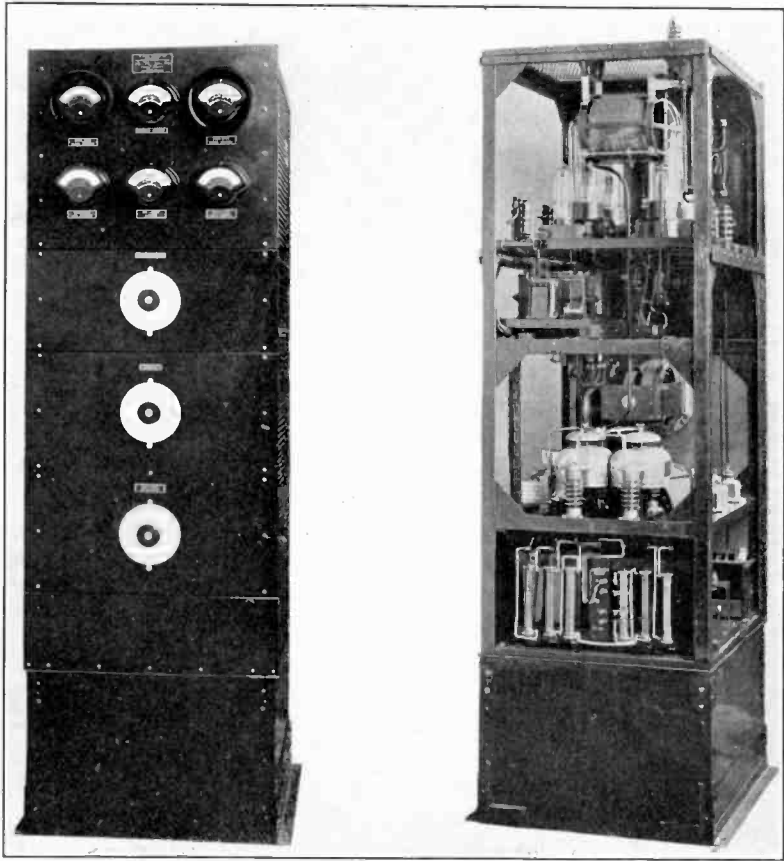
Figure 7—1-B Radio Transmitter

is employed which is coupled to the antenna circuit by means of a mutual capacity. This arrangement has been adopted primarily for the suppression of harmonics. The use of two tuned circuits and capacity coupling affords a high degree of discrimination against these higher frequencies, and radiation is confined almost entirely to the desired frequency band. The use of coupled circuits also results in somewhat better frequency stability than that representative of a transmitter employing the antenna as a part of the oscillator circuit, but dependence in this respect must be placed primarily in a rigid antenna structure.

Another feature of possible interest in connection with the radio frequency circuits is the antenna discharge resistor which appears on the schematic and may be seen in Figure 7 in the upper right-hand corner of the rear view. Experience has indi-

cated that with many antennas, if a series condenser is employed, thus interrupting the direct current path to ground, static charges will accumulate which may build up a voltage sufficient to break down the safety gap of the condenser. A number of stations have reported occasional interruptions arising from trouble of this nature. Observations to date indicate, however, that a relatively high resistance connected as shown is a complete solution to the problem.
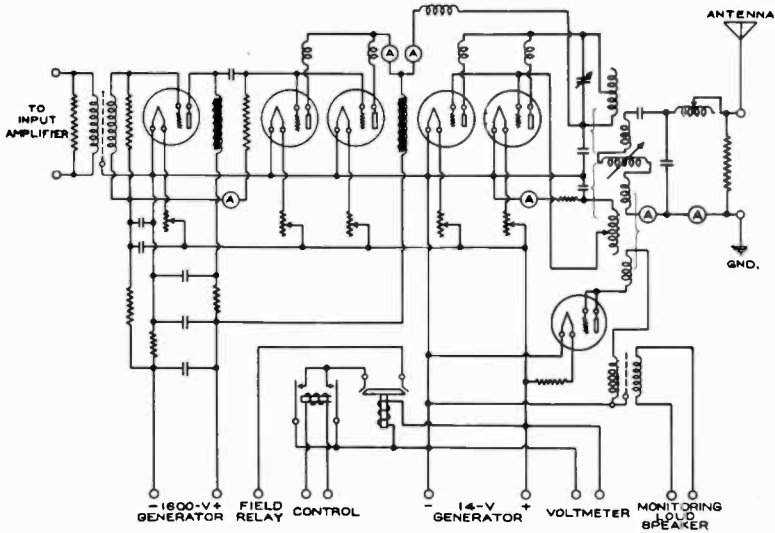
FIGURE 8—Schematic of 1-B Radio Transmitter

The audio frequency portion of the transmitter circuit, embracing the speech amplifier and the modulator, is essentially a two-stage, audio frequency amplifier. The output circuit of the last stage of this amplifier (the modulator) includes the plate circuit of the oscillator. The power supply to the latter may be considered to consist, therefore, of a direct component upon which is superimposed an audio frequency component corresponding to the desired transmission. Under these conditions the output of the oscillator varies in proportion and the envelope of the time-antenna current curve is a faithful reproduction of the signal voltage impressed upon the plate circuit. It is interesting to note in this connection that, within the usual operating limits, the distortion which can be traced to the radio frequency portion of the transmitter is negligible. In other words, by suitable design of the radio frequency circuits, the matter of insuring high quality

transmission may be resolved into proper treatment of the audio frequency circuits, and all of the accumulated experience of the telephone art brought to bear upon the problem.

A somewhat novel feature of the audio frequency circuit is the method employed for obtaining negative grid voltages for the speech amplifier and modulator tubes. A resistance is employed between the negative terminal of the plate generator and the filaments as shown in Figure 8. The grid circuit is connected through a filtering network to such a point on this resistance that the IR drop due to the plate current represents the required negative grid voltage. This method renders a grid battery unnecessary, an important consideration from a maintenance point of view.

The mechanical features of the transmitter are clearly shown in the photographs and require little explanation. The meters and controls are centralized upon the front panel. Meters are provided in the antenna and closed oscillatory circuits, in the oscillator plate and grid circuits, and in the modulator plate and grid circuits. When the equipment is in operation, the needle of the modulator plate meter swings at syllable frequency and affords a convenient, altho somewhat rough, means of adjusting the input amplifier to give the proper volume level. The function of the meter in the grid circuit of the modulator is to serve as an "overload" indicator. Overloading has already been mentioned and its causes briefly discussed. In the equipment as designed, it is occasioned by the peak signal voltage impressed upon the modulator grid exceeding the polarizing voltage in the positive direction. When this occurs, a momentary current will flow from the filament to the grid and the meter will give an indication. The operator is, therefore, afforded a positive means of detecting overloading and a definite operating procedure may be worked out to insure best results.

The three dials shown control the antenna tuning, the frequency, and the oscillator adjustment, respectively. The first is associated with the antenna tuning coil shown in the top of the set behind the tubes. The second operates upon the oscillator coil, which appears near the center of the rear view and serves to vary the inductance of the winding in the oscillatory circuit. The oscillator adjustment control is connected with a variable air condenser joined in parallel with the plate coil as previously described.

The condensers in the radio frequency circuits appear below the oscillator coil. The audio frequency coils and condensers are
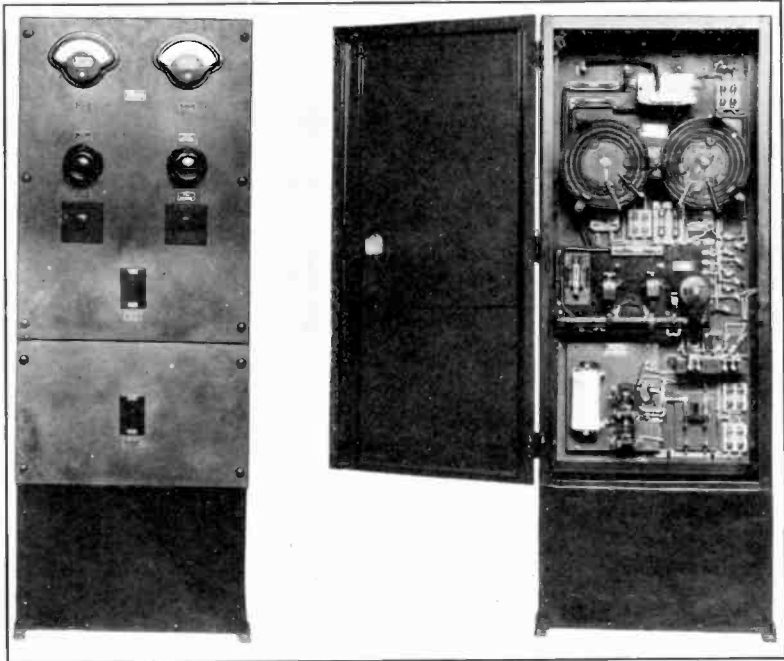
FIGURE 9—Power Panel

mounted under the tube compartment and in the bottom of the unit behind the resistance unit assembly.

The power apparatus for the radio transmitter is shown in Figure 9 and Figure 10. The motor-generator consists of three units mounted on a fabricated base: a suitable motor, a 1,600-volt 1.25-ampere generator, and a 16-volt 30-ampere generator. The high voltage generator is constructed with two 800-volt armature windings and employs two commutators. Current for
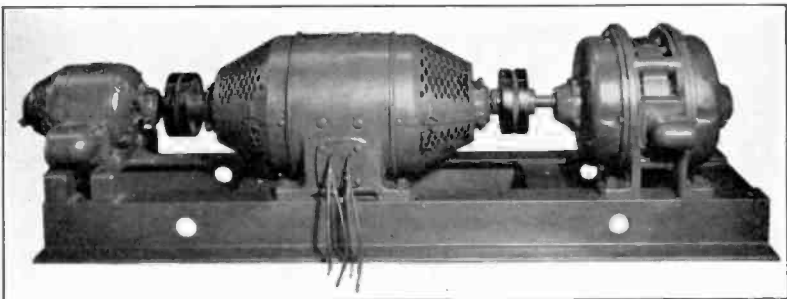
FIGURE 10—Motor Generator

the field is obtained from the low voltage unit, which is self-excited.

The power panel is of the cabinet type. It includes a suitable motor starter which is controlled from a push-button station on the panel and on the operator's desk, field rheostats and volt-meters for each generator, an electromagnetically operated switch in the plate circuit (which also serves as a circuit breaker), and rotary snap switches in the filament and high voltage field circuits. No live metal parts are exposed on the panel and, as further protection to the operator, a door switch is provided which prevents the motor from being started until the door of the cabinet is closed.

The radio receiver is shown in Figure 11. This unit consists
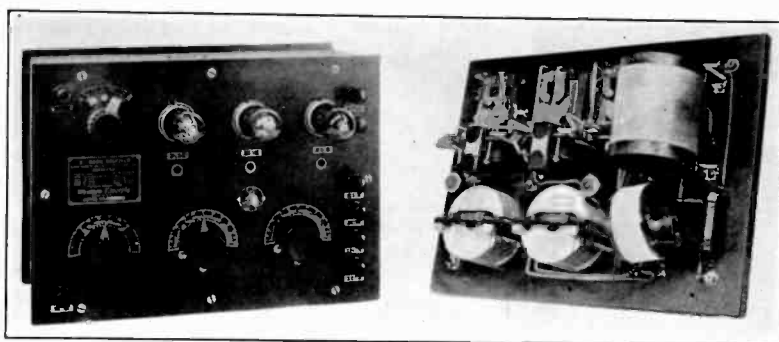
FIGURE 11—2-C Radio Receiver

of a detector and two stages of audio frequency amplification utilizing tubes similar to the VT-1. Coupled radio frequency circuits are employed without regeneration. Suitable jacks are provided, permitting the detector to be used alone or in combination with one or both of the amplifier stages as required. Power for the filament and plate circuits is obtained from the batteries that supply the speech input equipment.

In discussing the individual apparatus units, an attempt has been made to point out in each case the relationship to the other parts of the equipment. The description would hardly be complete, however, without some further indication of the nature of the system as a whole. In Figure 12, therefore, the general arrangement is represented diagrammatically. All of the units are shown and the various connections between them, the broadcasting circuits, the signal and control circuits, and the power circuits, are indicated by the different types of lines. Viewed in

combination with the detailed descriptions previously given, this drawing will enable the general aspects of the system to be readily understood.
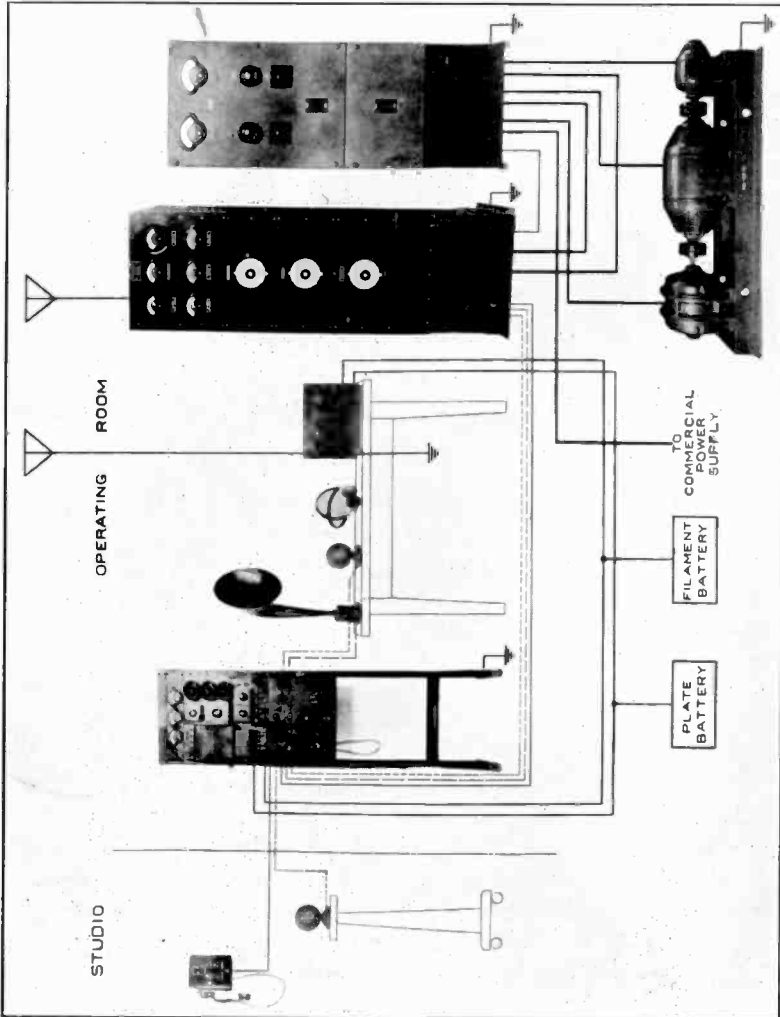
FIGURE 12—101-B Radio Telephone Broadcasting Equipment

On account of the severe transmission requirements imposed upon this class of apparatus, the performance capabilities of the equipment, especially its frequency-transmission characteristic, will probably be of considerable interest. The characteristic for the equipment when employing the carbon transmitter is shown

in Figure 13. This curve is a composite based upon single frequency measurements made on the individual apparatus units and may be considered to represent the relative magnitudes at the various frequencies of the effective alternating component
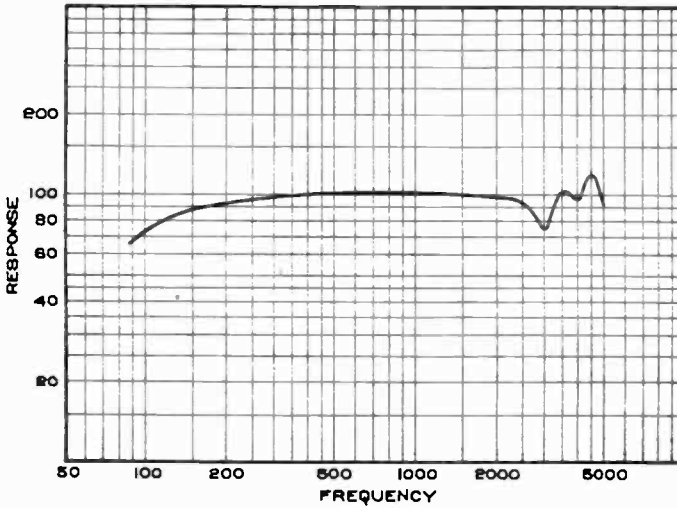
FIGURE 13—Frequency Response Characteristic with Carbon Transmitter

in a suitable rectifier circuit coupled to the antenna when the sound pressure impressed upon the diaphragm of the transmitter is maintained constant. In other words, if the effects of standing waves and other similar acoustical phenomena are eliminated, and if a pure tone of constant intensity but variable pitch is produced in the studio, this curve represents the relative intensity of the sound which might be expected from a distortionless receiving device. Logarithmic scales are employed for both ordinates and abscissas since there is considerable evidence to indicate that the response of the ear to variations in both intensity and pitch follows a logarithmic law. As will be noted, substantially uniform transmission efficiency is indicated except at the extreme ends of the frequency range where some slight deficiencies become evident. These can only be detected by careful measurement, however, and it is doubtful if their effect upon transmitted music would be apparent even to the artistically trained ear.

CONCLUSION

The general considerations involved in the design of a system for the high quality transmission of speech and music have been men-

tioned briefly, and the requirements to be met by the various apparatus units in a radio telephone broadcasting equipment have been discussed in some detail. A commercial 500-watt transmitting equipment has been described and its performance capabilities illustrated. It has been shown that apparatus can be produced commercially which will afford transmission at substantially uniform efficiency throughout the entire frequency range commonly employed in music.

From a quality point of view, at least, therefore, transmitting equipment for broadcasting applications may be said to have already reached an advanced stage of development. A standard of performance has been attained which it is not believed to be presumptuous to claim is an artistic, as well as a technical success. The next few years will, no doubt, bring forth contributions of great value, new and more efficient telephone transmitters, better vacuum tubes, more stable circuits, and less expensive apparatus, but in view of present attainments, radical improvement in the quality of the transmission can hardly be expected. If further improvement in quality is desired by the listener, therefore, it must be obtained elsewhere than in the transmitting equipment.

New York, N. Y.
January 16, 1924.

SUMMARY: The general transmission considerations applying to any system for the high quality transmission of speech or music are outlined briefly, and the specific requirements to be met by the various apparatus units in a radio broadcasting equipment are discussed in some detail. The standard Western Electric 500-watt broadcasting equipment, which has found application in some fifty of the larger stations in this country and abroad, is described. Its performance capabilities are illustrated and it is indicated that a standard of performance has been attained which renders possible reproductions not substantially different from the original.

AN ANALYSIS OF TWO TRIODE CIRCUITS*

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There seems to have been but little exact analysis of triode circuits, with the necessary experimental verification, without which the theoretical work is not complete. It is with the idea of helping to fill this gap that the following analyses and experiments are reported.

The plate current of a triode can be written

$$I_p = a \, (E_p + \mu \, E_g)^x$$

in which the symbols have their usual meaning. The exponent x is different in different types of tubes, and varies greatly in a given tube as the voltages used in the grid and plate are changed from small to large values. For the tube used in the following tests the curve represented by the above equation is given in Figure 1. The grid was held at zero potential and plate voltage changed as indicated.

The input circuit of a tube, grid-to-filament, has both conductance and capacitance. The former may be made negligible by negative bias and the latter is negligible at low frequencies. Such was the case in our tests.

The output circuit, plate-to-filament, also has resistance and capacitance; the latter was negligible in our tests, but of course the plate-filament resistance can seldom be neglected, and it was not, in our analyses and experiments. The resistance of the plate-filament circuit of the tubes has three different values, depending upon how the resistance is defined.

(a) Continuous current resistance, R_o. From the plate current equation $R_o = \dfrac{E_p}{I_p} = \dfrac{1}{a \, (E_p + \mu \, E_g)^{x-1}}$; or, if E_g zero,

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$$R_o = \frac{1}{a\,E_p{}^{x-1}} \qquad (1)$$

(b) Alternating current resistance, plate voltage alone varying.

$$R_p = \frac{\Delta E_p}{\Delta I_p} = \frac{\partial E_p}{\partial I_p} = \frac{1}{a\,x(E_p+\mu\,E_g)^{x-1}} = \frac{R_o}{x} \qquad (2)$$

STATIC CHARACTERISTIC FOR V.T.2 TUBE

I_f A.2 AMPS E_g 0
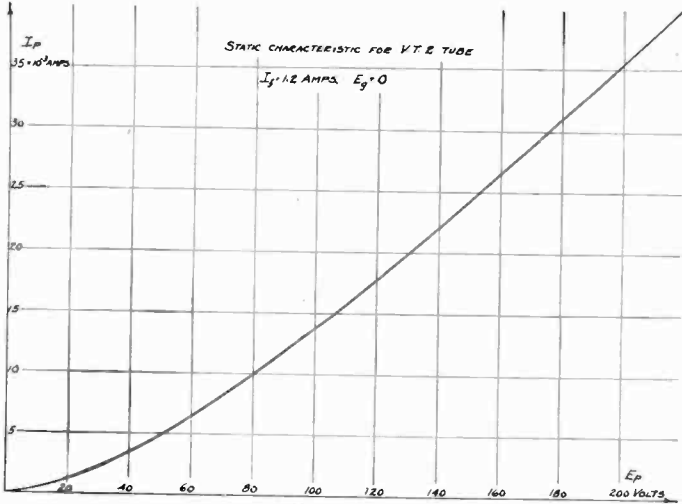
FIGURE 1

This resistance is constant only for small values of ΔE_p. For larger increments, the resistance must be written as a constant plus a series of harmonic terms. The value of the constant term increases with amplitude of ΔE_p, approaching R_o as a limit.

(c) Alternating current resistance, plate and grid voltages both varying.

$$d I_p = a\,x(d E_p+\mu\,d E_g)\,(E_p+\mu\,E_g)^{x-1}$$

$$R_p' = \frac{d E_p}{d I_p} = \frac{1}{a\,x(E_p+\mu\,E_g)^{x-1}} - \mu\frac{d E_g}{d I_p}$$

$$= R_p - \mu\frac{d E_g}{d I_p} \qquad (3)$$

Generally, E_g is varied by variations of I_p and it may be that the value of $\frac{d E_g}{d I_p}$ is either positive or negative, according to the manner in which E_g is made to change by changes in I_p. In case $\frac{d E_g}{d I_p}$ is positive, and sufficiently large, R_p' may be negative, that

is, althó the change in plate voltage $d\,E_p$ is positive, the accompanying change in plate current is negative.

In the following analyses, by the term "plate resistance," the value of R_p (case b) is always intended. This value is always positive and, in magnitude, is somewhat less than the "B" battery voltage divided by the steady value (average value) of the current furnished to the tube by this battery.

Oscillatory Circuit Connected to Plate

The first circuit to be considered is that given in Figure 2.
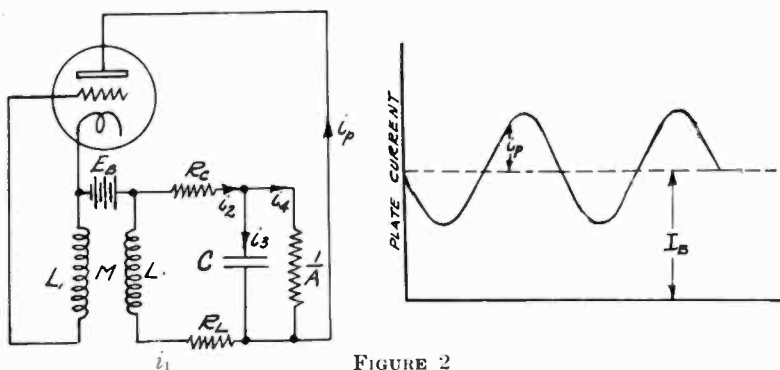
Figure 2

A small increment in plate voltage or grid voltage will produce currents in the circuit as indicated. It is supposed that the condenser has leakage, $\dfrac{1}{A}$, (conductance $= A$), and that the resistance of the condenser branch is R_c. For a good condenser, both of these may be neglected, but if the condenser happens to be an antenna, such is not possible. In the following equations e_p is the instantaneous value of the alternating component of the plate voltage, e_g is the instantaneous value of the alternating grid voltage, i_p is the instantaneous value of the alternating component of the plate current. In the analysis to follow, we assume the grid takes a negligible current; at low frequencies the charging current is negligible and by suitable bias the conduction current may also be made negligible. Then

$$i_p = \frac{e_p + \mu e_g}{R_p} \text{ or } i_p\,R_p = e_p + \mu e_g \tag{4}$$

$$e_g = -M\frac{d\,i_1}{d\,t} \text{ (This defines polarity of } M) \tag{5}$$

$$i_p = i_1 + i_2 \tag{6}$$

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$$e_p = -i_1 R_L - L\frac{d\,i_1}{d\,t} = -i_2\,R_c - \frac{i_4}{A} = -i_2\,R_c - \frac{q}{C} \qquad (7)$$

in which q is the instantaneous value of the charge on condenser C

But $\qquad \frac{d\,q}{d\,t} = i_3$ and $i_2 = i_3 + i_4$. So from (7)

$$\frac{1}{A}\frac{d\,i_4}{d\,t} = \frac{i_3}{C} \qquad (8)$$

And $\qquad i_2 = \frac{C}{A}\frac{d\,i_4}{d\,t} + i_4 \qquad (9)$

From 4, 5, 6, and 7,

$$R_p(i_1 + i_2) = -i_1 R_L - L\frac{d\,i_1}{d\,t} - \mu\,M\frac{d\,i_1}{d\,t} \qquad (10)$$

From which $\qquad i_2 = -i_1\left(1 + \frac{R_L}{R_p}\right) - \left(\frac{L + \mu\,M}{R_p}\right)\frac{d\,i_1}{d\,t} \qquad (11)$

From (9) and (11),

$$i_4 + \frac{C}{A}\frac{d\,i_4}{d\,t} = -i_1\left(1 + \frac{R_L}{R_p}\right) - \frac{L + \mu\,M}{R_p}\frac{d\,i_1}{d\,t} \qquad (12)$$

Using the second and third terms of (7), and substituting the value of i_2 from (9)

$$i_1 R_L + L\frac{d\,i_1}{d\,t} = i_4\left(R_c + \frac{1}{A}\right) + R_c\frac{C}{A}\frac{d\,i_4}{d\,t} \qquad (13)$$

Multiply (12) by R_c and use the value of $i_4\,R_c$ so obtained in equation (13)

$$\frac{i_4}{A} = i_1\left(R_L + R_c + \frac{R_c R_L}{R_p}\right) + \left[L + \frac{R_c}{R_p}(L + \mu\,M)\right]\frac{d\,i_1}{d\,t} \qquad (14)$$

Divide (12) by A, substitute for $\frac{i_4}{A}$ its value from (14)

$$\frac{C}{A^2}\frac{d\,i_4}{d\,t} = -i_1\left[R_L + \left(1 + \frac{R_L}{R_p}\right)\left(R_c + \frac{1}{A}\right)\right] - $$
$$\frac{d\,i_1}{d\,t}\left[L + \left(R_c + \frac{1}{A}\right)\frac{L + \mu\,M}{R_p}\right] \qquad (15)$$

Derive (14), multiply by $\frac{C}{A}$, and combine with (15).

$$\frac{d^2\,i_1}{d\,t^2}C[L\,R_p + R_c(L + \mu\,M)] + \frac{d\,i_1}{d\,t}[C(R_L R_p + R_c R_p + R_L R_c) +$$
$$L\,A\,R_p + (A\,R_c + 1)(L + \mu\,M)] + i_1[R_L\,R_p\,A +$$
$$(R_p + R_L)(A\,R_c + 1)] = 0 \qquad (16)$$

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$$\frac{d^2 i_1}{dt^2} + \frac{d i_1}{dt} \frac{(C R_L + A L)(R_p + R_c) + C R_c R_p + L + \mu M(A R_c + 1)}{C[L(R_p + R_c) + R_c \mu M]} +$$
$$i_1 \frac{A R_L(R_p + R_c) + A R_p R_c + R_p + R_L}{C[L(R_p + R_c) + R_c \mu M]} = 0 \qquad (17)$$

or
$$\frac{d i_1^2}{dt^2} + \frac{d i_1}{dt} \frac{a}{N} + i_1 \frac{b}{N} = 0 \qquad (18)$$

in which
$$a = (C R_L + A L)(R_p + R_c) + C R_c R_p + L + \mu M(A R_c + 1)$$
$$b = A R_L(R_p + R_c) + A R_p R_c + R_p + R_L$$
$$N = C[L(R_p + R_c) + R_c \mu M]$$

In case the conductance of the condenser is zero and no resistance is used in the capacity branch ($A = 0$, $R_c = 0$).

$$a = C R_L R_p + L + \mu M$$
$$b = R_p + R_L$$
$$N = C L R_p$$

The solution of (18) will be a sinusoidal, damped, current if $\frac{a^2}{4 N^2} < \frac{b}{N}$. If $\frac{a}{2 N}$ is zero any oscillations started keep up indefinitely. For oscillations to build up $\frac{a}{2 N}$ must be negative, which means that M must be negative, and in absolute value

$$M > \frac{1}{\mu(A R_c + 1)}[L + C R_c R_p + (C R_L + A L)(R_p + R_c)] \qquad (19)$$

For the perfect condenser, with $A = 0$ and $R_c = 0$, oscillations will occur if M is negative and satisfies the condition for absolute magnitude,

$$M > \frac{1}{\mu}(L + C R_L R_p)$$

The frequency of the oscillations, if they do occur, is given from equation (18) by the relation

$$(2 \pi f)^2 = \omega^2 = \frac{b}{N} - \frac{a^2}{4 N^2} \qquad (20)$$

or for the perfect condenser,

$$f = \frac{1}{4 \pi L} \sqrt{\frac{4 L}{C}\left(1 + \frac{R_L}{R_p}\right) - \left[R_L + \frac{1}{C R_p}(L + \mu M)\right]^2} \qquad (21)$$

The foregoing theory may be tested by opening the oscillatory circuit and putting it in a Wheatstone bridge, as shown in

Figure 3.   The voltage introduced for measuring, e_L, will be a sine wave, equal in magnitude to the drop in one arm of the bridge.
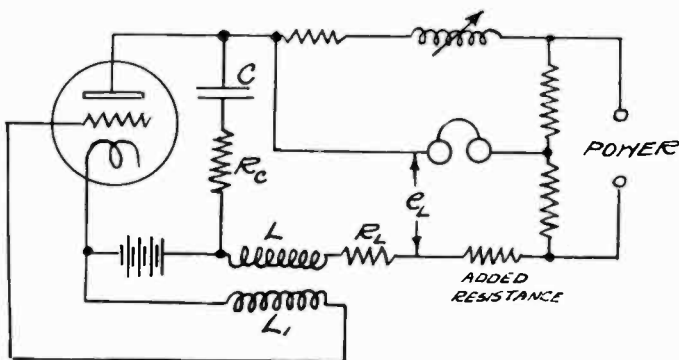

FIGURE 3

The derivation of the equation for this circuit is much the same as before, having, of course, the extra voltage, e_L, to take care of, however.

The equation, similar in form to (18), comes out

$$\frac{d^2 i_1}{d t^2}+\frac{d i_1}{d t}\frac{a}{N}+i_1\frac{b}{N}=\frac{1}{N}\left[\frac{d e_L}{dt}C(R_p+R_c)+e_L(R_pA+R_cA+1)\right](22)$$

Putting $e_L=E_L\varepsilon^{j\omega t}$, we know the current will be sinusoidal and representable by

$$i_1=F_1\varepsilon^{j\omega t}+j\,F_2\,\varepsilon^{j\omega t}$$

Getting the proper derivatives, substituting in equation (22) and equating real and imaginery terms, we find

$$F_1=E_L\frac{(b-N\,\omega^2)(R_p\,A+R_c\,A+1)+a\,\omega^2\,C(R_p+R_c)}{(b-N\,\omega^2)^2+(a\,\omega)^2}\qquad(23)$$

$$F_2=E_L\frac{\omega\,C(R_p+R_c)(b-N\,\omega^2)-a\,\omega(R_p\,A+R_c\,A+1)}{(b-N\,\omega^2)^2+(a\,\omega)^2}\qquad(24)$$

The impedance of the circuit Z_L is given by

$$Z_L=\frac{E_L}{I_1}=\frac{E_LF_1}{F_1{}^2+F_2{}^2}+j\frac{E_LF_2}{F_1{}^2+F_2{}^2}R=L'+j\omega L_L'\qquad(25)$$

So $\qquad R_L'=\dfrac{(b-N\,\omega^2)(R_p\,A+R_c\,A+1)+a\,\omega^2\,C(R_p+R_c)}{(R_p\,A+R_c\,A+1)^2+\omega^2\,C^2(R_p+R_c)^2}\qquad(26)$

$$L_L'=\frac{a(R_p\,A+R_c\,A+1)-C(R_p+R_c)(b-N\,\omega^2)}{(R_p\,A+R_c\,A+1)^2+\omega^2\,C^2(R_p+R_c)^2}\qquad(27)$$

In case $R_c = 0$ (which was true in our experiments)

$$R_L' = \frac{R_L(R_pA+1)^2 + R_p(R_pA+1) + \omega^2 C R_p[L(A R_p+1) + C R_L R_p + \mu M - L(A R+1)]}{(R_pA+1)^2 + \omega^2 C^2 R_p{}^2}$$

$$= R_L + \frac{R_p(R_pA+1) + \omega^2 C R_p\mu M}{(R_pA+1)^2 + \omega^2 C^2 R_p{}^2} \tag{28}$$

Putting in an equivalent series resistance, r, in place of A we have $\omega^2 C^2 r = A$ and the denominator of equation (28) becomes $R_p{}^2 \omega^2 C^2(1 + \omega^2 C^2 r) + 2 R_p r \omega^2 C^2 + 1$.

In the measurements we shall record we have, approximately, $R_p = 5,000$. $\omega = 8,000$, $C_{max} = 10^{-6}$, $r_{max.} = 2$, $C_{min.} = 0.2 \times 10^{-6}$ So $\omega^2 C^2_{max.} r_{max.} \ll 1$, $2 r_{max.} \ll R_p$, $R_p{}^2 \omega^2 C^2_{min.} \gg 1$, so the denominator of equation (28) becomes approximately $R_p{}^2 \omega^2 C^2$.

Then $\quad\quad R_L' = R_L + r + \dfrac{\mu M}{C R_p} + \dfrac{1}{\omega^2 C^2 R_p} \tag{29}$

By a similar transformation we get, from equation (27),

$$L_L' = L - \frac{1}{\omega^2 C} + \left(\frac{r}{R_p} + \frac{1}{\omega^2 C^2 R_p{}^2}\right)\mu M$$

For resonance $L_L' = 0$ and we get from the above equation

$$\omega^2 = \frac{C R_p{}^2 - \mu M}{C^2 R_p(L R_p + r \mu M)} = \frac{1 - \dfrac{\mu M}{C R_p{}^2}}{L C + C \dfrac{r}{R_p}\mu M}$$

$$= \frac{1}{L C}\left[1 - \frac{\mu M}{R_p}\left(\frac{1}{R_p C} + \frac{r}{L}\right)\right] \tag{30}$$

Above resonance, the circuit becomes capacitive and so a condenser must be used in the bridge to get a balance. This would open up the bridge circuit which has permitted the flow of continuous current, so we have to re-arrange the test circuit in accordance with Figure 4.

By derivations similar to those used for the circuit of Figure 3 we get, using R_c' to indicate the resistance as measured in the circuit of Figure 4.

$$R_c' = R_c + r + \frac{R_p R_L(R_p + R_L) + \omega^2 L R_p(L + \mu M)}{(R_p + R_L)^2 + \omega^2(L + \mu M)^2} \tag{31}$$

And as $R_L \ll R_p$ we may put

$$R_c' = R_c + r + \frac{R_p{}^2 R_L + \omega^2 L R_p(L + \mu M)}{R_p{}^2 + \omega^2(L + \mu M)^2} \tag{32}$$
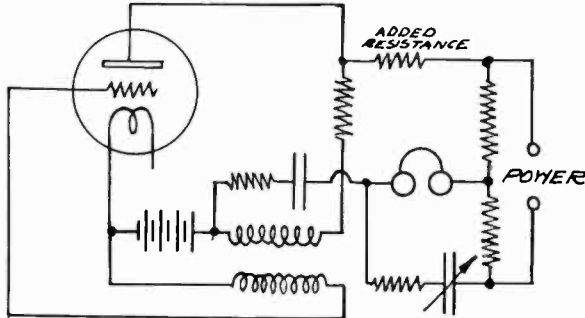
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FIGURE 4

This resistance, it is to be noted, is independent of the amount of capacity used in the circuit.

Resonant frequency is given by

$$\omega^2 = \frac{1}{LC}\left(1 + \frac{R_L}{R_p}\frac{\mu M}{L} + \frac{(L+\mu M)^2}{LCR_p^2}\right) \tag{33}$$

and the resistance as resonance is (if we put $R_c = 0$)

$$R_c'(\text{resonance}) = R_L + r + \frac{L+\mu M}{CR_p} \tag{34}$$

For measurement in the inductive branch the resistance at resonance becomes, after legitimate approximation in equation (29)

$$R_L'(\text{resonance}) = R_L + r + \frac{L+\mu M}{CR_p} \tag{35}$$

which is just the same as it is in the capacitive branch. It is to be noticed however, from equations (30) and (33), that the resonant frequency is higher when measuring in the capacitive branch than when measuring in the inductive branch.

VARIATION OF RESISTANCE WITH FREQUENCY, FOR RESONANT CONDITION,

At resonant frequency

$$R' = R + \frac{L+\mu M}{CR_p}\ (\text{by putting } R = R_L + r + R_c)$$

If we now make the ratio of inductances of the two coils (in Figures 3 and 4) constant, so that $L_1 = m^2 L$, if we keep the ratio of L to C constant so that $C = nL$, and as $M = K\sqrt{LL_1}$ we have, at resonance

$$R'(\text{resonance}) = R + \frac{L+\mu KmL}{nLR_p} = R + \frac{1+\mu Km}{nR_p} \tag{36}$$

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It thus seems that for a given tube the resistance of the oscillatory circuit will be independent of frequency, provided the ratio of L to C is kept constant, and a constant coupling is maintained between the coils, which are always selected with a definite ratio of inductances.

Thus suppose a certain combination gives -30 ohms resistance at $\omega = 16,000$. We change the size of wire used in the coils so as to get 4 times as many turns in the same size coil and keep the coils in the same relative position. The inductances will be increased 16 times and the coefficient of coupling will be the same as before. We increase the capacity 16 times. We drop the frequency so that $\omega = 1,000$ and the circuit will again have -30 ohms resistance.

Oscillatory Circuit Connected to Grid

The next circuit to investigate has the oscillatory circuit connected to the grid, as in Figure 5. Altho this circuit looks much like that previously used, the solution is somewhat different. We again assume that the grid current is sufficiently small to be neglected without appreciable error.

As preliminary equations we have

$$R_p i_p = e_p + \mu e_g$$

$$e_p = -M\frac{d i}{d t} - L_1\frac{d i_p}{d t}$$

$$e_g = -R_L i - L\frac{d i}{d t} - M\frac{d i_p}{d t}$$

$$= R_c i + \frac{i_2}{A} = R_c i + \frac{q}{C}$$

Now $\qquad \dfrac{d q}{d t} = i_1$, so $\dfrac{1}{A}\dfrac{d i_2}{d t} = \dfrac{i_1}{C}$ and $i = i_1 + i_2$.

From these equations we get by combination and differentiation

$$\frac{d i_p}{d t} = -\frac{1}{L_1}\left(e_p + M\frac{d i}{d t}\right) = -\frac{1}{M}\left(e_g + L\frac{d i}{d t} + R_L i\right)$$
$$= \frac{1}{R_p}\left(\frac{d e_p}{d t} + \mu\frac{d e_g}{d t}\right) \tag{37}$$

Using the second and third members of (37)

$$e_p = \frac{L_1}{M}\left(e_g + L\frac{d i}{d t} + R_L i\right) - M\frac{d i}{d t} \tag{38}$$

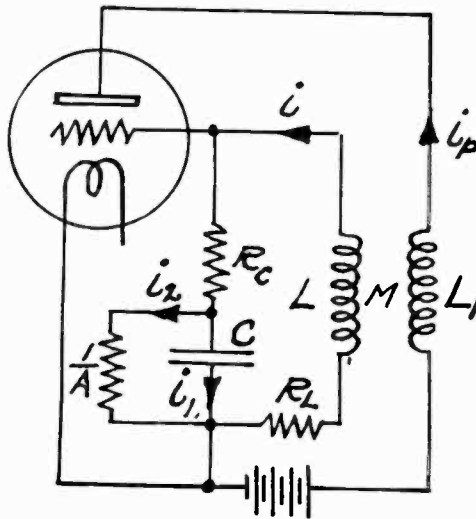Differentiating (38) and using this in the third and fourth members of (37)

FIGURE 5

$$-\frac{1}{M}\left(e_g+L\frac{d\,i}{d\,t}+R_L i\right)=\frac{1}{R_p}\left[\frac{L_1}{M}\left(\frac{d\,e_g}{d\,t}+L\frac{d^2 i}{d\,t^2}+R_L\frac{d\,i}{d\,t}\right)\right.$$
$$\left.-M\frac{d^2 i}{d\,t^2}+\mu\frac{d\,e_g}{d\,t}\right] \tag{39}$$

Combining terms

$$\frac{d^2 i}{d\,t^2}(L\,L_1-M^2)+\frac{d\,i}{d\,t}(L_1 R_L+R_p L)+i\,R_p R_L=-R_p e_g$$
$$-\frac{d\,e_g}{d\,t}(L_1+\mu M) \tag{40}$$

Now $\qquad e_g=R_c i+\dfrac{i_2}{A}=i\left(R_c+\dfrac{1}{A}\right)-\dfrac{i_1}{A}=R_c i+\dfrac{q}{C}$ $\qquad$ so

$$\frac{d\,e_g}{d\,t}=R_c\frac{d\,i}{d\,t}+\frac{i_1}{C} \qquad\qquad \text{hence}$$

$$\frac{1}{C}e_g+\frac{1}{A}\frac{d\,e_g}{d\,t}=\frac{i}{C}\left(R_c+\frac{1}{A}\right)+\frac{R_c}{A}\frac{d\,i}{d\,t} \tag{41}$$

By solving for e_g and $\dfrac{d\,e_g}{d\,t}$ in equation (40), we get

$$\frac{d^2 i}{d\,t^2}A(L\,L_1-M^2)+\frac{d\,i}{d\,t}(A\,L_1 R_L+A\,R_p L+C\,R_c R_p)$$
$$+i(A\,R_p R_L+A\,R_p R_c+R_p)=(C\,R_p-A(L_1+\mu\,M))\frac{d\,e_g}{d\,t} \tag{42}$$

and $\qquad \dfrac{d^2 i}{d\,t^2}(C(L\,L_1-M^2)+\dfrac{d\,i}{d\,t}(C\,L_1 R_L+C\,R_p L+C\,R_c L_1$

$$+C\,R_c\mu\,M)+i\{C\,R_p R_L+(A\,R+1)(L_1+\mu M)\}$$
$$=-\{C\,R_p-A(L_1+\mu M)\}e_g \tag{43}$$

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Differentiating (43) and adding the resultant equation to (42), we finally get

$$\frac{d^3 i}{d t^3} C(L L_1 - M^2) + \frac{d^2 i}{d t^2}[A(L L_1 - M^2) + C(R_L L_1 + R_p L)$$
$$+ C R_c(L_1 + \mu M)] + \frac{d i}{d t}[A(L_1 R_L + L R_p) + (A R_c + 1) \quad (44)$$
$$(L_1 + \mu M) + C R_p(R_c + R_L)] + i[A R_p(R_c + R_L) + R_p] = 0$$

Or
$$\frac{d^3 i}{d t^3} + \frac{d^2 i}{d t^2}\frac{a}{N} + \frac{d i}{d t}\frac{b}{N} + i\frac{c}{N} = 0 \quad (45)$$
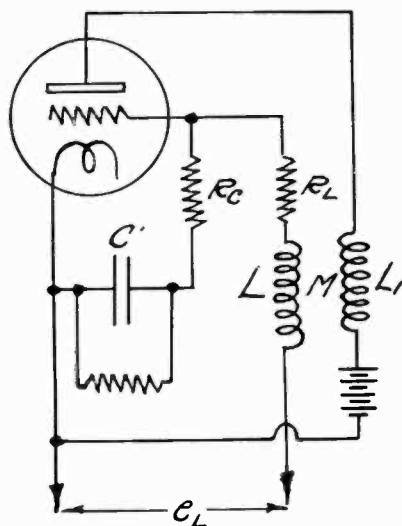
in which

$$a = A(L L_1 - M^2) + C(R_L L_1 + R_p L) + C R_c(L_1 + \mu M)$$
$$b = A(R_L L_1 + R_p L) + (A R_c + 1)(L_1 + \mu M) + C R_p(R_L + R_c)$$
$$c = A R_p(R_L + R_c) + R_p$$
$$N = C(L L_1 - M^2)$$

To test the theory of this circuit we imagine the circuit opened for measurement in a Wheatstone bridge as indicated in

To BRIDGE

FIGURE 6

Figure 6. An analysis carried out to find the impedance to be measured in the bridge yields the equation

$$\frac{d^3 i}{d t^3} + \frac{d^2 i}{d t^2}\frac{a}{N} + \frac{d i}{d t}\frac{b}{N} + i\frac{c}{N} = \frac{d^2 e_L}{d t^2}\frac{C L_1}{N} + \frac{d e_L}{d t}\frac{C R_p + A L_1}{N} + e_L\frac{A R_p}{N}$$

in which a, b, c, and N have the same meaning as before. Writing $e_L = E_L \epsilon^{j\omega t}$ as before and solving for the two components of current F_1 and F_2

$$F_1 = E_L \frac{(C\,R_p + A\,L_1)\omega^2(b - \omega^2\,N) + (A\,R_p - \omega^2\,C\,L_1)(c - \omega^2\,a)}{(c - \omega^2\,a) + \omega^2(b - \omega^2\,N)^2}$$

$$F_2 = E_L \frac{\omega(C\,R_p + A\,L_1)(c - \omega^2\,a) - (A\,R_p - \omega^2\,C\,L_1)\omega(b - \omega^2\,N)}{(c - \omega^2\,a) + \omega^2(b - \omega^2\,N)^2}$$

$$Z = \frac{E_L}{I} = \frac{E_L F_1}{F_1{}^2 + F_2{}^2} + j\frac{E_L F_2}{F_1{}^2 + F_2{}^2} = R_L' + j\,\omega\,L'$$

from which

$$R_L' = \frac{\omega^2(C\,R_p + A\,L_1)(b - \omega^2\,N) + (A\,R_p - \omega^2\,C\,L_1)(c - \omega^2\,a)}{\omega^2(C\,R_p + A\,L_1)^2 + (A\,R_p - \omega^2\,C\,L_1)^2} \quad (46)$$

$$L_L' = \frac{(A\,R_p - \omega^2\,C\,L_1)(b - \omega^2\,N) - (C\,R_p + A\,L_1)(c - \omega^2\,a)}{\omega^2(C\,R_p + A\,L_1)^2 + (A\,R_p - \omega^2\,C\,L_1)^2} \quad (47)$$

By substituting the values of a, b, c, and N, and making the same simplifying assumption as before ($A \ll \omega C$ and $R_c = 0$). we get

$$R_L' = R_L + r + \frac{R_p}{R_p{}^2 + \overline{\omega\,L_1}{}^2}\left(\overline{\omega M^2} + \frac{\mu\,M}{C}\right) \quad (48)$$

and

$$\omega\,L_L' = \left(\omega\,L - \frac{1}{\omega\,C}\right) - \frac{\omega\,L_1}{R_p{}^2 + \overline{\omega\,L_1}{}^2}\left(\overline{\omega\,M^2} + \frac{\mu\,M}{C}\right) \quad (49)$$

If the circuit is opened in the capacity branch, the resistance R_c' will be found independent of the value of the capacity, as it was for the first circuit analyzed (equation (32)).

The resonant frequency is given by the relation

$$\omega^4\,L_1\,C(L\,L_1 - M^2) - \omega^2(L_1{}^2 + \mu\,M\,L_1 - R_p{}^2\,C\,L) - R_p{}^2 = 0 \quad (50)$$

If M is small, the resonant frequency is given approximately by the relation $\omega_o{}^2 = \dfrac{1}{L\,C}$.

Calling the resonant frequency (or more accurately $2\pi \times$ the frequency), ω, and putting $\dfrac{\omega_o{}^2 - \omega^2}{\omega^2} = x$ we get

$$x^2 + x\left(1 + \frac{\omega_o{}^2\,L_1}{R_p{}^2}(L_1 + \mu\,M)\right) + \frac{\omega_o{}^2\,L_1}{R_p{}^2}\left(\frac{M^2}{L} + \mu\,M\right) = 0 \quad (51)$$

Then using the general equations (48) and (49) (after putting $r = 0$), we find from equation (49) the value of

$$\frac{1}{R_p{}^2 + \overline{\omega\,L_1}{}^2}\left(\overline{\omega\,M^2} + \frac{\mu\,M}{C}\right)$$

and use this value in (48) and so get as the resistance of the circuit at resonance (for any value of M)

$$R_L' (\text{resonance}) = R_L + \frac{R_p}{\omega L_1}\left(\omega L - \frac{1}{\omega C}\right)$$

$$= R_L - R_p \frac{L}{L_1} x \tag{52}$$

Can M Be Too Large to Produce Oscillations?

We determine the values of L_1 for which R_L' (resonance) is zero with the maximum possible value of M. We have

$$M = -K\sqrt{L L_1}$$

and using this in equations (51) and (52) we get (for R_L' (resonance) $= 0$).

$$L_1{}^2 - L_1{}^{3/2} K \mu L^{1/2} + L_1\left(\frac{R_p R_L}{\omega_o{}^2 L} + \frac{K^2 L R_p}{R_L}\right) \\ - L_1{}^{1/2}\frac{R_p}{R_L} K\mu L^{3/2} + \frac{R_p{}^2}{\omega_o{}^2} = 0 \tag{53}$$

In our tests we had $\mu = 6.75$, $L = 0.03$, $\omega_o{}^2 = 10^8$, $R_p = 5,000$, $R_L = 3$, and maximum available $K = 0.8$.

So $\quad L_1{}^2 - 0.935 L_1{}^{3/2} + 32 L_1 - 46.8 L_1{}^{1/2} + 0.25 = 0$
which gives $\sqrt{L_1} = 1.425$ or $L_1 = 2.03$ henrys. So if we should put more than 2.03 henrys in the tickler coil, with $K = 0.8$, the set would not oscillate.

In a somewhat simpler derivation we can write (as approximately true) equation (53) in the form

$$K^2 \frac{R_p}{R_L} L L_1 = K \mu \frac{R_p}{R_L} L_1{}^{1/2} L^{3/2}$$

which gives

$$L_1 = \frac{\mu^2}{K^2} L$$

For $K = 0.8$ and $\mu = 6.75$ this gives the upper limit of L_1 as 70 times L, or if $L = 0.03$ the maximum value of L_1 which can be used is 2.1 henrys, which is practically the same as obtained by solution of the more complex equation.

In general $\overline{\omega L_1{}^2} \ll R_p{}^2$ so that equation (48) may be written

$$R_L' = R_L + \frac{1}{R_p}\left(\overline{\omega M^2} + \frac{\mu M}{C}\right) \tag{54}$$

As M is negative the effect of this term $\dfrac{\mu M}{C}$ is to give negative

resistance and the effect of $\overline{\omega M^2}$ is to give positive resistance. Hence at some value of M the two terms will just neutralize and the effect of the triode on the resistance of the circuit is zero. The value of M at which this equality occurs will evidently increase as C decreases. It is given by

$$\omega^2 M = \frac{\mu}{C} \quad \text{or} \quad M = \frac{\mu}{\omega^2 C} \tag{55}$$

VARIATION OF RESONANT FREQUENCY AS M VARIES, M BEING LARGE

If we consider $L_1 = 2$ henrys and $M = 0.2$ and the other constants same as given above, the resonant frequency will vary appreciably as M is varied. We have to solve equation (51) to get the ratio of the resonant frequency to that given by $\omega_o{}^2 = \dfrac{1}{L\,C}$ This effect will depend upon the frequency and was calculated for various values of M, and three frequencies; the results are
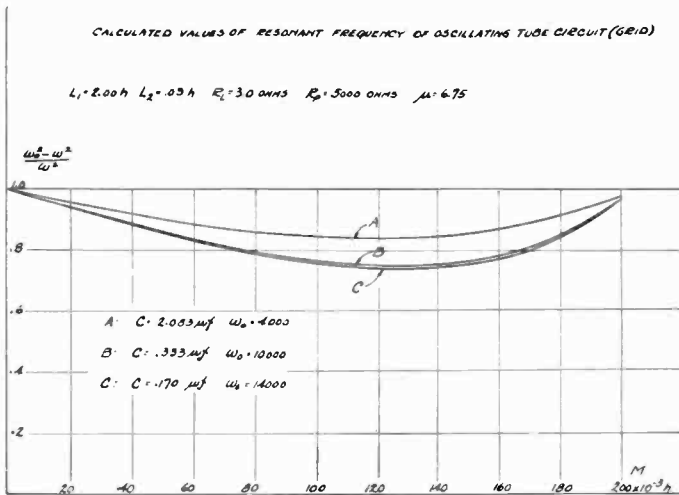

FIGURE 7

given in Figure 7. The value of the resistance of the oscillatory circuit, at the resonant frequency as fixed by the curves of Figure 7, is given in Figure 8. It is seen that this circuit, with the constants given, would oscillate until M approached the value of about 0.2 henry and with greater values of tickler coil (coupling being kept constant), the tube would refuse to oscillate.
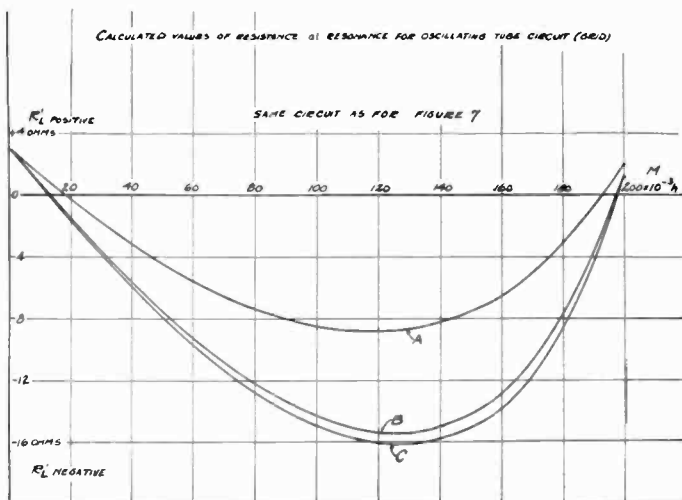
FIGURE 8

RESISTANCE AND REACTANCE OF THE CIRCUIT AS FREQUENCY IS VARIED

Using equations (48) and (49), the resistance and reactance of the oscillatory circuit were calculated, using the following conditions:

$R_L = 3$, $R_p = 5,000$, $\mu = 6.75$, $L = 0.03$, and for the values of C, 2.083 $\mu f.$, 0.333 $\mu f.$, and 0.170 $\mu f.$, which give resonance for $\omega = 4,000$, 10,000, and 24,000 radians, this covering the frequency range for which the circuit seemed suitable. Two values of L_1 were used.

$$\text{Case } a \qquad L_1 = 0.04$$
$$\text{Case } b \qquad L_1 = 2.0$$

The results, for resistance, are shown in Figures 9-11 and for reactance in Figures 12-14. The curves for reactance are nearly independent of the value of L_1, so only those for $L = 0.04$ henry are shown.

EXPERIMENTAL INVESTIGATION

The characteristics of the two triode circuits, the theory of which has been given in the previous paragraphs, were experimentally determined and the results are given herewith.

The circuit was arranged as in Figure 3, a Wheatstone bridge being used to get the resistance and reactance of the circuit. As long as the circuit had a sufficiently large capacity the impressed frequency (that of the power used on the bridge) was above the
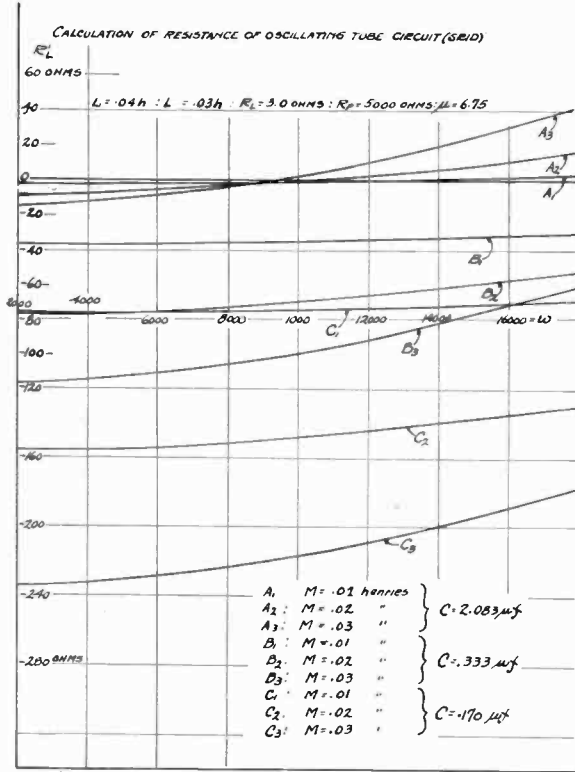
CALCULATION OF RESISTANCE OF OSCILLATING TUBE CIRCUIT (GRID)

R_L
60 OHMS
40
20
0
-20
-40
-60
-80
-100
-120
-160
-200
-240
-280 OHMS

$L = .04h$: $L = .03h$: $R_L = 3.0$ OHMS : $R_P = 5000$ OHMS : $\mu = 6.75$

A_3
A_2
A_1
B_1
B_2
B_3
C_1
C_2
C_3

4000   6000   8000   1000   12000   14000   16000 = ω

A_1 $M = .01$ henries
A_2 : $M = .02$ "
A_3 : $M = .03$ "
B_1 : $M = .01$ "
B_2 $M = .02$ "
B_3 : $M = .03$ "
C_1 $M = .01$ "
C_2 $M = .02$ "
C_3 : $M = .03$ '

$C = 2.083 \mu f$
$C = .333 \mu f$
$C = .170 \mu f$

FIGURE 9

resonant frequency of the circuit, its reactance was inductive, and a variable inductance was used in the bridge to get the reactance balance. With capacity sufficiently small to make the impressed frequency lower than the natural frequency of the circuit it was necessary to use a condenser to balance the bridge. As the plate current of the tube would, in this case, have to flow thru the telephones (the capacity arm of the bridge being open) and as this would ruin the telephones and make the plate voltage excessively low, the oscillatory circuit was opened in the capacity branch, thus permitting the plate current to flow to the plate without going thru the bridge at all. This is indicated in Figure 4.

A vacuum tube oscillator, giving a frequency of about 1,200 cycles was used as a source of power for the bridge. Measurements were made on the circuits for about fifteen values of M between zero and its maximum possible value, this representing about 80 percent coupling. Various condensers, the capacity and series resistance of which had been accurately determined,
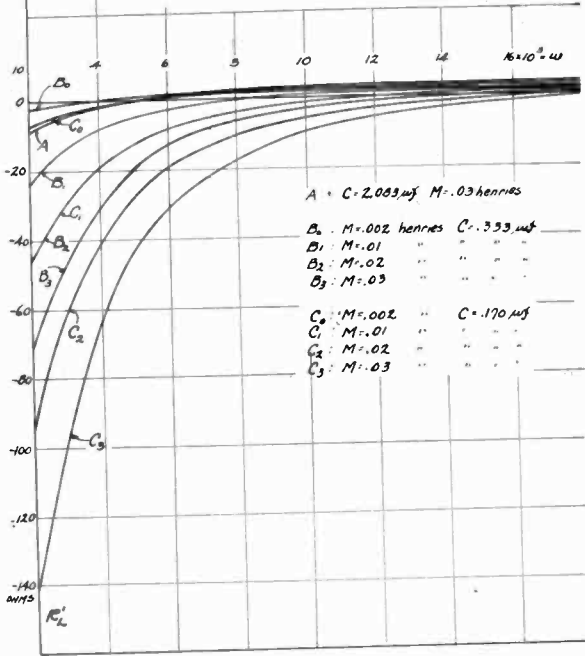
FIGURE 10

were used in the oscillatory circuit; their values ranged from seven microfarads to one-tenth microfarad. The value of the equivalent series resistance of the condensers involves both R_c and A of the theory, as no extra resistance was used in the capacity branch.

The experimental results have a precision of better than 1 percent in most cases; in the only cases where a greater discrepancy than this occurred it was later found that the tube had ionized and had thereby changed the plate circuit resistance. When first starting the tests the tube had a resistance (alternating current) of 5,115 ohms, but during the progress of the work, it ionized a few times, and its final value of resistance was only 4,300 ohms.

After performing a series of measurements, with fixed value of C and varying M, the oscillatory circuit, where it had been opened for bridge measurement, was closed, and the value of M was increased until oscillations started, to see how closely the
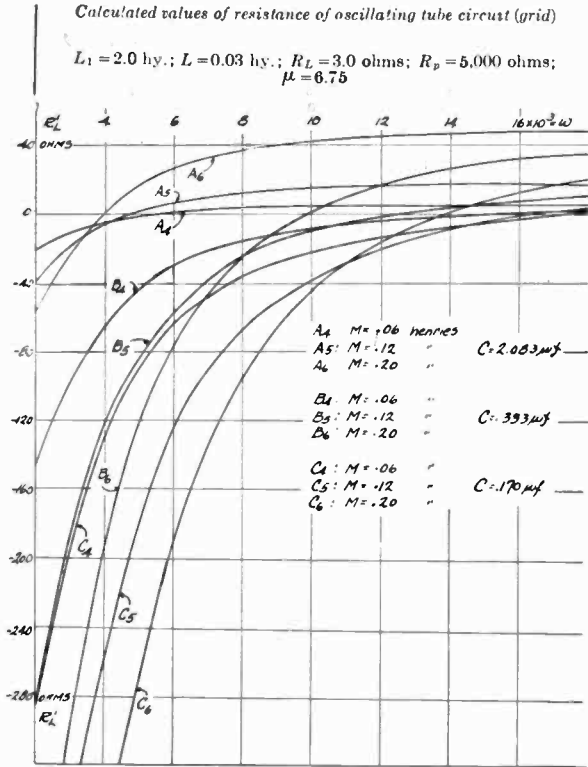
Calculated values of resistance of oscillating tube circuit (grid)

$L_1 = 2.0$ hy.; $L = 0.03$ hy.; $R_L = 3.0$ ohms; $R_p = 5,000$ ohms; $\mu = 6.75$

A_4 $M = .06$ henries
A_5: $M = .12$ " $C = 2.083 \mu f$
A_6 $M = .20$ "

B_4 $M = .06$ "
B_5: $M = .12$ " $C = 393 \mu f$
B_6: $M = .20$ "

C_4: $M = .06$ "
C_5: $M = .12$ " $C = 170 \mu f$
C_6: $M = .20$ "

FIGURE 11

condition of zero resistance corresponded to the condition for starting oscillations. As free oscillations must occur at a frequency fixed by the circuit constants, it is necessary that the circuit show negative resistance at the circuit's natural frequency before oscillations will start.

Thus altho the bridge measurement might show negative resistance, oscillations might not start when the circuit was closed because at the *natural* frequency the resistance may be positive. But if, in the bridge, the circuit showed zero reactance and negative resistance, it would always start to oscillate as soon as the circuit was closed.

Thus when the impressed frequency was the same as the natural frequency and the coupling was sufficiently weak that its effect on the natural frequency was negligible, the total resistance of the oscillatory circuit as measured in the bridge might look like that in Figure 15; if, with the oscillatory circuit closed, the coupling was increased until M was equal to 5.2 millihenrys, the circuit would at once begin to oscillate.
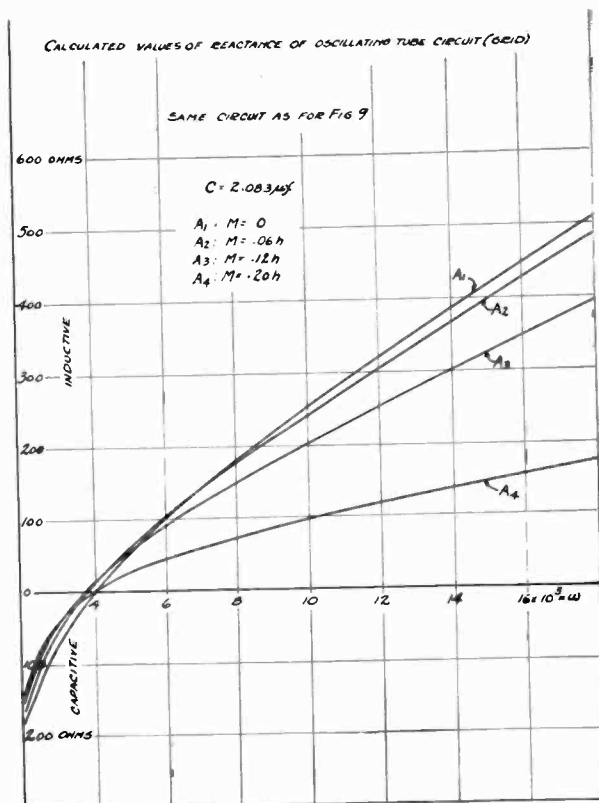
Figure 12

The results of the bridge measurements are shown in Figures 16-19. In Figure 16 are shown the curves of the resistance of the oscillatory circuit when this is in the plate circuit of the triode and the bridge is in the inductive branch of the circuit. The resistance for this case is given theoretically by equation (29); this is seen to predict a linear relation between resistance and mutual inductance and the experimental results show this to be so. The equation predicts that, for circuit constants as used in this test, the smaller the capacity the tighter must the coupling be made to reduce the circuit resistance to zero, and the curves substantiate the theory.

Curve 8 of Figure 16 shows the results when just enough capacity was used in the oscillatory circuit to make this resonant for the impressed frequency, so that the bridge showed zero reactance. With such a capacity the tube circuit will start to oscillate by itself, at this frequency, if the value of mutual in-
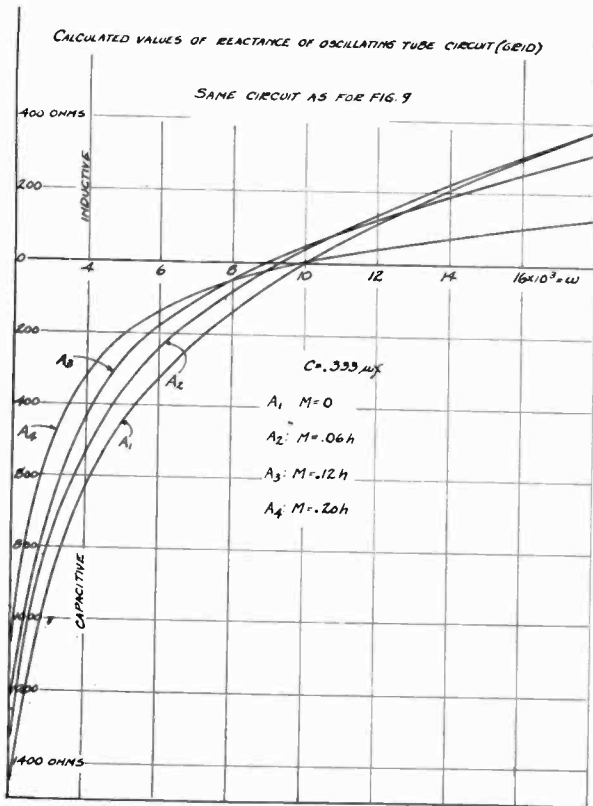
FIGURE 13

ductance is increased to 5.5 millihenrys or more. Having in mind the slight discrepancy between theory and experiment, due to the approximations made in simplifying the theory, experiment checks exactly with theory. As nearly as could be measured, oscillations started at once when the value of M demanded by equation (29), to make R_L' equal to zero, was reached.

The values of resistance used in plotting the curves of Figure 16 have had the equivalent series resistance of the condenser taken out, as this was entirely different for the different condensers used. The values plotted on the curve sheet, therefore, are $R_L' - r$, where R_L' is the circuit resistance measured in the bridge and r is the equivalent series resistance of the condenser used.

For the highest value of coupling used ($M = 27$ millihenrys) the values of $R_L' - r$ have been calculated from equation (29), for the various values of C used. These values, together with the

CALCULATED VALUES OF REACTANCE OF OSCILLATING TUBE CIRCUIT (GRID)

SAME CIRCUIT AS FOR FIG. 9

$C = .170 \mu f$

$A_1: M = 0$
$A_2: M = .03h$
$A_3: M = .06h$
$A_4: M = .12h$
$A_5: M = .20h$
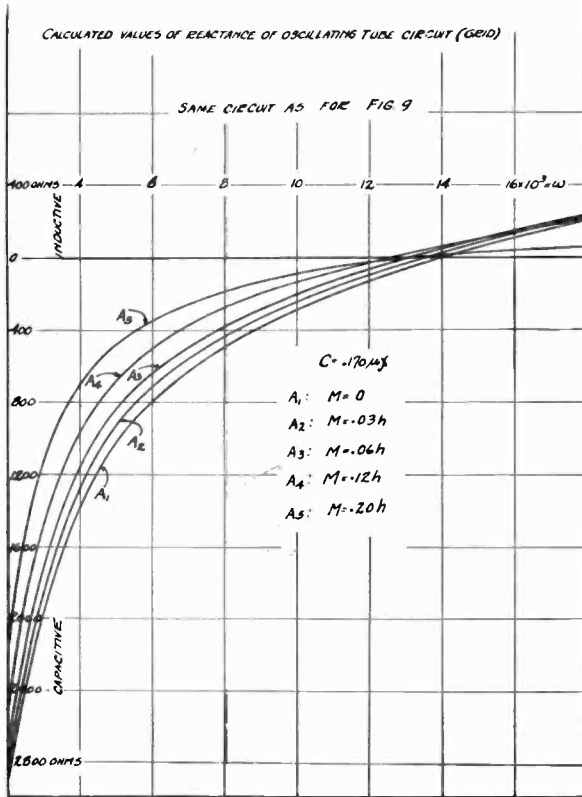
FIGURE 14

experimentally determined values, have been tabulated on the curve sheet. The agreement can be seen to be remarkably close, especially when it is remembered that 200 ohms of resistance was permanently left connected in the oscillatory circuit to prevent oscillations while measurements were being made. The agree-
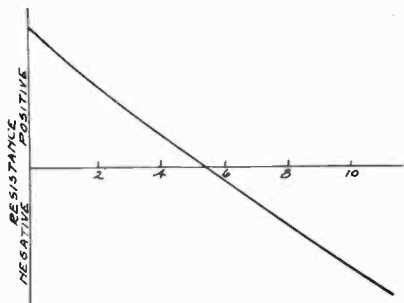

FIGURE 15

$L_1 = 39.83 \times 10^{-3}$ henries

$L_2 = 29.24 \times 10^{-3}$ "

$\omega = 8.19 \times 10^3$

CIRCUIT USED FOR MEASURING RESISTANCE OF OSCILLATORY CIRCUIT

CHECKS ON VALUES OF R

$R_L' - r$

20 OHMS

$R_L' = R_2 + r + \dfrac{\mu M}{C R_p} + \dfrac{1}{\omega^2 C^2 R_p}$

AND $r = 0$ WITH CONDENSER

OF NO LEAKAGE

	R_{CAL}	R_{EX}
8	55.62	56.75
7	50.17	48.90
6	45.04	45.20
5	41.73	41.10
4	37.57	37.60
3	34.45	35.60
2	32.14	33.40
1	29.23	29.25

POWER

28×10^{-3} henry

C IN μf

1 : 1.0183

2 . .9257

3 . .8616

4 . .7877

5 . .7041

6 . .6347

7 . .5744

8 . .5095

RESONANCE

FIGURE 16

ment of theory with the experimentally determined values is generally closer than $\frac{1}{2}$ percent.

When the capacity used was greater than that required for resonance it was necessary to introduce the bridge in the capacity branch of the oscillatory circuit as noted before; the theoretically derived resistance for this case is given by equation (32). This equation predicts that the resistance should be independent of the capacity used in the oscillatory circuit and such is seen to be the case for the experimental curves given in Figure 17. The equivalent series resistance of the condenser has again been subtracted from the measured value of circuit resistance. Here also the divergency of the experimental values from those predicted by theory is very small in nearly all cases, being much less than 1 percent.

When the oscillatory circuit is connected to the grid, theory shows the resistance should vary with M as given in equation (48). In Figures (18) and (19) are shown curves of the measured value
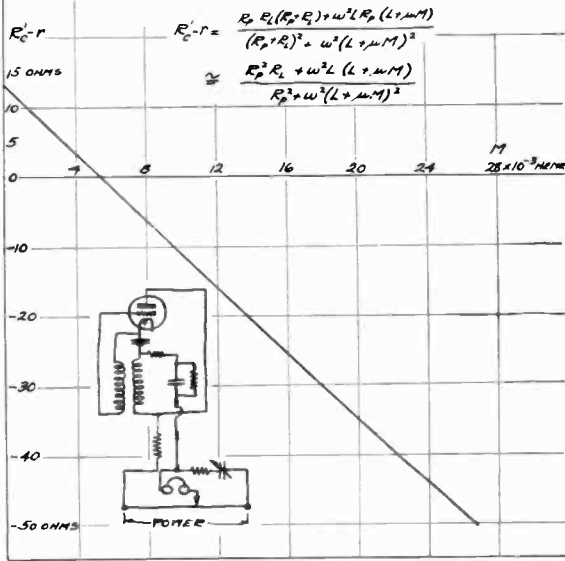
FIGURE 17

of the resistance of the circuit after $(R_L + r)$ has been taken out, these curves then give the experimentally determined values of the third term of the right-hand side of equation (48). It is here seen that the effect of coupling is much different than was the case when the oscillatory circuit was connected to the plate. The coupling introduces negative resistance into the oscillatory circuit for low values of M (this being negative in polarity) and positive resistance for high values of M. The amount of negative resistance developed in the circuit varies with the capacity used much more than was the former case. In Figure 18 the resistance is shown for the larger condensers used to bring out clearly the peculiar form of this curve and in Figure 19 are shown curves for the complete range of condensers used in the experiments; for the smallest capacity used, 0.0903 microfarad, the negative resistance generated with tightest coupling, exceeded 400 ohms. Other circuits with properly adjusted constants give thousands of ohms of negative resistance.

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It is, of course, to be always remembered that the negative resistance cannot actually exist as the sole resistance of the circuit; such a circuit would in general be unstable. These values of negative resistance are the values given by the interaction of the tube circuits and generally exist only if the positive resistance of the circuit is greater than this negative value, that is, the resistance of any stable circuit, as a whole, must be positive.
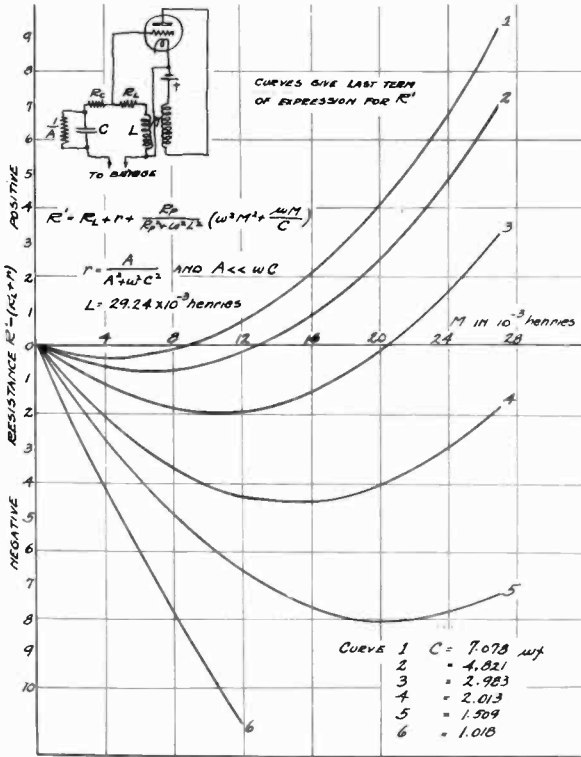
FIGURE 18

The experimental determinations given in this test for the oscillating condition were generally good to four significant figures, a precision easily obtainable with the alternating current bridge; the departure of these values from the theoretical values was always such as could have been predicted, being due to approximations in the theory or to a slow change in circuit constants as the tests progressed. Also R_L and r were determined only for one frequency and the measuring frequency differed

$$R' = R_L + r + \frac{R_p}{R_p^2 + \omega^2 L^2}\left(\omega^2 M^2 + \frac{\omega M}{C}\right)$$
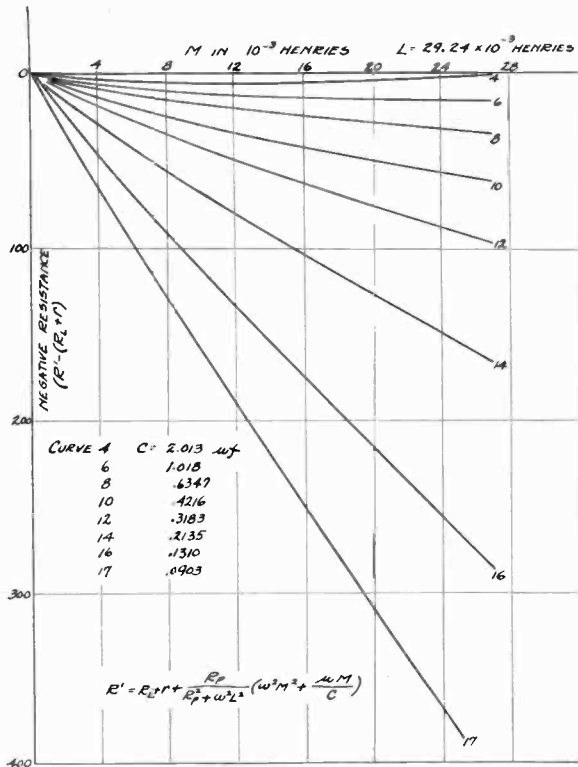
FIGURE 19

from this. Greater precision is possible, but it is felt that the results given fully substantiate the theory of these two circuits.

SUMMARY: Two typical triode circuits, used for obtaining regeneration by mutual induction, are examined analytically, with the idea of testing the results of the analysis by experiment.

As the experiments were to be performed with a one thousand cycle current (because of the greater accuracy of measurement at this frequency) it was permissible to neglect the condensive coupling due to the tube elements, and furthermore, as the conditions for the *starting* of oscillations were to be determined, the current taken by the grid could be neglected as it is only one or two microamperes in the non-oscillatory state.

Measurements made on the circuits, by using the alternating current bridge, confirm the theoretically derived relations to a remarkable degree of precision, showing that triode circuits are susceptible of accurate analysis.

DISCUSSION

Lewis M. Hull (by letter): The painstaking analysis and ingenious experimental methods set forth in this paper deserve praise. But it appears to me that the writers have directed their efforts away from those problems of triode operation which are of importance in modern radio practice. Their analyses are concerned solely with the first-order effects in two typical circuits: the tuned-grid and tuned-plate circuit, with inductive feed-back coupling. By first-order analysis I mean, in general, consideration of quasi-plane sections of the triode characteristic surface, which leads to the usual linear circuital equations with *constant* coefficients. The results of investigations of the two circuits discussed in this paper are applicable mainly to two general processes in tuned triode circuits: first, regenerative amplification of externally impressed oscillations, and second, self-generation of sustained oscillations. If we turn back to the literature on this subject, we find that the elementary results obtainable by first-order analysis alone have been derived again and again. With regard to regenerative amplification we find the classical papers of Vallauri, seven years old ("Elettrotecnica," **4**, 335, 350, 1917), of Ballantine (this journal, **7**, 2, page 155, 1919), and, more recently, the thoro analysis of Chaffee (this journal **12**, 3, 299, 1924). In fact, the framework of the typical first-order analysis of regenerative circuits has become essentially text-book material for modern engineering courses, in a class with the alternating-current transformer and the induction motor. As far back as 1918, moreover, Hazeltine carried the first-order analysis to its useful limit in the problem of generation of sustained oscillations by a triode (this journal, **6**, 2, 63, 1918). All of this work reaches the usual conclusions on negative resistance, conditions for starting oscillations, variations of national frequency with circuit constants, etc. To be sure, no previous investigator has measured those particular first-order effects described in the present paper. But the point I wish to emphasize is that it is unfortunate that much talent and diligence should still be applied, in the present day, to ferreting out the details of various circuit combinations on the basis of the old linear equations, when the real problems, which combine theoretical interest with immediate practical importance, are concerned almost wholly with the departures of the current-voltage surfaces of the triode from simple planes.

For instance, the radio technician is not so vitally interested in the properties of the negative resistance which a feedback

coupling introduces into a tuned circuit *provided* the assumption of linear characteristics is fulfilled: Armstrong discussed this negative resistance pretty thoroly without any reference to the calculus at all. The engineer wants to know why the ultimate current obtainable in a tuned receiving circuit without local oscillation is not independent of the original resistance of that circuit, as the linear negative-resistance equations predict. The answer to such a question can be found only by assuming the resistance of some element of the circuit to be a function of the *amplitude* of *oscillation*, which is the condition that actually obtains in most practical circuits. I have investigated this question mathematically, and have obtained an approximate solution of the equations for a variable grid-filament resistance (appropriate in the case of a regenerative detector tube) and for a variable plate-filament resistance, R_p. By considering variations with the amplitude in either of these two triode coefficients alone, it is possible to explain the experimental phenomenon of a progressive decrease in the generative amplification as the original or basic loss resistance in the tuned circuit is increased.

In the case of the triode generator it has been shown repeatedly, not only that the slope of the characteristic surface at the operating point (expressed in terms of amplification "constant" and plate-resistance) has little significance in determining the amplitude of oscillation, but that stable oscillations are actually impossible on plane sections of this surface. I wish to point out that the literature contains several *complete* discussions of the whole problem of oscillation with a triode having the necessary curved characteristics. Many years ago Barkhausen supplied a volume of analysis on the oscillating arc, much of which is directly applicable to the triode problem ("Das Problem der Schwingungs-erzeugungen," Pries, Leipzig, 1907). Van der Pol ("Radio Review" I, 701, 1920), gave an excellent discussion of the non-linear problem, using Lorentz's method of perturbations, but without experimental verification. Appleton and Van der Pol ("Phil. Mag.," 42, 201, 1921), carried the analysis further and obtained experimental verification of their results on triode oscillations over a wide curved range of the voltage-current characteristics. Two years ago I published in these Proceedings (**10**, 5, 373, 1922), a complete mathematical theory of sustained oscillations in a triode generating circuit which are of such wide amplitude as to extend far into the saturation region of the characteristic surface, and thus apply to conditions in which the triode is most efficient as a power converter. I consider that my method is

perhaps superior to Van der Pol's, insofar as it permits either exactness or approximation, as the occasion demands, and allows a prediction of the alternating-current power obtainable from a tube, and of the circuit constants necessary for maximum power output or maximum efficiency, all of which have been checked repeatedly in experimental practice. The extra cooling of a power-tube filament, due to the latent heat of evaporation of electron current, was considered, with the resulting errors which this effect introduces into the "static characteristics" of a power triode. Accordingly, a bridge method was developed for measuring at low frequencies the "generation constants" of a triode, from which its power output in any mono-periodic generating circuit are calculated. The methods described in my paper are of most value in connection with triodes of small power capacity; I have never applied them in practice to tubes larger than the conventional "five-watt" type. But the outstanding point is that the elementary non-linear viewpoint has long since been passed, in application to practical operation of triode circuits. In view of the quantity of theoretical and concordant experimental work which has already been published on the curved characteristics of the triode, the real "gaps" which remain to be filled in do not appear to be touched by the present paper. Undoubtedly many other investigators in this country have worked out methods for handling the curvature of triode characteristics in the circuit equations; they would perform a further scientific service by publishing their results in these PROCEEDINGS.

J. H. Morecroft (by letter): It seems as tho Dr. Hull had failed to read the first paragraph of our paper, otherwise most of his reiteration is difficult to understand. The excellent work of Armstrong, Hazeltine, Ballantine, himself, and other investigators, as published in the PROCEEDINGS, does not give experimental proof of the theory (insofar as I can see), and, unimportant as it may seem to a research man of Dr. Hull's type, the experimentally obtained curves, substantiating the theory, are often of more value to the general worker and student than the theory itself.

Not all papers can be original in the sense that they deal with a subject never touched before—not all of us can be Faradays or Hertz's. In 1919, I published in the "American Institute of Electrical Engineers Proceedings," a paper on the triode as a power converter, giving operating curves exactly like those given by Dr. Hull in his 1922 paper in these PROCEEDINGS, and many

other curves perhaps more important than those Dr. Hull gave three years later. But that did not make his paper of no value! Five years ago, in my radio text, I discussed in an elementary way the effects of the variation of the plate circuit resistance of a triode, but that doesn't make valueless any contribution Dr. Hull may add, due to research activities he is carrying out at present.

The mathematically inclined research worker will do well to remember that not all people think as he does; some have to progress by routes which involve more detailed effort and repetition than he thinks worth while.

EXPERIMENTAL DETERMINATION OF THE FUNDAMENTAL DYNAMIC CHARACTERISTICS OF A TRIODE*

By

Eijiro Takagishi

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CONTENTS

I. General Considerations

Because of the recent development in vacuum tube technique, not only the field of radio communication but also several other branches of science have made such rapid progress that one could hardly have foreseen these wonderful changes even a few years ago. Nevertheless, progress has been rather slow in the experimental determination of the real functioning of the triode.

The writer believes that more efficient and correct utilization of the triode can be attained only thru exact knowledge of its characteristics. Up to the present, the characteristics of the triode in its static condition have been much investigated; while on the other hand, when dealing with its performance as a generator or an amplifier of oscillating current or potential, the dynamic characteristics become important in order fully to understand the operation of the system.

As is well known, the current flowing in a triode in actual operation is not free from distortion, and large harmonic com-

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ponents are present in addition to the fundamental. When a triode is used in an electric circuit which can be actually tuned to a resonant frequency, the relation between the various electrical quantities is ordinarily important only insofar as it relates to their fundamental frequency.

With this idea in mind the writer carried out some experiments for measuring the fundamental components of electrical currents in triode circuits under conditions similar to those found in practice. The results are described below, but it is very regrettable that the writer is unable to give them quantitatively, because all of the records of the investigations were destroyed during the recent conflagration in Tokyo.

II. Methods of Measuring Fundamental Frequency Components in Distorted Electric Currents

Three methods may be considered for measuring the fundamental component in a distorted electric current. These are:

(a) Tuned circuit method
(b) Wattmeter method
(c) Vibration galvanometer method.

In the first method, there is used a tuned circuit consisting of inductance and capacity, the natural frequency of which circuit is equal to that of the current to be measured. This method is convenient for the measurement of currents of higher frequencies but is not so practically suitable for that of lower frequencies.

The second method utilizes the following characteristics of a wattmeter. When sinusoidal currents of the same frequency are made to flow thru both of the two coils of a wattmeter, it always indicates their product, and even when one of them is replaced by a distorted current, the wattmeter indication remains unchanged so far as the fundamental of the distorted current does not change both in amplitude and phase. The wattmeter can be calibrated by using a sinusoidal current obtained from an ordinary power source and it can then be readily used for the measurement of the fundamental component of any distorted current.

This method is ingenious in theory and also simple and convenient for experiment, but it should be borne in mind that even a slight distortion in the auxiliary electric current causes a large error, especially when the current to be measured contains a large proportion of harmonics. For example, only five percent of the third harmonic in the auxiliary current and three hundred

percent in that to be measured give rise to an error of fifteen percent in wattmeter reading.

The last method depends upon the principle of alternating current potentiometer, using as an indicator a vibration galvanometer, which is resonant to a fundamental frequency.

When a sinusoidal current i_s flows thru a conductor A (Figure 1), and the current i_d which is to be measured flows thru the other conductor B, and a vibration galvanometer is bridged across the two resistances R and r (which are inserted in each conductor), the galvanometer will cease to vibrate at a definite value of r/R. Then the fundamental component of i_d is found to be $i_s \dfrac{R}{r}$. By the use of a galvanometer which does not respond to any other harmonics, the zero point will remain unchanged as long as the amplitude of the fundamental in i_d is kept constant. Even in the case when the ammeter reading of i_s is used directly, instead of its fundamental component, the percentage error is actually small. For example, if i_s contains five percent of harmonic components, the error as a whole has a value of only 0.12 percent.
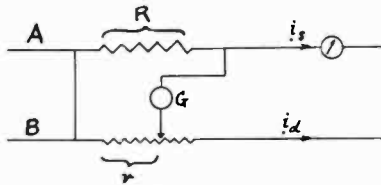
FIGURE 1

III. MEASURING ARRANGEMENT FOR THE DETERMINATION OF THE FUNDAMENTAL DYNAMIC CHARACTERISTICS OF A TRIODE

Suppose that a triode V (Figure 2) behaves as a generator of radio frequency current, impedances Z_g and Z_a being placed in the grid and anode circuits.

If the voltage drops in Z_g and Z_a are e_g and e_a, we may substi-
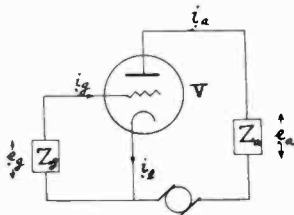
FIGURE 2

tute for each of the impedances a generator of an emf., which is equal to that across the respective impedances with respect to both amplitude and phase. The practical experimental arrangement, which is based on this idea, is shown in Figure 3, the following abbreviations being used:

V—triode, the characteristics of which are to be studied,

B_a—battery or direct current generator in anode circuit,

B_g—battery in grid circuit,

B_k—battery in cathode circuit,

T_a—transformer to supply anode circuit with e_a,

T_g—transformer to supply grid circuit with e_g,

T_l—transformer in the local circuit,

F_g—phase transformer in grid circuit,

F_l—phase transformer in local circuit,

r_a, r_g, r_l—resistances in anode, grid, and local circuits, respectively,

r_p—potentiometer,

A_a, A_g, A_k, A_l—ammeters in anode, grid, cathode, and local circuits, respectively,

V_g, V_a—voltmeters in grid and anode circuits,

G—vibration galvanometer,

S—change-over switch,

t—ground connection.

The phase transformer is such that single phase voltage in any phase may be drawn from the rotor, three-phase voltage being applied in the stator and the rotor being turned. The electromotive forces e_a and e_g are applied to the anode and grid circuits by means of the transformers T_a and T_g. There then flows in each circuit a current of complicated wave form, which becomes more distorted as the applied voltage grows. The fundamental alternating components in the distorted currents are denoted by i_a and i_g. When i_a and i_g are made to flow thru r_a and r_g, the voltage drops are proportional to the currents flowing. The currents are then measured by potentiometer method, the drops being balanced to the sinusoidal voltage drop in r_p. The phase transformer F_l supplies, thru a transformer T_l, a current of pure sine form flowing in a local circuit which includes a potentiometer resistance r_p.

We first adjust the local current to a certain amount (for example, one ampere). The vibration galvanometer G is bridged across r_a and a part of r_p. When the galvanometer just ceases to

vibrate, the fundamental voltage drops in r_a and the included part of r_p are equal to each other, that is, i_a is found by the following equation

$$i_a = i_l \left(\frac{r_p}{r_a} \right), \tag{1}$$

where i_a signifies here the fundamental of the distorted alternating current and i the sinusoidal current in the local circuit.
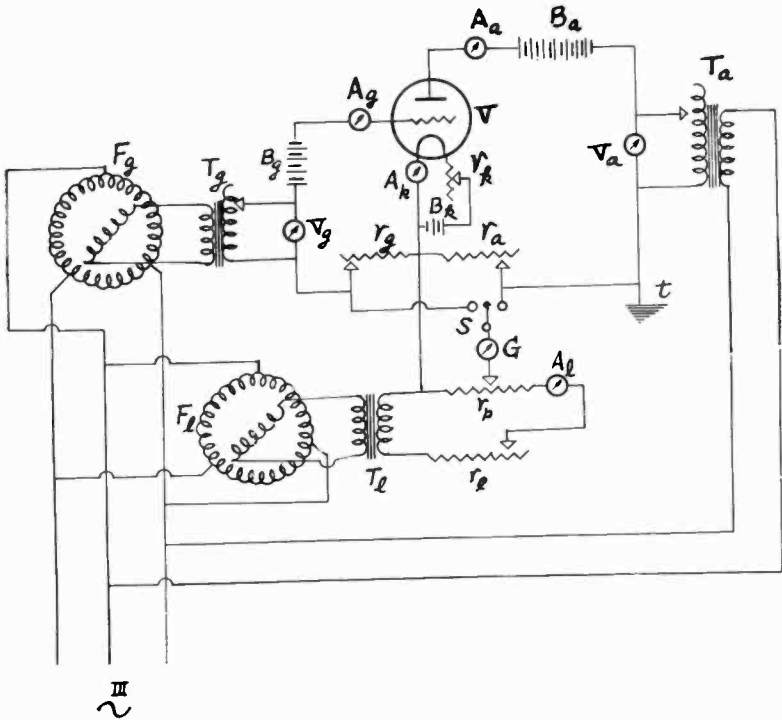
FIGURE 3

In cases where both amplitude and phase of e_a and e_g are varied to various extents, the current i_a can be measured easily and correctly by adjusting F_g, F_l, and so on. The distorted current i_g in the grid circuit can also be measured similarly. The writer has made some experiments with the case where e_a and e_g are in opposite phase, the arrangement being shown in Figure 4, which was chosen because he did not possess a phase transformer F_g, as required in the arrangement of Figure 3. A transformer T was used, which has two sections, S_a and S_g in the secondary, the former being for the anode and the latter for the grid circuit. L_{g1} and L_{g2} are for adjustment of the phase of e_g; L_{g1} serves to

produce a lag and L_{g_2} a lead relative to the terminal voltage at S_g. Resistance R_g serves for fine adjustment of the amplitude of e_g.
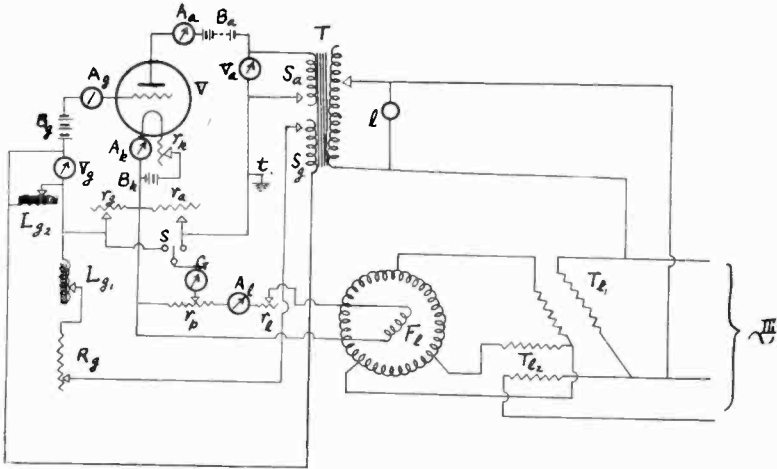
As the phase transformer F_l, the writer made use of a direct current shunt motor of one-half horsepower, the whole winding of its armature being divided into three symmetrical sections in order to render it available for three-phase supply, and two new windings, each having twenty turns, being placed on the existing field coils. 100 volts of three-phase supply were too high for safe direct application to the armature, and two transformers were used, by which 100 volts were stepped down to about 20 volts. Thus the induced 8 volts in the secondary was nearly cotnstan for all positions of the rotor. This emf. produced an electric current of less than one ampere in the local circuit. The instruments and arrangements used will be explained below in more detail.

Referring to Figure 4:

T—primary 100 volts, 6 steps, secondary voltages finely adjustable, secondary max. 1,000 volts, S_g max. 250 volts,

l—100-watt incandescent lamp, used for suppression of self-oscillation of the transformer T associated with the triode V,

T_{l_1}, T_{l_2}—single-phase transformer, primary 100 volts, secondary 20 volts,

V_a—alternating current voltmeter, max. 1,200 volts.

V_g—alternating voltmeter, hot-wire type, 100 volts, max., about 600 ohms resistance,

A_a—direct current milli-ammeter, 500 milli-ampers max.,

A_g—direct current milli-ammeter, 150 milli-amperes max.,

A_k—direct current ammeter, 5 amperes, max.,

A_l—alternating current ammeter, 1 ampere max.,

r_p—5 ohms, about 50 cm. (19.4 inches) long, straight, B. and S. number 32,

r_l—20 ohms, 2 amperes,

r_a—22 ohm—44 ohm—220 ohm—and 440 ohm—taps producing 2.2 volts of potential drop by 100 m.a., 50 m.a., 10 m.a., and 5 m.a. of anode alternating current, respectively.

r_g—similar construction to r_a,

R_g—8 ohms, 2 amperes,

L_{g_1}—small iron core inductance coil for fine adjustment of phase,

L_{g_2}—large iron core inductance coil for fine adjustment of phase,

B_a—dry battery, max. 1,500 volts,

V—triode, manufactured by the Tokyo Electric Company, Ltd., 250 watts, cathode current 3 amperes,

B_k—accumulators, 20 volts,

G—vibration galvanometer, re-constructed from a 800-ohm telegraph relay; instead of the tongue, a vibrator of thin iron plate, was used, this tongue being about 0.08 mm. (0.03 inch) thick, adjusted to vibrate vigorously at a resonant frequency of 50 cycles, and operative down to 0.1 volt or 0.05 milli-ampere.

The current in the local circuit was maintained constant at 0.44 ampere in order that fifty centimeters (19.7 inches) in length of the potentiometer resistance wire may have a voltage drop of just 2.2 volts. If 22 ohms of resistance are used for r_a, then i_a is measured by

$$i_a = 100 \times \frac{\text{(length of resistance wire } r_p \text{ in cm.)}}{50 \text{ cm.}}$$

with 44 ohms for r_a

$$i_a = 50 \times \frac{\text{(length of resistance wire } r \text{ in cm.)}}{50 \text{ cm.}}$$

with 220 ohms

$$_a = 10 \times \frac{\text{(length of resistance wire } r \text{ in cm.)}}{50 \text{ cm.}}$$

and with 440 ohms

$$i_a = 5 \times \frac{\text{(length of resistance wire } r \text{ in cm.)}}{50 \text{ c.m.}}$$

(2)

For the measurement of larger currents, the current i_l was increased to 0.66 ampere.

The process of measurement will be explained for the above-mentioned arrangement. We first keep the triode V in its normal condition, for example, with 750 volts of direct anode potential, 2.8 amperes of cathode current, and we then apply voltages from the transformers to the anode and grid so as to make e_a and e_g opposite in phase to each other. e_g and e_a cannot always be kept in exactly opposite phase, since the phase varies according to the amount of load on the transformers. To take care of this condition, L_{g1} and L_{g2} serve to adjust the phase precisely. We then throw the switch S to the left, slide the contactor of the potentiometer, and at the same time turn the rotor of the phase transformer F_l until the galvanometer ceases to vibrate, at the same time keeping the alternating current i_l in the local circuit at a constant value. We then throw the switch S to the right and observe whether the null point of the galvanometer be found on the potentiometer or not. Generally this is not the case, for i_a and i_g are not exactly in opposite phase. In that case, it is necessary to adjust the phase of e_g by L_{g1}, L_g, R_g, and so on, and find anew the desired condition. For example, if e_g lags a little behind e_a, i_a also lags behind it. It is then necessary to make e_g lead but a little in order that i_a be in phase with i_l; in that condition, however, the phase of e_a does not coincide with that of e_g, but leads it by a small amount. Thus if e_g leads quite a little, the desired condition will readily be found, where all the voltages and currents e_a, e_g, i_l, and so on, are in the same phase. Under these conditions, we must throw the switch to the left and then to the right, and the null point will be found in both cases, and thus i_a and i_g may be measured directly on the scale of the potentiometer.

Typical results of the measurement are shown in Figures 5, 6, 7, 8, and 9.

IV. Dynamic Characteristics of a Triode

Figure 5 shows the relations between effective values of the fundamental components of anode current (A), grid current (G), and their total (E) against applied grid voltage at a definite ratio of e_g and e_a, which latter are in exactly opposite phase. It is interesting to note that curve A has a peak, past which the curve falls rapidly, while curve G begins to rise suddenly at that point and reaches a high value. This indicates that, in such conditions, the grid absorbs the larger part of the electrons emitted from the incandescent cathode.
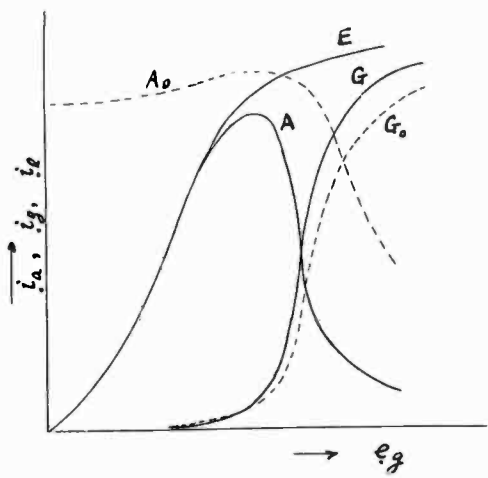
FIGURE 5
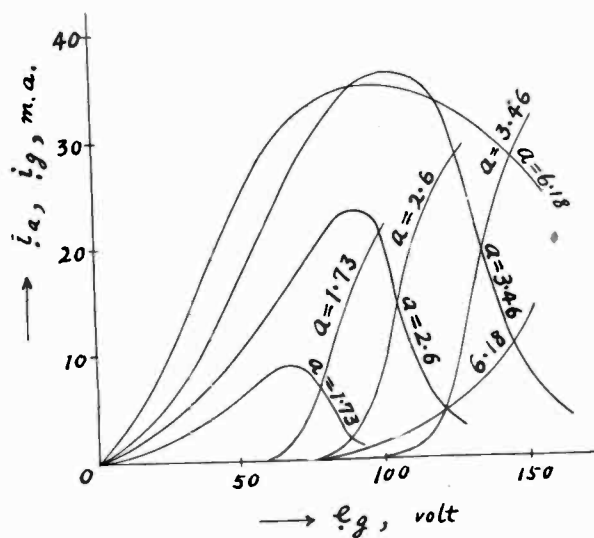
FIGURE 6
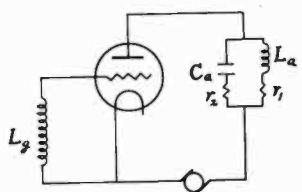
FIGURE 7

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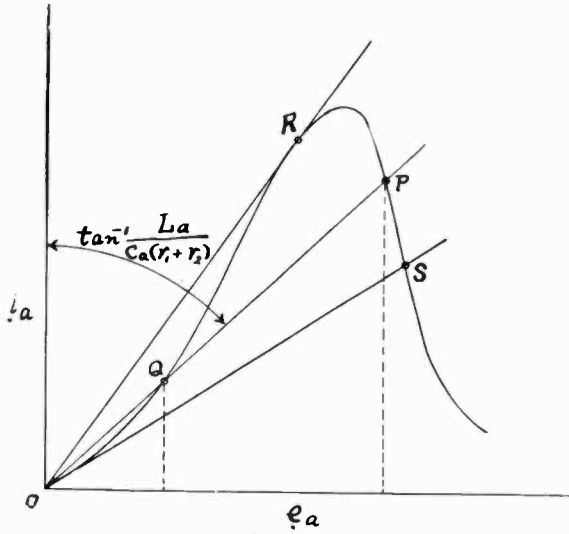
FIGURE 8

As to curve E, however, it never drops. In the figure, it is seen that grid current varies widely, depending upon the alternating voltage applied to the anode, and in the following equation for grid current a factor ν cannot be neglected,

$$i_g = g \ (e_g + \nu \, e_a) \tag{3}$$
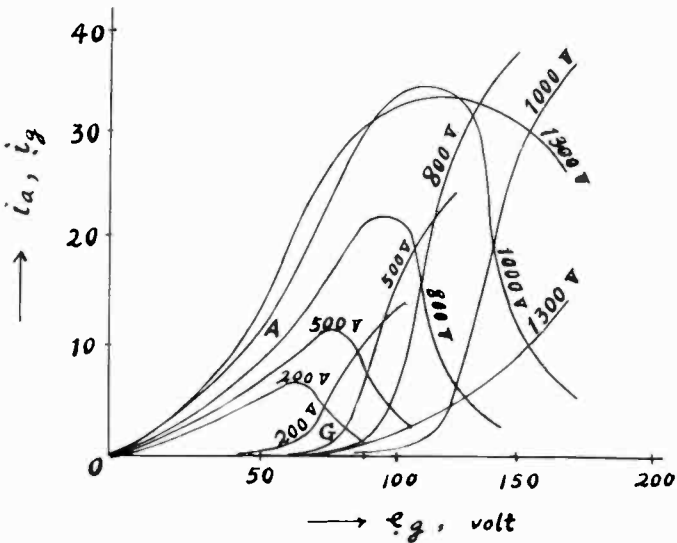
FIGURE 9

The presence of grid current and the factor ν are both important in studying the operation of a triode. Their exact relations are not yet worked out.

Dotted lines A and G in the figure show the corresponding values of direct current in the anode and grid circuit. Curves A and G in the figure change in form for different values of e_g/e_a, as shown in Figure 6. The values of the ratio e_g/e_a multiplied by μ is denoted by a, where μ is so-called voltage amplification factor. In the present case, μ was measured and found to have a nearly constant value of 26, and a was selected between 1.73 and 6.18.

The characteristic curves explained above will be of great service for the predetermination of the stable oscillating condition of the triode. As a simple example, let a system oscillate with the circuit arrangement as shown in Figure 7, using a triode, the characteristic of which is known to have the form shown in Figure 8 for a certain value of a. In this case, the expression $\mu \dfrac{M}{L_a}$ can be used approximately for a, M being the mutual inductance between L_a and L_g.

Strictly speaking, the wave length at which the system is maintained in oscillation is not exactly equal to the value calculable when considering inductance and capacity only, but in the present case, for the sake of simplicity, the only capacity besides the condenser capacity which was considered in the determination of the wave length was that of the lead wires, which eliminated an appreciable error. Let C_a signify the resultant total capacity. Then the oscillating circuit consisting of L_a, C_a, r_1, and r_2 offers an equivalent resistance equal to $L_a/C_a\,(r_1+r_2)$, at the wave length calculated from L_a and C_a. We draw the straight line $O\,Q\,P$ (Figure 8), which has a slope equal to $tan^{-1}L_a/C_a(r_1+r_2)$, and intersects the dynamic curve at two points P and Q. One of these two points, P corresponds to the condition found in practice, where the values of e_a and i_a, and consequently that of the oscillation current are definite. If C_a is made larger and larger, the straight line $O\,Q\,P$ rises more and more steeply, and at a point R becomes tangent externally to the curve, and there oscillation dies away. To start oscillation again, it is necessary to give the straight line a greater slope, in other words, to make C_a smaller. Then the system begins to oscillate at a point S, $O\,S$ being tangent internally to the curve. In this case e_a and consequently e_g have rather larger values than those in the former case, but so far as the amplitude of the oscil-

lation current in the resonant circuit is concerned, it is not proportional to e_a, but is rather smaller, because then the equivalent resistance has a larger value.

This phenomenon is already known as" Ziehen"[1] in Germany, and has been investigated in that country altho the investigators do not use dynamic characteristics as explained above, with the exception of Dr. Hans Georg Möller,[2] who, however, did not show the falling characteristics for the anode.

From the characteristics of a triode and its associated circuits, it will be fully understood that this phenomenon accompanies changes in wave length and that it takes place only when the characteristic has an inflection point on the left-hand side. The values of L_a, M, L_a/C_a (r_1+r_2), and so on, being measured and calculated, the writer predetermined, with C_a, $M_,$, and so on, as variables, the values of the oscillation currents according to the methods explained above and observed good agreement. As the direct voltage E_a on the anode is varied between 200 volts and 1,300 volts, the corresponding alternating anode and grid currents vary according to the curves in Figure 9, by which means of which modulation in the anode circuit can be predetermined. If the heating current thru the cathode is increased, those curves grow higher, as might be expected. If E_g were varied, similar curves might be expected. Unfortunately the great fire in Tokyo intervened before these measurements were started. It is regrettable that the results are, for that reason, given only qualitatively (against the writer's inclination).

The writer, however, believes that he will be able to publish the further details of the experiments later on. The above-mentioned instance is only for the special case where e_a and e_g are in opposite phase, while in other cases, matters are not so simple, but become more interesting and important.

V. Voltage Amplification Factor or Constant

The anode current is usually expressed by the following equation

$$i_a = h \left(e_a + \mu e_g\right) \tag{4}$$

where h is the variable conductance of a triode and μ is a factor called "voltage amplification factor," which is supposed to be constant with respect to a triode. In the equation it is seen that i_a is always zero when e_a plus e_g is equal to zero, and following

[1] "Ziehen" may be translated as "pulling" (intermittently) or "tugging."
[2] Dr. Hans Georg Möller, "Die Elektronenröhren und ihre Technischen Anwendung."

this idea the writer measured this factor with the arrangement already mentioned, the results being shown in Figure 10. The dotted lines show the relation between e_a and e_g under the condition that i_a is zero, that is to say, the slope of any straight line $O P$ defines the value of μ in that condition.. The lines are curved as shown, in other words μ is not constant.
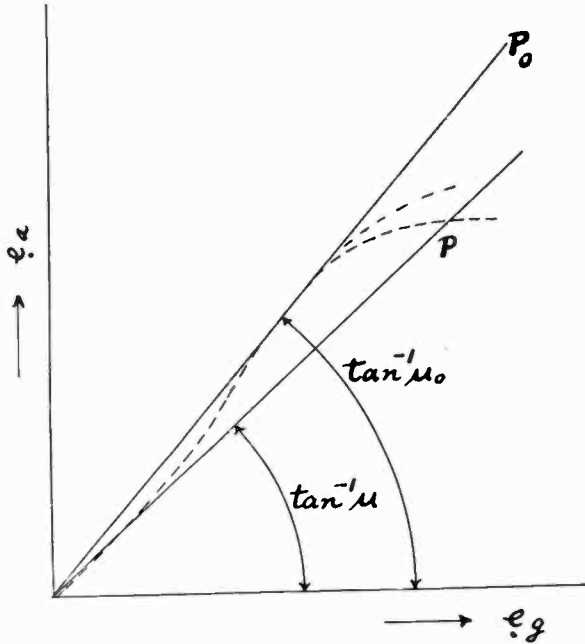
The cause of this is merely the existence of grid current, which is proved by the following experiments. The writer assumed the correctness of the equation

$$i_l = g_o \left(e_a + \mu_o \, e_g \right) \tag{5}$$

where i_l is the fundamental component of current emitted from the cathode, and g_o and μ are the factors just defined. The measurement reveals the fact that the relation between e_a and e_g under the condition that i_l is zero is shown by a straight line $O P_o$, which remains unaffected by great changes in the values of the variables. Thus the newly established constant of voltage amplification is really constant for a given triode. The writer desires to express his obligation to Mr. E. Yokoyama, under whose direction he carried out the present experiments, and also to

express his deepest indebtedness to Mr. S. Kawazoe for his valuable assistance thruout the whole series of experiments.

Synopsis

1. In studying the action of a triode, the static characteristics have been much considered, but little attention has been paid to the fundamental dynamic characteristics, which show the relation of much distorted currents and potential drops in a triode with reference to their fundamental components only.

2. A method of measuring the fundamental component in distorted electric current was adopted to determine the above mentioned characteristics of a triode. The experiment was confined to a special case, where grid and anode voltages are kept exactly in opposite phase, because all the apparatus and the arrangements had been destroyed by the recent conflagration in Tokyo before further experiments were executed. The circuitual arrangements of the measurements are fully described. Some of the fundamental dynamic characteristic curves of a triode thus determined are shown and explained.

3. It has been found by the experiment that the anode current-grid voltage characteristic curve has a falling region in a working condition and that the current which flows thru the grid circuit generally has a large value.

4. Some characteristics of a triode in stable oscillation were predetermined by the help of the characteristic curves and were in a good coincidence with those directly measured, the triode being operated as an oscillator. The phenomenon of "Ziehen" was briefly mentioned.

5. It was observed in the experiments that a voltage amplification factor or constant as at present defined is not of constant value, but is subjected to variation with applied voltages. A new constant has been established, which relates to the emission current from the incandescent cathode of a triode instead of the anode current.

6. It is much regretted by him that the writer is unable to give the results of the present researches quantitatively because of the recent conflagration at Tokyo, but he expects to go ahead with the experiment in the near future.

SUMMARY: By picking out the fundamental component only from the distorted current in a triode while functioning, the writer has found first new forms of dynamic characteristics for a triode, which are utilizable for solving various problems not hitherto satisfactorily solved (for example, modulation in radio telephony, the phenomenon of "Ziehen," or "pulling," and so on), and secondly that the grid current may reach a large value and that the factor or constant ν merits attention. Some experiments have been carried out relative to the voltage amplification factor for which a new expression has been found. The work is being continued.

A HIGH EFFICIENCY VACUUM TUBE OSCILLATING CIRCUIT*

By

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INTRODUCTION

In preceding issues of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, methods of calculating the performance of vacuum tube oscillators have been developed.[1] Except in Chapter VII of the cited papers, sinusoidal voltages only were considered. Chapter VII develops the method of calculating the increase in efficiency and output when the circuit is made responsive to more than the fundamental frequency.

A circuit called the "push-pull" circuit has been developed in which two tubes are operated 180° out of phase. Several circuits of this sort have also been suggested in which the idea that each tube should carry current half of the time and that the current form should be square in each tube (the summation current being held constant by an inductance in the supply lead), has been added to the two-tube idea.

The writer has spent considerable time reducing this form of circuit to a design basis. In doing this it has been necessary to make several changes in the original circuit. The present paper develops this study to a point where intermediate frequency circuits of very high efficiency and large specific output can be designed with reasonable certainty that they will work as desired.

SQUARE WAVE OSCILLATOR

With the usual type of sine-wave oscillating circuit, current is drawn thru the tube for a very small part of the cycle when high efficiency is sought. The reason for this is that, since the plate voltage is varying sinusoidally, it will remain at a minimum for a very short time. During the remainder of the cycle the

drop is high so that any current passed is accompanied by large loss. However, since the purpose of this discussion is to develop the design of a circuit of high output we will deliberately increase the angle during which current flows and then study ways of reducing the losses. Since a given thermionic valve has a definite amount of emission, the largest input will be obtained by taking the full emission while the current flows. For reasons which will appear later, the circuit will be made symmetrical so that each of two tubes may draw current of constant value during one-half of the cycle. Such a circuit is shown in Figure 1. This circuit of Figure 1 is not usefully operative, so no grid control is shown. Figure 2 shows the plate drop e_p, plate current i_p, and
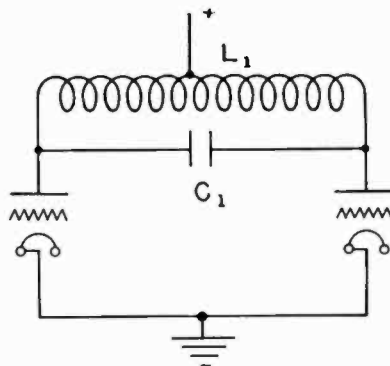
FIGURE 1

loss w_p^r for one of the two tubes. It is apparent that the losses at points "a" are much higher than at points "b," because the voltage drop across the tube is high at those points. At points "b" the drop need be only sufficient to overcome the space charge of the tube.
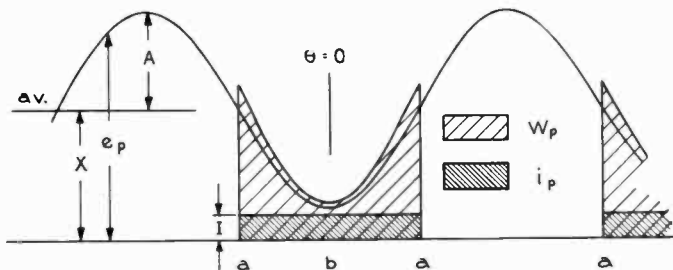
FIGURE 2

The power input to the circuit is the product of current I and volts X, both of which are constant. The plate voltage is

$$e_p = X - A \cos \theta$$

Instantaneous loss is $W = e_p I = I (X - A \cos \theta)$

$$\text{Efficiency} = \frac{I X - \frac{1}{\pi} I \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} [X - A \cos \theta] \, d\theta}{I X}$$

$$= \frac{A}{\pi X} \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} \cos \theta \, d\theta$$

$$= \frac{2 A}{\pi X}$$

Thus the efficiency is 0.637 times the ratio of counter emf· maximum to impressed direct voltage. A is always less than X.

Since we are trying to improve the efficiency without changing the current wave, we must proceed by changing the voltage wave. From Figure 2 it appears that the loss could be reduced to a minimum if the plate voltage e_p were kept at the minimum value necessary to overcome space charge for the entire period during which current flows. This would mean a voltage wave such as shown in Figure 3.
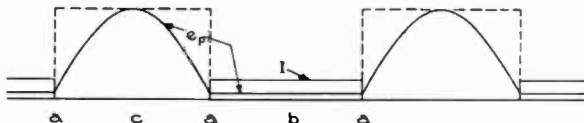
FIGURE 3

It is quite immaterial what the plate voltage is for the period "a c a," since no current is flowing. The important thing is to have e_p a minimum for the period "a b a" during which current is flowing. Whether the values of e_p should follow the full or dotted line for the period "a c a" is determined by the method used to control the potential in the period "a b a."

By the method described in Chapter VII, of the previously cited papers, a symmetrical square voltage wave could be obtained by employing a considerable number of trap circuits tuned to various odd harmonics. The circuit would, however, be extremely complicated and difficult to handle.

The circuit of Figure 1 may be added to as shown in Figure 4, by the addition of the choke coil L_3. If L_3 be

made very large it will maintain the current at a constant value regardless of any cyclic voltage change. If the grid excitation is such as to pass this constant current for one-half cycle thru each tube with low plate potential, the remaining drop must be absorbed by the inductance L_3. Thus, with proper grid excitation, the voltage e_p could vary in accordance with the full line of Figure 3. Figure 5 shows the various voltage rela-
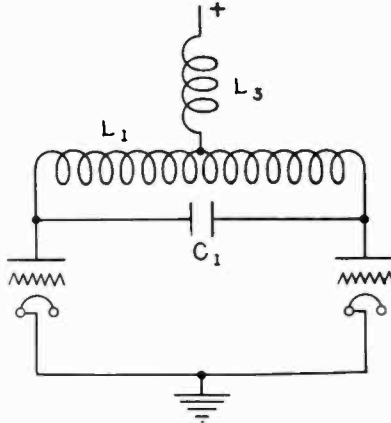
FIGURE 4

tionships, e_{p_1} and e_{p_2} are the plate potentials of the two tubes while e_{L_3} is the average potential at the mid-point of L_1. X is the average value of e_{L_3}, since there can be no constant difference of potential across an inductance such as L_3. The difference between e_{p_1} and e_{p_2} is the potential difference across L_1 and C_1 and is sinusoidal, since L_1 and C_1 constitute a resonant circuit. With this arrangement the energy put into the circuit at times "a" is stored in L_3 and does not appear as tube loss. At points "c" and "b" it is passed on to the oscillating circuit with no loss of efficiency.
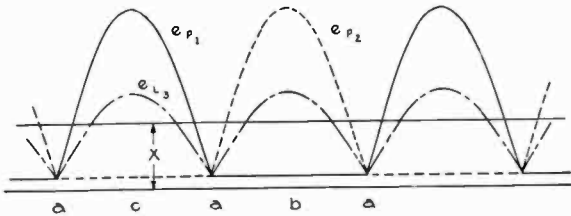
FIGURE 5

The necessary condition to secure this highly efficient voltage relationship is that the tube control must not limit the current while current is flowing. The valve must be either wide open or entirely closed. Examination of the tube characteristic, Figure 6, shows that plate current and grid current are roughly proportional. A control system which will produce square blocks of grid current will then presumably produce square blocks of plate current at constant plate volts. Square blocks of grid
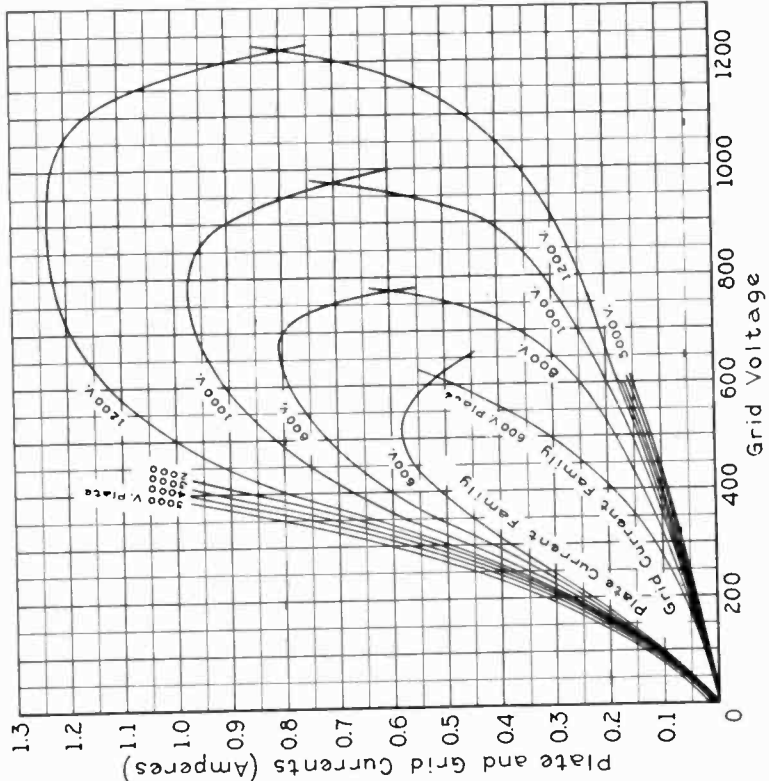
FIGURE 6—Characteristics of a UV-206 Radiotron with High Emission

current can be obtained by the arrangement shown in Figure 7. Energy is supplied to the grid by coupling between L_1 and L_2. The total grid current is kept constant by inductance L_4.

This grid current naturally flows from that grid which is most positive at any instant, and since the total is constant, each grid receives current in square blocks as required.

If L_3 and L_4 are not large enough to keep the total current constant in grid and plate circuits, there will be cyclic "ripple"

variations in each of these circuits as shown in Figure 8. As in Figure 5, X is the impressed potential while e_{L3} is the potential of the mid point of L_1 and represents the counter emf. of the oscillator. At points "a" the counter-electromotive force is zero.
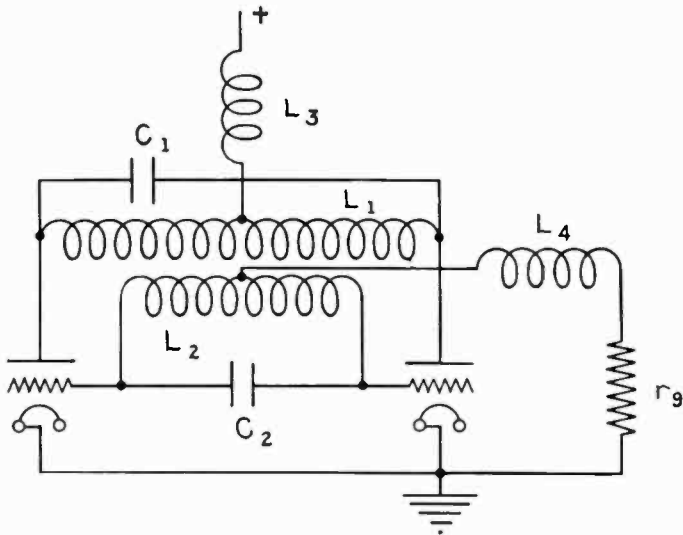
FIGURE 7
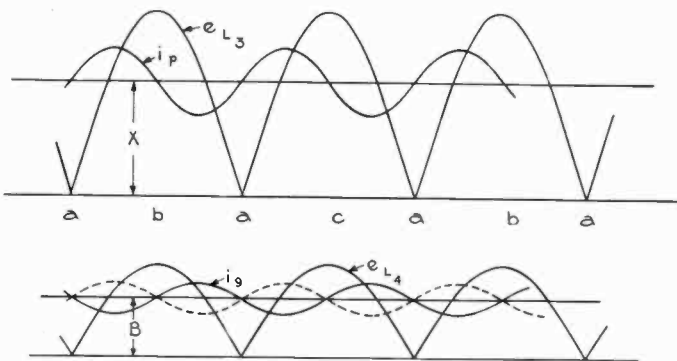
FIGURE 8

There is, therefore, nothing to prevent a rise in current except the inductance L_3. The current will therefore be rising at the points "a." To preserve the average, the counter emf. at times "b" and "c" is higher than the impressed potential X. The plate current i_p is, therefore, falling at "b" and "c."

By similar reasoning, e_{L4} is the emf. at mid-point of L_2, due to the rectifier action of the grids. Since this potential is depended

upon to make the grid current flow and since the potential is zero at points "a," the grid current is falling at these points and rising at points "b" and "c." The amplitude of the variations is dependent upon inductance $L_\circ$. The point of importance is that while i_p is rising, i_g is falling, and vice versa. In Figure 9 the tube losses for various values of grid voltage and plate current are shown. The corresponding plate voltage and grid current are found from the characteristic of Figure 6 from which Figure 9 was derived.

Suppose the required plate current for a given load to be 0.7 amperes, then the most economical grid voltage is 520, and the plate voltage is about 750, and the grid current about 0.225. The circuit
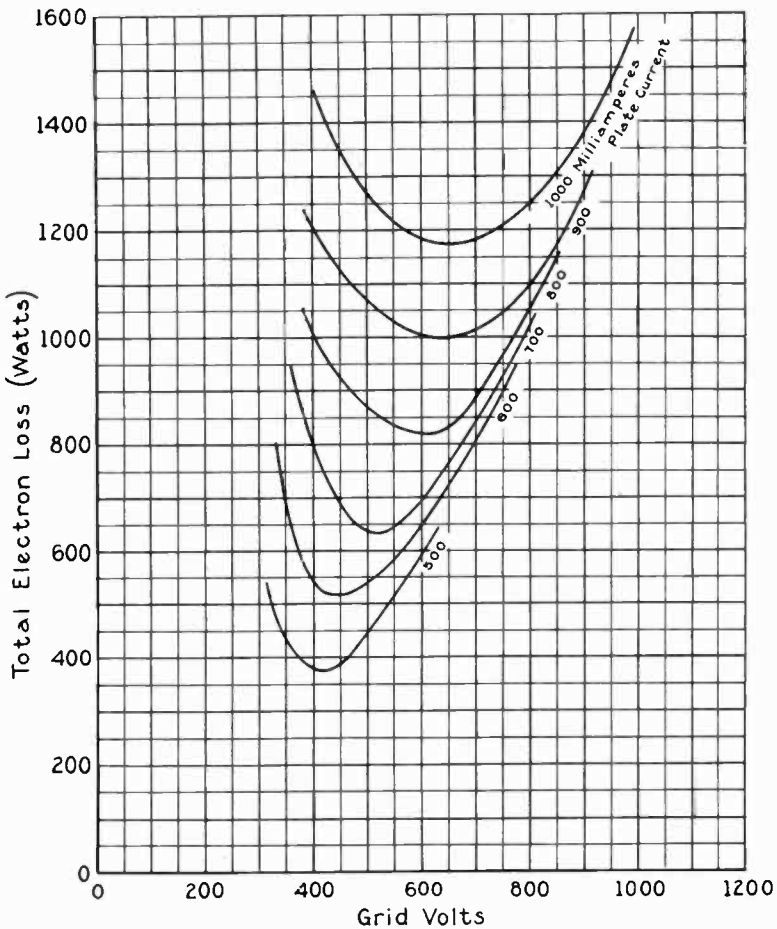
FIGURE 9—Space Charge Loss of the UV-206 Radiotron, the Characteristics of which are shown in Figure 6

of Figure 7 would logically be designed so that these currents and voltages would be obtained. But if, for some reason, the currents vary, as shown in Figure 8, the plate current rising to, say 1 ampere, and the grid current falling to, say, 0.150, the plate potential must rise to nearly 1,200 volts, giving a loss of 1,275 watts as compared with 630 under the correct conditions.

On the corresponding down swing of plate current to 0.4 and up swing of grid current to 0.3, the loss falls to 270. The average loss is thus considerably higher if the values of L_3 and L_4 are such as to permit a material change in plate and grid currents during the cycle. It is therefore important with this circuit to keep variations in both plate and grid currents to a minimum.

With high frequency and high voltage, the design of inductances L_3 and L_4 becomes increasingly difficult. For instance, if the power factor of the circuit $L_1 C_1$ is 10 percent, the energy current is one-tenth of the circulating current. The direct current is roughly equal to this energy current, so that if the current ripple in the plate circuit is not to exceed 10 percent of the direct current, then the ohmic impedance of L_3 must be 100 times L_1. The lowest frequency impressed upon L_3 is twice the fundamental, so that L_3 must be 50 L_1 in henrys. At radio frequencies and with apparatus of reasonably small size, it is almost impossible to design the inductance L_3 so that its natural resonant frequency will be sufficiently above the frequency at which it is required to work. Such a circuit as that in Figure 7 is thus practically limited to a few hundreds of cycles.

Where conditions are such that the circuit of Figure 7 can be used, it is very desirable because it combines high efficiency with high output. The test of suitability is to find what inductance L_3 must have to keep the ripple within satisfactory limits and then see if it is possible to build such an inductance.

Rather than develop the design of the foregoing circuit in greater detail, a circuit suitable for a somewhat higher frequency will be described. The design of that circuit will include the work necessary to design such a circuit as that of Figure 7. Figure 10 is the circuit in question.

The arrangement in Figure 10 is precisely the same as that in Figure 7, except that L_3 and L_4 are magnetically coupled. By this means the voltages induced by the plate current ripple in L_3 may be utilized to cause a grid current similar to the plate current to flow. The grid current ripple i_g in Figure 8 is thus shifted 180° to the position shown dotted. In this way it is possible, even with a considerable ripple, to conform constantly to the conditions

of minimum tube loss as obtained from Figures 6 and 9.

L_3 and L_4 are either actually connected thru a capacity as C_3, or the direct current power supply acts as such a condenser. If there should be no oscillations in C_1 L_1 and C_2 L_2, it is apparent that L_3 and L_4 may function as the plate and grid inductances of an ordinary Hartley circuit, the capacity coupling of which is furnished by the capacities between the coils representing L_1 and L_2 and such other parts of the main oscillating circuits as are exposed to one another. This circuit may support parasitic oscillations in which the two tubes operate in parallel and the frequency is the natural frequency of the circuits for this condition. The remedy for such parasitic oscillations is to have the fundamental circuit the lower decrement circuit of the two.
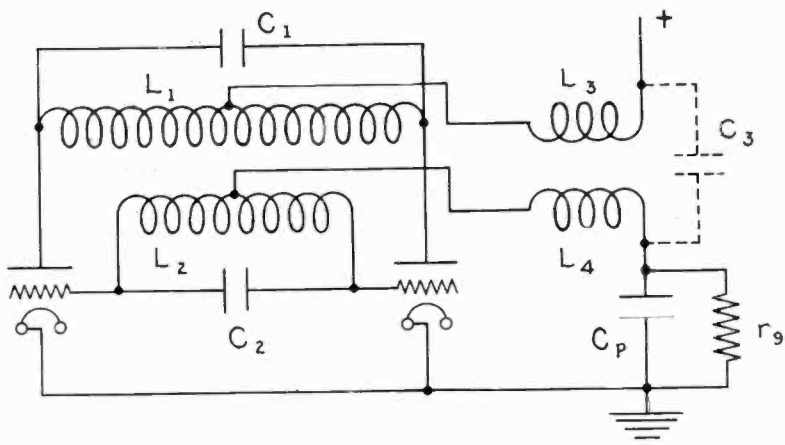
FIGURE 10

If this is not naturally the case, it may be made so by placing some resistance in series with C_p. The stored energy in the parasitic circuit with good design is small and its decrement is therefore fairly high. Another cause which tends to produce parasitic currents in any circuit is insufficient emission, and this must also be avoided.

The following sections of this paper are devoted to a discussion in detail of the design of the circuit Figure 10, with numerical example carried thru to illustrate how the various steps are taken.

RATING OF CIRCUIT

The rating of a circuit of the double-tube, square-wave type is very easy to predetermine. A given type of tube will be able

safely to withstand the stresses caused by a certain limiting direct voltage and the temperatures due to the passage of a certain current thru it for half of each cycle. The product of this voltage and current is the input from the direct current source. The losses occurring in the tubes are readily determined, and those occurring in the coil systems may be estimated. Hence the output rating is quickly obtained.

Distribution of Energy Thruout Circuit

The ratio of the energy stored in an oscillating circuit to the energy dissipated per cycle is $\frac{1}{2\pi}$ times the ratio of reactive volt-amperes to watts. Hence, if it is desired to store twice as much energy in a circuit as is dissipated per cycle, the volt-amperes must be 4π times the watts. This has been found to be a very good criterion for the lower limit in the design of simple oscillating circuits, and it is doubtful if it is wise to make any reduction here; for altho energy is now introduced into the circuit in two pulses per cycle instead of one, it is more essential that the voltages thruout the circuit hold to a definite shape, and this is accomplished by utilizing the fairly pure sine wave obtained with moderate energy storage.

The volt-amperes given by the relation just discussed is the total for the circuit. Part of this is in the primary circuit, part is in the secondary circuit, and part is held by the two circuits in common. Another way of stating this is to say that the volt-amperes in the two condensers represent the total energy, and their sum is equal to the sum of the volt-amperes in the two coils, due to their self inductances plus a term involving both currents and the mutual inductance.

As the current required to commutate the grids is drawn from the condenser in the secondary circuit, it is obvious that this is one point in the circuit where it would be well to store an amount of energy quite large with respect to the local energy losses. Several circuits have been designed using 30-volt-amperes in the grid condenser for each watt lost at this point, and this seems to be a very good value. It certainly is generous, and at the same time does not ordinarily tie up too large a portion of the total energy.

Voltages in Plate and Grid Coils

The ratio of average to peak voltage for a double wave rectifier is $\frac{2}{\pi}$. The grid circuit is obviously an ordinary recti-

fier, and the plate circuit is the same circuit reversed for the reception of energy at constant potential.

Due to the action of the grid choke, the voltage applied to the grid resistance is the theoretical average voltage minus the voltage drop in the grids themselves. Similarly, the plate choke takes the alternating voltage in the plate coil and presents it to the direct voltage source of energy as a steady counter electromotive force to be overcome. The input voltage must, of course, be equal to this counter emf., plus the plate drop.

Let E_p = plate drop under full load current.

E_g = grid drop under full load current.

E_1 = alternating voltage across primary coil (r. m. s. value).

E_2 = alternating current across secondary coil (r. m. s. value).

E_r = voltage applied to grid leak resistance (r_g).

E_c = voltage of constant potential source.

Then, remembering that only one-half the voltage of either coil is used at a time,

$$E_c = \frac{1}{2} \times \frac{2}{\pi} \times \sqrt{2}\, E_1 + E_p = \frac{\sqrt{2}}{\pi} E_1 + E_p \tag{1}$$

and

$$E_r = \frac{\sqrt{2}}{\pi} E_2 - E_g \tag{2}$$

It will be observed that these values are not entirely exact, as commutation of current from one tube to the other does not occur instantly at the time of zero coil voltage and other slight variations occur, but the values obtained will be found to be of quite sufficient accuracy for all practical needs.

The direct voltage of the supply source will, of course, be known, and this will determine E_1. The values of E_2 and E_r are not fixed until some assumption is made regarding the time required for the grid current to be transferred from one tube to the other. In general, this period will be short, but if it is made too short E_r will be found to be so high as to cause unwarranted losses in the grid resistance. The whole problem of commutation of the grid currents is quite elaborate and will be discussed later. For the present it will be sufficient to assume that commutation occupies a time interval represented by an angle θ to either side of the time when the voltage across the grid condenser is zero. See Figure 11. When commutation starts there is the drop E_g in one grid and zero drop across the other. The total voltage between grid terminals measured thru the grids is E_g.

At the same time, this voltage as measured thru the grid condenser or coil is $\sqrt{2}\,E_2 \sin\theta$, and, as these two must be equal,

$$E_g = \sqrt{2}E_2 \sin\theta \tag{3}$$

As E_g will be known from the tube characteristics, the value of E_2 is determined when θ is assumed. As the volt-amperes in the primary and secondary condensers have already been determined, and the voltages are now known, the currents are easily obtained, thus giving the terminal conditions to be met by the plate and grid coils, which together form an air cored transformer.

Calculation of Coupling Between Plate and Grid Circuits

As the terminal conditions of the plate and grid coils are known, and if the coefficient of coupling be assumed, it will be found that all voltages and currents thruout this part of the circuit may be readily calculated. Since the maximum possible coefficient of coupling of such a transformer can be estimated from its volt-ampere and voltage ratings with a fair degree of accuracy for a given type of construction after a little experience, this is the logical factor to use as a basis of calculation. A high coefficient of coupling is desirable, as it makes for easy energy interchange between the circuits and prevents any appreciable phase differences. (This latter point is more fully covered in another section.) A very high coefficient of coupling, however, is obtained only at the expense of the insulation between coils. In general, it is usually found that a coefficient of coupling between 0.35 and 0.50 will give satisfactory results.

Proceeding to the method of calculation, let

$P_1 = I_1{}^2 \omega L_1$ be the wattless power in the primary, due to its own current and inductance.

$P_2 = I_2{}^2 \omega L_2$ be the wattless power in the secondary, due to its own current and inductance.

$C_2 = I_1 I_2 \omega M_o$ be the wattless power which is held between the two circuits.

$Q_1 = I_1{}^2/\omega C_1$ be the total primary volt-amperes.

$Q_2 = I_2{}^2/\omega C_2$ be the total secondary volt amperes.

$$\tag{4}$$

Since the only real power transferred thru the coupling is that utilized in the grids, and this is very small in comparison with the wattless volt-amperes involved, and since the coefficient of coupling is high, the primary and secondary currents may be as-

sumed to be in phase in calculating their magnitudes; which will also make the primary and secondary voltages in phase.
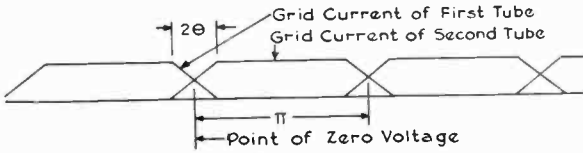
FIGURE 11

In order to fix the phase relationships definitely in mind, so as to avoid getting any term reversed in sign, let us refer to Figure 12. Current enters the two coils at a_1 and a_2 in the same
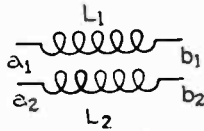
FIGURE 12

phase, and let us assume that at some instant the current in L_1 is developing its maximum effort to make a_1 positive with respect to b_1. At the same instant let the current in L_2 also be developing its maximum effort to the same end. Then both currents at the same instant are also exerting their peak effort to make a_2 positive with respect to b_2. Let these be the conditions assumed for positive values of P_1, P_2, C, and so on. Then

$$P_1 + C = Q_1 \tag{5}$$

$$P_2 + C = Q_2 \tag{6}$$

$$P_1 P_3 = \frac{C^2}{K^2} \tag{7}$$

where K is the coefficient of coupling. This last equation is obtained as follows:

$$M_o = K\sqrt{L_1 L_2} \quad \text{(by definition)}$$

$$I_1 I_2 \omega M_o = K\sqrt{I_1^2 \omega L_1 \cdot I_2^2 \omega L_2}$$

$$C = K\sqrt{P_1 P_2}$$

$$P_1 P_3 = \frac{C^2}{K^2}$$

Multiplying (5) by (6)

$$P_1 C + P_2 C + P_1 P_2 + C^2 = Q_1 Q_2 \tag{8}$$

Substituting from (7)

$$P_1 C + P_2 C + \left(1 + \frac{1}{K^2}\right) C^2 = Q_1 Q_2 \tag{9}$$

Multiplying (5) by C

$$P_1 C = Q_1 C - C^2 \tag{10}$$

Multiplying (6) by C

$$P_2 C = Q_2 C - C^2 \tag{11}$$

Substituting (10) and (11) in (9)

$$(Q_1 + Q_2) C + \left(\frac{1}{K^2} - 1\right) C^2 = Q_1 Q_2$$

$$\left(\frac{1}{K^2} - 1\right) C^2 + (Q_1 + Q_2) C - Q_1 Q_2 = 0$$

$$C = \frac{-(Q_1 + Q_2) \pm \sqrt{(Q_1 + Q_2)^2 + 4 Q_1 Q_2 \left(\frac{1}{K^2} - 1\right)}}{2\left(\frac{1}{K^2} - 1\right)} \tag{12}$$

Inspection of the original equations will show that P_1 and P_2 may now be readily determined, thus giving the values of L_1, L_2, and M_o. As C_1 and C_2 (primary and secondary capacities) and r_g are easily derived from the results of the preceding paragraph, there remains only the load resistance, pass condenser for r_g, and inductance in the harmonic transformer to be determined.

If the circuit is loaded as shown in Figure 13, it is only necessary in calculating r_L to pick a resistance which will absorb the output power at the known current I_1.
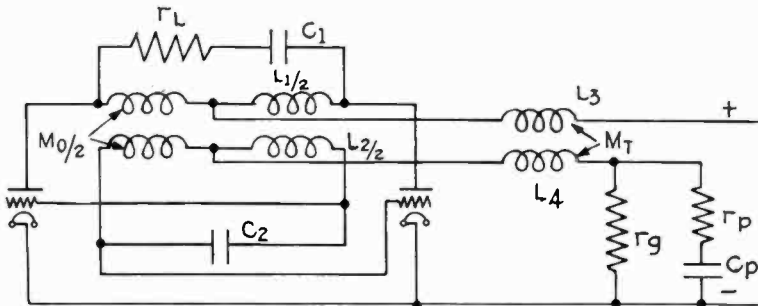
FIGURE 13—Diagram of Square Wave Circuit showing Nomenclature

The function of C_p is to pass the small amount of alternating current which gets thru the harmonic transformer so that it does not have to go thru r_g. The corresponding current ~~which~~ flows

to the power supply thru L_3 will undoubtedly be short-circuited in this manner and it is desirable to have both paths of this same nature. The value of C_p is not critical; it must merely present a relatively low impedance to a double-frequency current of magnitude equal to a fraction of the grid current. The function of r_p in series with it is to discourage parasitic oscillations; its value is best found by trial.

THE HARMONIC TRANSFORMER

The function of the harmonic transformer is to pass the steady direct currents from the power source and grid leak resistance and to hold up the double frequency and its harmonic voltages which always appear wherever rectification takes place. Now it is impossible to do this without drawing some exciting current and it will be found that this exciting current must be quite large if bad internal tuning effects are to be avoided. If separate chokes are used in the two circuits the plate ripple will be increasing the plate current while the grid ripple will be decreasing the grid current. This is a condition giving bad additional losses in the tubes which could be reduced almost to zero if the two currents rose and fell together. This latter state may be obtained by coupling the two chokes together into a transformer and making the plate ripple current drive the grid current ripple in the right phase relation.

Before advancing any further, it is well to see what the magnitudes of the voltages are which it is proposed to hold back. The fundamental component of the ripple (that is the double output frequency component) for double wave rectifiers is $\dfrac{2}{3\pi}$ times the voltage across the outer terminals of the alternating current winding. Hence, neglecting all other components, and designating by E_3 and E_4 the voltages across L_3 and L_4 (see Figure 13),

$$E_3 = \frac{2}{3\pi} E_1 \quad \text{and} \quad E_4 = \frac{2}{3\pi} E_2 \tag{13}$$

As a further approximation assume that all the impedance presented to these voltages is that located in the harmonic transformer. The ripple currents will be taken as a certain percent of the direct currents. Their relative magnitude will depend on the particular circuit used. Some circuits due to relatively high direct currents and low voltages can be run with smaller ripples than others. The limit occurs due to internal resonance of the coils.

We now have a transformer with given terminal conditions to be met, and the solution is the same as for the transformer in the oscillating circuit. Upon examination of the relative polarities of the voltages and currents, however, it is discovered if the previous nomenclature is held to, that the actual numerical value of the voltage applied to L_4 will be negative, whereas, before, everything had a positive numerical value. This has an important practical bearing on the problem as it puts a limit on the coefficient of coupling.

Let X_3 = reactance of coil L_3.
X_4 = reactance of coil L_4.
X_M = mutual reactance.
I_3 = alternating current thru L_3.
I_4 = alternating current thru L_4.

Then

$$I_3 X_3 + I_4 X_M = E_3 \tag{14}$$

$$I_4 X_4 + I_3 X_M = E_4 \tag{15}$$

$$X_3 X_4 = \frac{X_M{}^2}{K^2} \tag{16}$$

Multiplying (14) and (15) together and substituting from (16):

$$I_3{}^2 X_3 X_M + I_4{}^2 X_4 X_M + I_3 I_4 X_M{}^2 \left(1 + \frac{1}{K^2}\right) = E_3 E_4 \tag{17}$$

Multiplying (14) by $I_3 X_M$ and transposing

$$I_3{}^2 X_3 X_M = E_3 I_3 X_M - I_3 I_4 X_M{}^2 \tag{18}$$

Multiplying (15) by $I_4 X_M$ and transposing

$$I_4{}^2 X_4 X_M = E_4 I_4 X_M - I_3 I_4 X_M{}^2 \tag{19}$$

Substituting (18) and (19) in (17)

$$I_3 I_4 \left(\frac{1}{K^2} - 1\right) X_M{}^2 + (E_3 I_3 + E_4 I_4) X_M - E_3 E_4 = 0 \tag{20}$$

From which

$$X_M = \frac{-(E_3 I_3 + E_4 I_4) \pm \sqrt{(E_3 I_3 + E_4 I_4)^2 + 4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)}}{2 I_3 I_4 \left(\frac{1}{K^2} - 1\right)} \tag{21}$$

In order that X_M may be a real number

$$(E_3 I_3 + E_4 I_4)^2 > -4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)$$

$$E_3{}^2 I_3{}^2 + E_4{}^2 I_4{}^2 + 2 E_3 E_4 I_3 I_4 > -4 E_3 E_4 I_3 I_4 \left(\frac{1}{K^2} - 1\right)$$

$$E_3{}^2 I_3{}^2 + E_4{}^2 I_4{}^2 - 2 E_3 E_4 I_3 I_4 > -\frac{4 E_3 E_4 I_3 I_4}{K^2}$$

Now the left-hand side is the square of a real number and hence positive. Also K^2 must be positive. Hence:

$$K^2 > \frac{-4\,E_3\,E_4\,I_3\,I_4}{E_3{}^2\,I_3{}^2 + E_4{}^2\,I_4{}' - 2\,E_3\,E_4\,I_3\,I_4} \tag{22}$$

The limits on K are now known. If it is impossible or impracticable to build a set of coils having a coefficient of coupling within these values, it will be necessary to give new values to some of the assumed constants on which the circuit as been calculated. Changing the assumed ripples in the plate and grid currents will change only the actual values of X_3, X_4, and X_M, leaving their relative values unchanged. This is not apt to give any relief in the situation and earlier assumptions must be scrutinized, and of these the only ones which can be changed to advantage are those affecting the voltage in the grid circuit, for if this can be lowered, the requirements imposed on the harmonic transformer will be easier to meet as will be seen by examination of equation (22). As soon as a practical value of K is determined upon, the values of X_M, X_3, and X_4 follow from equations (21), (14), and (15), taking them in the order named.

COMMUTATION OF GRID CURRENTS

Among the assumptions which were made earlier in this article was one to the effect that transfer of grid currents would be accomplished as tho this transfer had no effect on the voltage which would otherwise exist in the secondary circuit. It is now proposed to examine into the truth of this assumption. It is not practical to go very far into the investigation, however, so it will merely be assumed that during a short period of commutation, the current in the secondary coil remains constant, which will be very nearly the case, and the first assumption verified on this basis.

$i_o =$ maximum of current flowing thru inductance L_2 and assume a constant current of this value to flow thru L_2 during commutation. (This is in addition to the direct components of the grid currents.)

$i_1 =$ current in grid from which current is being transferred.

$i_2 =$ current in grid to which current is being transferred.

$i_g =$ steady value of grid current during the active part of the cycle. (This assumes complete smoothing by the harmonic transformer.)

$r =$ resistance of two grids in series

Then it can be shown that

$$i_2 = i_o \left(1 - \varepsilon^{-\frac{t}{rC_2}}\right) + \frac{i_g}{2}\left(1 - \varepsilon^{-\frac{t}{rC_2}}\right) = \left(i_o + \frac{i_g}{2}\right)\left(1 - \varepsilon^{-\frac{t}{rC_2}}\right) \quad (23)$$

and
$$i_1 = -i_o\left(1 - \varepsilon^{-\frac{t}{rC_2}}\right) + \frac{i_g}{2}\left(1 + \varepsilon^{-\frac{t}{rC_2}}\right) \quad (24)$$

By setting i_2 equal to i_g it is possible to calculate the time at which transfer is completed, and in the usual case no appreciable difference is shown between this value and that given by the first assumption.

Losses in Tubes During Commutation

The loss in the plate circuit during commutation may be unduly increased by using too large a lap angle. This is a matter which probably cannot be calculated accurately, because of the lack of exact knowledge of many things which happen in an extremely short interval of time, but, if certain reasonable assumptions are made, at least the order of magnitude of the additional losses may be estimated. First let it be assumed that there is no phase displacement between the grid excitation and the plate voltage under which conditions commutation is effected by the grids. Then at the start and finish of the period of commutation, the tube which is carrying the load current should have only the voltage drop across it actually required to pass the current. The other tube will have the remainder of the voltage of the primary circuit impressed upon it. While it is extremely difficult to say exactly how the voltage will divide between the tubes during commutation the most likely division will be of the nature shown in Figure 14.
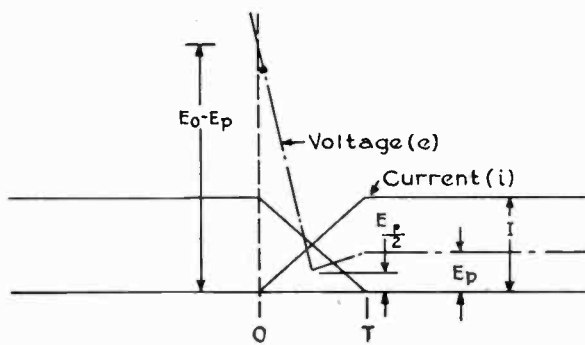
FIGURE 14

Let E_o be the voltage in the primary circuit at either end of the period of commutation.

E_p be the steady plate drop while either tube is carrying all the load current.

e be the instantaneous voltage across the tube into which the current is being commutated.

i be the instantaneous current thru the same tube.

Then assume that

$$i = \frac{I}{T} t$$

$$e = (E_o - E_p) - \frac{2\left(E_o - \frac{3}{2}E_p\right)}{T} t \quad \text{for} \quad 0 < t < \frac{T}{2}$$

and

$$e = \frac{E_p}{T} t \quad \text{for} \quad \frac{T}{2} < t < T$$

Measuring t in degrees and calling the losses in both tubes W_1

$$W_1 = \frac{4}{360} \left\{ \frac{(E_o - E_p) I}{T} \int_o^{\frac{T}{2}} t \, dt - \frac{2(E_o - \frac{3}{2}E_p) I}{T^2} \int_o^{\frac{T}{2}} t^2 dt + \frac{I E_p}{T^2} \int_{\frac{T}{2}}^{T} t^2 \, dt \right\}$$

$$= \frac{I T}{2,160} \left\{ E_o + 7 E_p \right\}$$

Now let $W_2 =$ watts lost in one tube during time $2\,T$ if it carried the steady current I. Then

$$W_2 = \frac{2\,T}{360} \times I \times E_p$$

When $W_1 = W_2$ there are no additional losses due to commutation

$$\frac{2 E_p}{360} = \frac{E_o + 7 E_p}{2,160}$$

or

$$E_o = 5 E_p$$

and if $E_o < 5 E_p$ it is unlikely that there will be any serious losses due to the commutation even tho our assumptions may not be obtained in practice.

Vector Diagram of Circuit

Up to this point the phase displacement between the grid excitation and plate voltage has been assumed to be zero. While it is usually very small it is often desirable to know its value. This can be determined by referring to Figure 15. The subcripts refer to the different parts of the circuit shown in Figure 13. The root-mean-square components of plate and grid currents are designated I_p and I_g.

Starting with I_{L_2} we know that E_2 leads it by $90° + \theta_1$ in order that this coil may be a source of power for the secondary. Icr leads E_2 by $90°$ and I_g must have such a value that the vector sum of I_{L_2}, Ic_2 and I_g is zero. E_2 consists of the voltage induced in L_2 by its own current ($I_{L_2} X_{L_2}$), the voltage induced in L_2 by the primary current ($I_{L_1} X_{M_0}$), and the resistance drop in L_2.
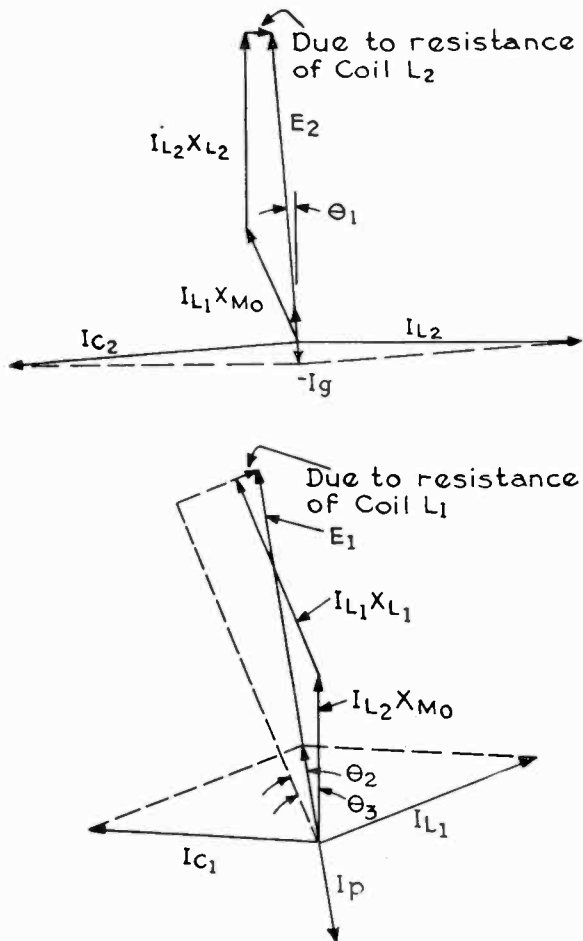
Figure 15—Vector Diagram of Circuit shown in Figure 13

In the primary circuit we have its own current I_{L_1} and the voltage induced in the primary coil by the secondary current ($I_{L_2} X_{M_0}$). Adding to the latter the voltage induced in the primary coil by its own current ($I_{L_1} X_{L_1}$) and the resistance drop in the same coil gives E_1, the primary voltage. I_p must then be

180° out of phase with E_1 in order to represent input to the circuit and Ic_1 must combine with I_{L_1} and I_p to give a vector sum of zero.

Let $Q_1 =$ total primary reactive volt-amperes.

$Q_2 =$ total secondary reactive volt-amperes.

$C =$ reactive volt-amperes common to both circuits.

$P_g =$ watts lost in grids.

Then if the resistance of the coils themselves be neglected

$$\left.\begin{array}{l} \theta_1 = tan^{-1}\dfrac{P_g}{Q_2} \\[2mm] \theta_2 = tan^{-1}\dfrac{P_g}{Q_1} \\[2mm] \theta_3 = tan^{-1}\dfrac{P_g}{C} \end{array}\right\} \qquad (25)$$

and the angle ϕ by which the grid excitation leads the plate voltage is given by

$$\phi = \theta_1 + \theta_2 - \theta_3 \qquad (26)$$

PRODUCTION OF LEAD ANGLE

In circuits using soft tubes, that is tubes which have control only to the extent that current flow may be held back if once stopped by external sources, it is necessary that the grid excitation shall lead the plate voltage in phase by such an amount that commutation is completed before the point of reversal of the commutating voltage is reached. This may be accomplished by placing a condenser in each plate lead bridged by a choke of high inductance to pass the direct component of plate current.

The displacement of voltages produced by such a device may be seen by referring to Figure 16. The tubes are here represented as a source of alternating current power feeding the oscillating circuit which, being tuned, may be represented as a pure resistance. The circuit is completed by the two phasing condensers. That part consisting of the oscillating circuit and phasing condensers is so drawn that the lengths and directions of the different impedances are representative of the voltages across the parts.

The grid voltage will be practically in phase with the voltage across the oscillating circuit; and any difference may be calculated by the method just shown. Hence the difference in phase between grid excitation and plate voltage is known as soon as the impedances are known; or, if the effective resistance of the oscillating circuit is known, the size of the condensers required to produce a given phase displacement may be calculated. Since

the voltage across the oscillating circuit and the power output are known, its effective resistance is readily calculated.

Osc. Cir.

Grid Voltage

Plate Voltage
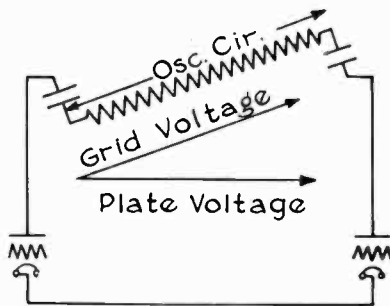
FIGURE 16

EXAMPLE ILLUSTRATING METHOD OF CALCULATION

The circuit to be calculated is to use two General Electric Company 50-watt tubes having "XL" filaments and is to draw 500 milliamperes from a 1,000 volt direct current source for input. The frequency is to be 12,000 cycles. Investigation of the tubes show that they will pass 500 milliamperes plate current with a plate drop of about 165 volts, a grid drop of about 125 volts, and a grid current of 85 millamperes. This gives:

Applied voltage (E_c)		$= 1,000$ V.
Plate drop under full load current	$(E_p) =$	165 V.
Grid drop under full load current	$(E_g) =$	125 V.
Input current	$(I) =$	500 ma.
Grid current	$(i_g) =$	85 ma.
Input		$= 500$ watts

Assuming 80 percent efficiency, the output will be 400 watts, and as the reactive power is to be 4π times the watts, there will be 5,000 volt-amperes in both circuits.

From equation (1) the voltage in the primary circuit (E_1) is

$$E_1 = \frac{\pi}{\sqrt{2}} (E_c - E_p)$$

$$= \frac{\pi}{\sqrt{2}} (1,000 - 165) = 1,850 \; volts \;\; (r.m.s.)$$

Let it be assumed that 10° will be sufficient time for commutation in the grid circuit. Then the r.m.s. voltage in the secondary circuit (E_2) is given by equation (3)

$$E_2 = \frac{E_g}{\sqrt{2} \sin\theta} = \frac{125}{\sqrt{2} \times \sin 5°} = 1,015 \; volts \;\; (r.m.s.)$$

and the voltage across the grid resistance (E_r) is given by equation (2).

$$E_r = \frac{\sqrt{2}}{\pi} E_2 - E_g$$

$$= \frac{\sqrt{2}}{\pi} \times 1,015 - 125 = 333 \; volts.$$

At 85 ma., this represents 28 watts loss, and the impedance of the grid resistance is

$$r_g = \frac{333}{0.085} = 3,920 \; ohms.$$

If it is assumed that 2,000 volt-amperes out of the total of 5,000 will be in the secondary circuit, the losses in the secondary circuit will be

Loss in grids $2 \times 125 \times 0.085$	$= 21$ watts
Loss in circuit due to resistance (at 1 percent power factor)	$= 20$ watts
Loss in r_g	$= 28$ watts
Total loss in secondary circuit	$= 69$ watts

This gives thirty volt-amperes in the secondary circuit for every watt lost, which appears to be a good value.

We can now show that commutation will occur in substantially the 10° assumed. This may be accomplished by the use of equation (23).

$$i_o = \sqrt{2} \times \frac{2,000}{E_2} = 2.79 \; amperes$$

$$\frac{1}{2 \pi f c_2} = \frac{E_2}{I_2} = \frac{1015}{0.707 \times 2.79}$$

from which $C_2 = 2.57 \times 10^{-8}$ microfarad

$$r = \frac{2 E_g}{I_g} = \frac{2 \times 125}{0.085} = 2,940 \; ohms.$$

Substituting in equation (23) for the finish of transfer.

$$i_2 = \left(i_o + \frac{i_g}{2} \right) \left(1 - \varepsilon^{-\frac{t}{r C_2}} \right)$$

$$0.085 = (2.79 + 0.043) \left(1 - \varepsilon^{-\frac{t}{r C_2}} \right)$$

from which

$$\frac{t}{r C_2} = 0.0304$$

$$t = 0.0304 \; r \, C_2 = 0.0304 \times 2,940 \times 2.57 \times 10^{-8}$$

$$= 2.295 \times 10^{-6} \; second$$

$$12,000 \; cycles = 4.32 \times 10^6 \; degrees/second.$$

Therefore, the period of transfer covers
$$2.295 \times 4.32 = 9.9°$$

It has already been decided to divide the reactive power in the circuit so that there are 3,000 volt-amperes in the primary (Q_1) and 2,000 volt-amperes in the secondary (Q_2). Hence if the coefficient of coupling between primary and secondary (K) be assumed, the value of the reactive power common to both circuits (C) may be obtained from equation (12).

$$C = \frac{-(Q_1 + Q_2) \pm \sqrt{(Q_1 + Q_2)^2 + 4 Q_1 Q_2 \left(\frac{1}{K^2} - 1\right)}}{2 \left(\frac{1}{K^2} - 1\right)}$$

and if $\quad K = 0.50, \quad$ then $\quad \left(\dfrac{1}{K^2} - 1\right) = 3, \quad$ and

$$C = \frac{-5,000 \pm \sqrt{5,000^2 + 4 \times 3,000 \times 2,000 \times 3}}{2 \times 3}$$

$= +808$ or $-2,475$; and the positive value is the one desired in this case.

Since P_g the loss due to the grids is 69 watts, the phase displacement between grid and plate voltages may be calculated by equations (25) and (26).

$$\theta_1 = tan^{-1} \frac{P_g}{Q_2} = tan^{-1} \frac{69}{2,000} = 2°$$

$$\theta_2 = tan^{-1} \frac{P_g}{Q_1} = tan^{-1} \frac{69}{3,000} = 1°20'$$

$$\theta_3 = tan^{-1} \frac{P_g}{C} = tan^{-1} \frac{69}{808} = 4°55'$$

$$\phi = \theta_1 + \theta_2 - \theta_3 = -(1° 35')$$

Calculation of Constants of Main Oscillating Circuit

E_1, the primary voltage, has been shown to be 1,850 volts r.m.s.

E_2, the secondary voltage, has been shown to be 1,015 volts r.m.s. Since there are 3,000 volt-amperes in the primary, the current is

$$I_1 = \frac{3,000}{E_1} = \frac{3,000}{1,850} = 1,622 \ ampere \ r.m.s.,$$

and since there are 2,000 volt-amperes in the secondary, the current here is

$$I_2 = \frac{2,000}{E_2} = \frac{2,000}{1,015} = 1.97 \ ampere \ r.m.s.$$

There are 808 volt-amperes common to both circuits or

$$\omega\, M_o\, I_1\, I_2 = 808$$

$\omega = 75{,}400$ for 12,000 cycles, hence

$$\omega\, M_o = \frac{808}{1.622 \times 1.97} = 253 \; ohms$$

$$M_o = \frac{253}{75{,}400} = 3.36 \; milli\text{-}henrys.$$

The volt-amperes in the primary coil due to its self inductance
are

$$\omega\, L_1\, I_1{}^2 = 3{,}000 - 808 = 2{,}192$$

$$\omega L_1 = \frac{2{,}192}{(1.622)^2} = 833 \; ohms.$$

$$L_1 = \frac{833}{75{,}400} = 11.04 \; milli\text{-}henrys.$$

The volt-amperes in the secondary coil due to its self-inductance are

$$\omega L_2 I_2{}^2 = 2{,}000 - 808 = 1{,}192$$

$$\omega L_2 = \frac{1{,}192}{(1.97)^2} = 307 \; ohms$$

$$L_2 = \frac{307}{75{,}400} = 4.08 \; milli\text{-}henrys.$$

The volt-amperes in the primary capacity are

$$\frac{I_1{}^2}{\omega\, C_1} = 3{,}000$$

$$\frac{1}{\omega\, C_1} = \frac{3{,}000}{(1.622)^2} = 1{,}140 \; ohms.$$

$$C_1 = 0.01163 \; microfarad.$$

The volt-amperes in the secondary capacity are

$$\frac{I_2{}^2}{\omega\, C_2} = 2{,}000$$

$$\frac{1}{\omega\, C_2} = \frac{2{,}000}{(1.97)^2} = 516 \; ohms.$$

$$C_2 = 0.0257 \; microfarard.$$

The voltages across the harmonic transformer are given by
equations (13).

$$E_3 = \frac{2}{3\,\pi}\, E_1 = \frac{2}{3\,\pi} \times 1{,}850 = 394 \; volts \; r.m.s$$
$$= 557 \; volts \; (peak)$$

$$E_4 = \frac{2}{3\,\pi}\, E_2 = \frac{2}{3\,\pi} \times 1{,}015 = 215 \; volts \; r.m.s.$$
$$= 306 \; volts \; (peak)$$

and this latter will be negative in sign under the assumed system of nomenclature.

Let it be assumed that these voltages cause ripple currents to flow (I_3 and I_4) which have peaks equal to 20 percent of the direct currents. Hence

$I_3 = 0.2 \times I = 0.2 \times 0.5 = 0.1$ *ampere* *(peak)* $= 0.0707$ *ampere* *r.m.s.*

$I_4 = 0.2 \times i_g = 0.2 \times 0.085 = 0.017$ *ampere* *(peak)* $= 0.012$ *ampere* *r.m.s.*

Using equation (22) to find the minimum possible coupling gives

$$K^2 > \frac{-4}{\dfrac{E_3 I_3}{E_4 I_4} + \dfrac{E_4 I_4}{E_3 I_3} - 2}$$

$$K^2 > \frac{-4}{\dfrac{394 \times .0707}{215 \times .012} - \dfrac{215 \times .012}{394 \times .0707} - 2}$$

$$K^2 > 0.313.$$

This gives a value of 0.56 as the minimum for K, and substituting this value in equation (21) gives

$$X_M = \frac{-(E_3 I_3 + E_4 I_4) \pm \sqrt{(E_3 I_3 + E_4 I_4)^2 + 4 E_3 E_4 I_3 I_4 \left(\dfrac{1}{K^2} - 1\right)}}{2 I_2 I_4 \left(\dfrac{1}{K^2} - 1\right)}$$

$$= \frac{-(394 \times 0.0707 - 215 \times 0.012) \pm 0}{2 \times 0.0707 \times 0.012 \times (3.19 - 1)} = -6{,}780 \ ohms.$$

Knowing X_M, the values of X_3 and X_4, the reactances of coils L_3 and L_4, are obtained from equations (14) and (15)

$$X_3 = \frac{E_3 - I_4 X_M}{I_3}$$

$$= \frac{394 + 0.012 \times 6{,}780}{0.0707} = 6{,}720 \ ohms.$$

$$X_4 = \frac{E_4 - I_3 X_M}{I_4}$$

$$= \frac{-215 + 0.0707 \times 6{,}780}{0.012} = 21{,}900 \ ohms.$$

At 24,000 cycles (150,800 radians/seconds) these impedances give

$$M_T = -45 \ milli\text{-}henrys$$
$$L_3 = 44.6 \ milli\text{-}henrys$$
$$L_4 = 145.3 \ milli\text{-}henrys$$

Experimental Performance of Small Circuit

In order to show the probable variations between actual performance and performance as calculated, the circuit which has just been analyzed was set up in the laboratory. The tubes had been selected and tested before the calculations were started, and the actual and calculated values of the other parts of the circuit are given in Table I.

TABLE I

12,000 ∿ Square Wave Circuit Using 2-50 Watt "XL" Filament Tubes

Part	Calculated Value	Actual Value	Resistance
C_1	0.01163 mfd.	0.012 mfd.	. . . .
C_2	0.0257 mfd.	0.026 mfd.	. . . .
C_p	. . . .	2.0 mfd.	. . . .
L_1	11.04 milli-henrys	11.2 mh.	6 ohms
L_2	4.08 milli-henrys	4.0 mh.	3.33 ohms
L_3	44.6 milli-henrys	44.67 mh.	36.5 ohms
L_4	145.3 milli-henrys	146.6 mh.	119 ohms
M_o	+3.36 milli-henrys	+3.43 mh.	. . . .
M_T	−45 milli-henrys	−44.9 mh.	. . . .
r_L	. . . .	. . . .	. . . .
r_a	3,920 ohms	4,100 ohms	. . . .
r_p	. . . .	500 ohms	. . . .

The harmonic transformer tuned with its internal capacity plus the capacity between the primary and secondary coils of the oscillating circuit at about 78,000 cycles.

The performance of the circuit is shown in Figure 17. The output as measured along the axis of abscissas is the actual useful output in the resistance r_L. Now the calculations were based on 500 watts input, and from the experimental curves this corresponds to 329 watts output. Drawing a vertical line thru this value, and inserting on it the calculated values for the different currents and the tube losses shows that the performance is practically as calculated.

<div style="text-align: right">

Research Laboratory,
General Electric Company,
Schenectady, New York.

</div>

August 21, 1923.
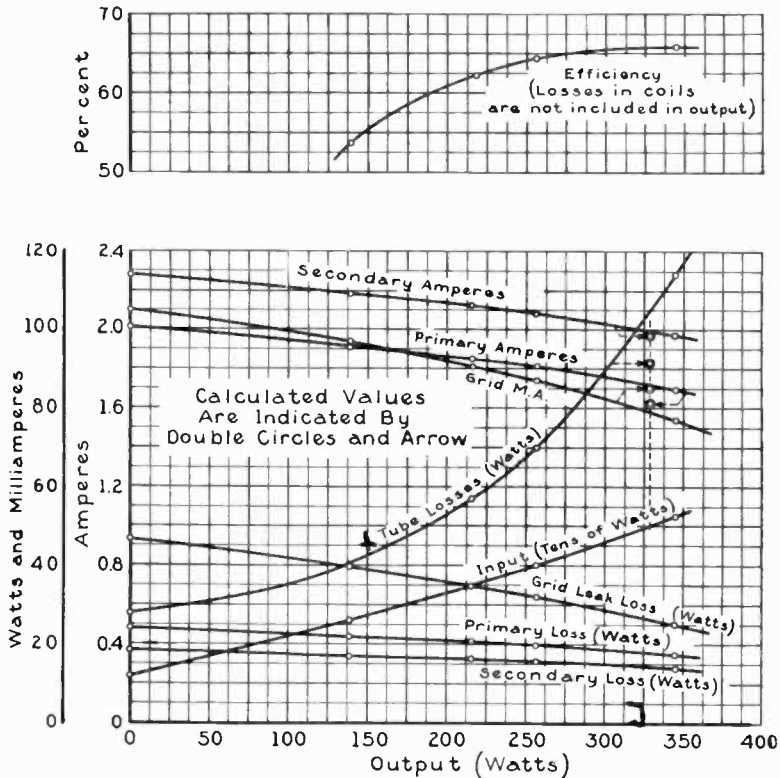
FIGURE 17—Performance of a Square-Wave Circuit with Constants given in Table 1 at 1,000 volts direct potential

SUMMARY: A modified push-pull circuit is described in which approximately square waves of current are passed thru the two tubes alternately. High efficiency is obtained by superimposing upon the grid circuit a voltage proportional to current variations in the plate circuit. The circuit gives high output combined with high efficiency without shortening the life of the tubes.

"ON THE CALCULATION OF THE INDUCTANCES AND CAPACITIES FOR A MULTI-RANGE . . . SERIES OF TUNED CIRCUITS . . . "—By J. ERSKINE MURRAY

By

RALPH R. BATCHER

Referring to the paper from J. Erskine-Murray in the August, 1924, issue, "On the Calculation of the Inductances and Capacities for a Multi-Range or Other Consecutive Series of Tuned Transmitting or Receiving Circuits, the Total Range and Accuracy Required Being Given," the writer has devised another method which has proven of practical value for several years, and which seems considerably simpler than the method outlined.

Since many of the present-day circuits require the design of a multi-range oscillating circuit which includes the antenna capacity and inductance, this problem will be considered first.

The first case will be that of a circuit tuned with a variable condenser having a maximum and minimum capacity of C_{max} and C_{min}, respectively.

The capacity of the antenna is designated C_a, and the maximum and minimum wave lengths by λ_{max} and λ_{min}.

$$R = \frac{C_{max}}{C_{min}}$$

$$S = \frac{C_{max}}{C_a}$$

$$T = \frac{\lambda_{max}}{\lambda_{min}}$$

All of these values are generally known at the start, except possibly S. In many cases the receiver is supposed to work with a number of aerials, in which case the writer generally uses the value $S = 4$, which is the practical limit.

The total number of taps on the inductance necessary can then be computed from the formula:

$$N = \frac{\log T}{\log 0.90 \sqrt{\dfrac{R+S}{1+S}}}$$

where the factor 0.90 insures a 10 percent overlap. The derivation is not difficult, but will not be given here.

Having determined the number of taps (N) required, the location of each is given from the formula:

The last tap $\quad\quad$ = Total inductance = L (max.)
Next to the last tap = $L_{max} \div x$
Second from last $\quad$ = $L_{max} \div x^2$
Third from last $\quad$ = $L_{max} \div x^3$
nth from last $\quad\quad$ = $L_{max} \div x^n$

The value for L_{max} can be determined from

$$L_{max} = \lambda^2 \frac{1+S}{3,552 \, C_{max}}$$

In some cases in computing the two lowest taps, it may be desirable to consider the inductance of the aerial. In most cases sufficient accuracy will be obtained if the equivalent concentrated inductance of the aerial is subtracted from the values of inductance obtained by the above formulas. A number of cases have been solved using a more exact formula, but the additional accuracy obtained rarely warranted the extra amount of work. The simpler case, of where the antenna is not included in the oscillating circuit, corresponding to the problem considered by Mr. Erskine-Murray, is solved in a similar manner.

Such a circuit might, for example, be a secondary circuit of which the above coil is the primary or a wavemeter. Naturally the same wave length bands must be covered.

$$\text{Then} \quad \frac{L_{sec}}{L_{pri}} = \frac{M}{S+1} \quad \text{where}$$

M is the ratio of maximum condenser capacities of each circuit, that is,

$$M = \frac{C_{pri}}{C_{sec}}$$

The number of taps is given by the formula

$$N_s = \frac{log \, T}{log \, 0.90 \sqrt{\dfrac{R \, D+1}{D+1}}}$$

and the locations of the taps are determined in the same way described except that now

$$x_s = 0.90 \sqrt{\frac{R \, D+1}{D+1}}$$

where R is the condenser capacity ratio as before, and D is the

ratio of the minimum capacity of the condenser to the distributed capacity of the attached circuit, such as that due to leads, coils, vacuum tubes, and so on.

Using this method for locating the taps, the wave lengths will only be found in one location (except at the ends) and the minimum number of taps possible is used, so that tuning is much easier. In the broadcast receiver field, it may be mentioned, the use of the "straight line frequency" condenser simplifies tuning to a remarkable degree since all frequency bands are separated by an equal amount of dial displacement. The accuracy of tuning, outlined by Mr. Erskine-Murray, is constant over the whole range.

August 8, 1924.

ISSUED JUNE 24, 1924—AUGUST 19, 1924

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,505,055—A. R. Molins, filed May 19, 1922, issued August 12,
1924.

Fig.2.   *Fig.2ª*
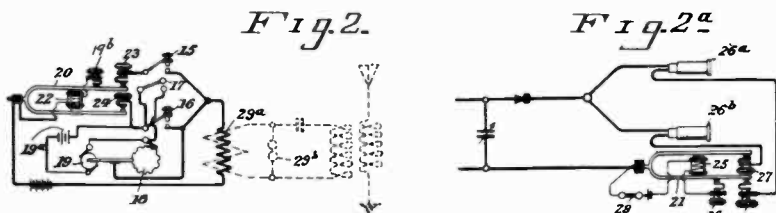
NUMBER 1,505,055—Secret System for Radio Telegraphy

SECRET SYSTEM FOR RADIO TELEGRAPHY, wherein a tuning
fork is provided at both the transmitting and receiving stations
and arranged to vibrate in synchronism to close sets of contacts
connected in different circuits whereby one series of signals may
be radiated between the spaces of another series of signals. The
messages are therefore transmitted in mixed relation and sepa-
rated by a synchronized device at the distant receiving station.

The following patent of interest issued in 1922 has been added
to this digest:

1,438,828—H. W. Houck, filed March 29, 1920, issued December
12, 1922.

METHOD AND APPARATUS FOR SELECTIVELY TRANSFERRING
ELECTRICAL OSCILLATORY ENERGY, of any frequency or fre-
quencies lying within a continuous band of frequencies from one
electrical system to another. This patent shows an electron tube

*Received by the Editor, September 12, 1924. Readers will note that the
first patent, 1,505,055, was issued August 12, 1924, and inadvertently omitted
from an earlier digest.

amplifier in which the input and output circuits of the several tubes are coupled by means of a series of oscillatory circuits tuned to different frequencies. Each circuit is resonant to a different frequency so that the circuit has a highly efficient collective range of resonance which includes the band of frequencies.
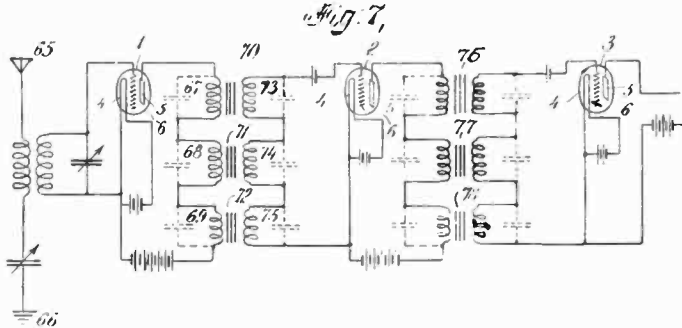
NUMBER 1,438,828—Method and Apparatus for Selectively Transferring Electrical Oscillatory Energy

1,498,536—S. N. Baruch, filed August 2, 1920, issued June 24, 1924.

APPARATUS FOR PRODUCING ELECTRICAL OSCILLATIONS comprising an arc generator where the electrodes are disposed on a horizontal plane between a vertically disposed magnetic field arranged to produce a field transversely of the arc. The lower magnetic pole is provided with a cup-shaped extension which carries a quantity of mercury for the production of a mercury vapor in the arc chamber.

1,498,898—A. L. Young, filed July 25, 1922, issued June 24, 1924.

SPREADER FOR ELECTRICAL WIRES formed in a cage type of antenna. A circular spreader is provided for spacing the antenna wires of a cage antenna in a cylindrical frame-like body. A detachable binding wire is passed around the circular spreader and is drawn taut to secure the wires with reference to the spreader

1,499,310—J. Bethenod, filed August 2, 1921, issued June 24, 1924.

CALLING SYSTEM FOR RADIO SIGNALING, where a thermoelectric element is connected to the receiving circuit and heated by the incoming signaling energy to generate a direct current for operating a ballistic galvanometer. The galvanometer has con-

siderable inertia and is unresponsive to short impulses. In order to operate the calling device, an extended signal is transmitted over such a period of time that the thermo-electric element becomes sufficiently heated to produce the necessary continuous energy for operating the galvanometer, which in turn closes the local calling circuit.

1,499,331—M. C. Batsel, filed December 11, 1922, issued July 1, 1924. Assigned to Westinghouse Electric and Manufacturing Company of Pittsburgh, Pennsylvania.

RADIO RECEIVING CIRCUIT employing the feed-back principle where the input and output circuits of an electron tube each are provided with variable inductors for providing the feed-back coupling there between. Additional inductance is provided in each of the circuits for establishing such additional feed-back coupling that the degree of regeneration is substantially independent of all adjustments of the inductor in the input circuit of the tube. The circuit arrangement provides a compact radio receiving set.

1,499,403—W. H. Priess, filed December 5, 1924, issued July 1, 1924. Assigned, Wireless Specialty Apparatus Company, Boston, Massachusetts.

ELECTRICAL CONDENSER for high power operation having a plurality of stacks of sheets carried within a metallic casing by which the stacks are connected in series. The opposite terminals for the condenser extend thru the metallic casing and are well insulated therefrom.

1,499,404—William H. Priess, filed December 5, 1921, issued July 1, 1924. Assigned, Wireless Specialty Apparatus Company, Boston, Massachusetts.

ELECTRICAL CONDENSER for use in radio transmission wherein maximum heat radiation from a plurality of stacks may be secured so that the mica dielectric may be operated at a desirable point on its phase angle temperature curve. To allow for rapid heat conduction and radiation the condenser is arranged in a plurality of relatively short stacks and maintained under pressure between heat radiating plates.

1,499,865—H. E. Hallborg, filed September 1, 1923, issued July 1, 1924.

HIGH SPEED SIGNALING SYSTEM, wherein a plurality of sets

of keys are employed with an armature for actuating each of the
keys with negligible electrical time lag. A plurality of opening
coils and a plurality of closing coils are arranged to actuate each
of the armatures. The electrical inertia of the control circuit
which connects the opening and closing coils is extremely small,
permitting high speed operation of the key.

Fig. 1
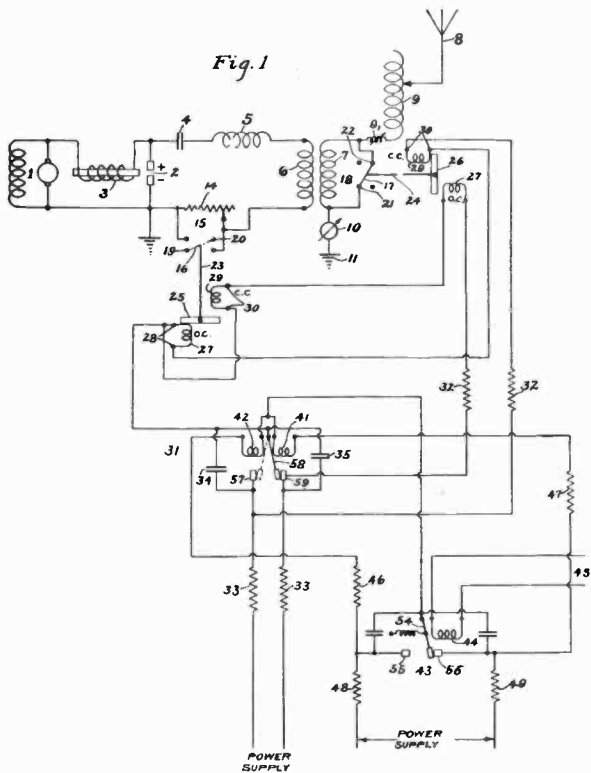
NUMBER 1,499,865—High Speed Signaling System

1,500,476—F. K. Vreeland, filed July 28, 1920, issued July 8,
1924.

INTERFERENCE PREVENTION IN RADIO RECEPTION, having a
pair of energy-collecting systems electrically coupled one with
the other. The system is tuned to the desired signaling frequency
and then another current of interfering frequency whose effect
on the receiving system is opposite to any interfering effect which
may be present when the signal is produced. These opposite
effects are balanced in intensity and phase so that the resultant
effect on the receiver is nil. This balancing of the interfering

signals is accomplished without perceptible reduction in the signal strength of the energy desired to be received.

1,500,528—F. F. Rathbun, filed July 7, 1922, issued July 8, 1924.

VARIABLE PLATE ELECTRIC CONDENSER having cast-in stationary and movable plates for facilitating the protection of the instrument. The stationary plates are supported at three points about the periphery thereof by cast-metal poured into slotted tubular members. The movable plates are similarly supported by a slotted tubular member in which molten metal is poured over the plates.

1,501,104—L. Epenschied, filed September 26, 1919, issued July 15, 1924. Assigned to American Telephone and Telegraph Company, of New York.

MULTIPLEX RADIO SYSTEM, having a plurality of transmitting channels and a corresponding number of receiving channels. Different frequencies are superposed on the different channels, and the effect of these frequencies one upon the other is substantially eliminated by circuits provided in the different channels.

1,501,132—P. B. Murphy, filed May 4, 1921, issued July 15, 1924. Assigned to Western Electric Company, of New York.

SIGNALING SYSTEM employing a carrier wave having an alternate frequency source for modulating the radio frequency wave. According to this invention, at the transmitting station two oscillation generators each designed to supply a current of pre-determined frequency are alternately paralyzed at a lower frequency by a control oscillator, and at the receiving station the signal translating device is supplied with impulses alternating at double the frequency of the control oscillator. The several frequencies are employed at the receiver for controlling the receiving device.

1,501,240—K. Rottgardt, filed August 26, 1921, issued July 15, 1924.

MANUFACTURE AND OPERATION OF VACUUM TUBES, where the electrodes are freed of occluded gases by heating the elements to incandescence and introducing a hydrocarbon vapor into the tube and then evacuating the tube. In this manner the highest possible vacuum is secured and the last remnants of gas are removed from the metal.

Fig 1.
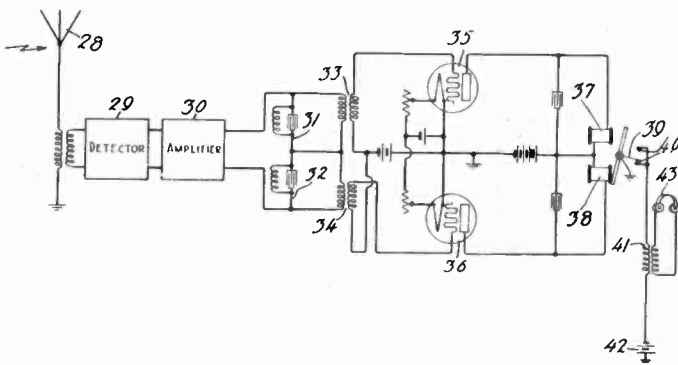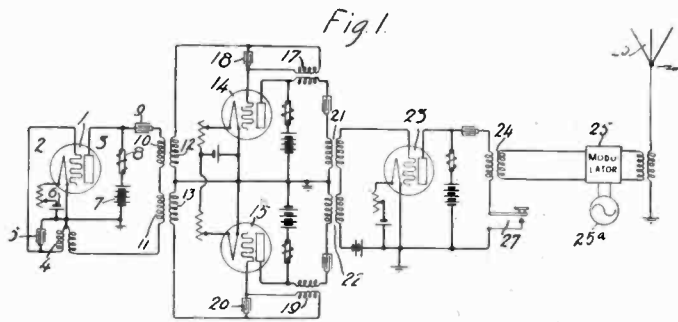
NUMBER 1,501,132—Signaling System

1,501,402—G. Jacobs, filed September 22, 1922, issued July 15, 1924.

VARIABLE CONDENSER, having a broad adjustment member and a fine adjustment member where the rotary plates are set in a tubular insulator having circumferential slots and grooves therein. The stationary plates project into the grooves and the rotary plates may be inter-leaved between the stationary plates.

1,501,543—L. A. Hammarlund, filed November 28, 1922, issued July 15, 1924. Assigned to Hammarlund Manufacturing Company, of New York.

VARIABLE CONDENSER, having broad adjustment and fine adjustment means consisting of a rock shaft on which the movable plates are mounted and a swinging arm operated by a cam for frictionally engaging the rock shaft and moving the same through small angles. The rotary plates are inter-leaved between the stationary plates and the relative position changed both by means of the rock shaft and the secondary swinging arm.

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1,501,664—A. Hund, filed September 18, 1924, issued July 15, 1924. Assigned to General Electric Company, of New York.

RADIO SIGNALING SYSTEM for telephone transmission where modulation is secured by means of an electron tube circuit connected across points in an inductance series connected in the antenna. The electron tube circuit is controlled by a telephone transmitter for varying the amplitude of the transmitted signals.

1,501,683—A. A. Oswald, filed December 6, 1920, issued July 15, 1924. Assigned to Western Electric Company, of New York.

REMOTE CONTROL SYSTEM, having a tramsmitting device arranged to transmit impulses at timed intervals and a receiving device arranged to respond only to impulses received at definite instants and of a duration between two fixed limits. The system is substantially secret whereby a particular control may be locked at a receiver upon the transmission of a definite signal.

1,501,684—A. A. Oswald, filed December 7, 1920, issued July 15, 1924. Assigned to Western Electric Company, of New York.
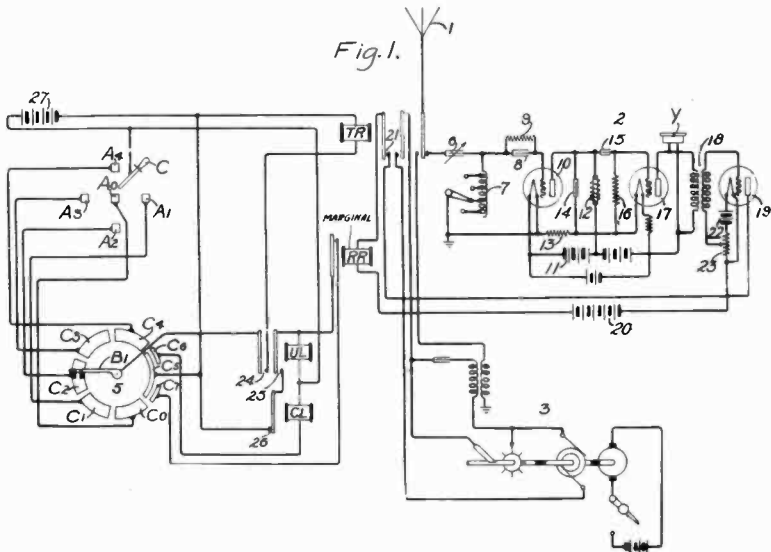
NUMBER 1,501,684—Remote Control System

REMOTE CONTROL SYSTEM, wherein a rotary distributor at both a radio transmitter and receiver are maintained in synchronism by synchronizing impulses sent out from the transmitting station. A circuit arrangement is provided which pre-

vents operation of the controlled devices except when the moving mechanism is in synchronism.

1,501,711—R. A. Heising, filed September 29, 1919, issued July 15, 1924. Assigned to Western Electric Company.
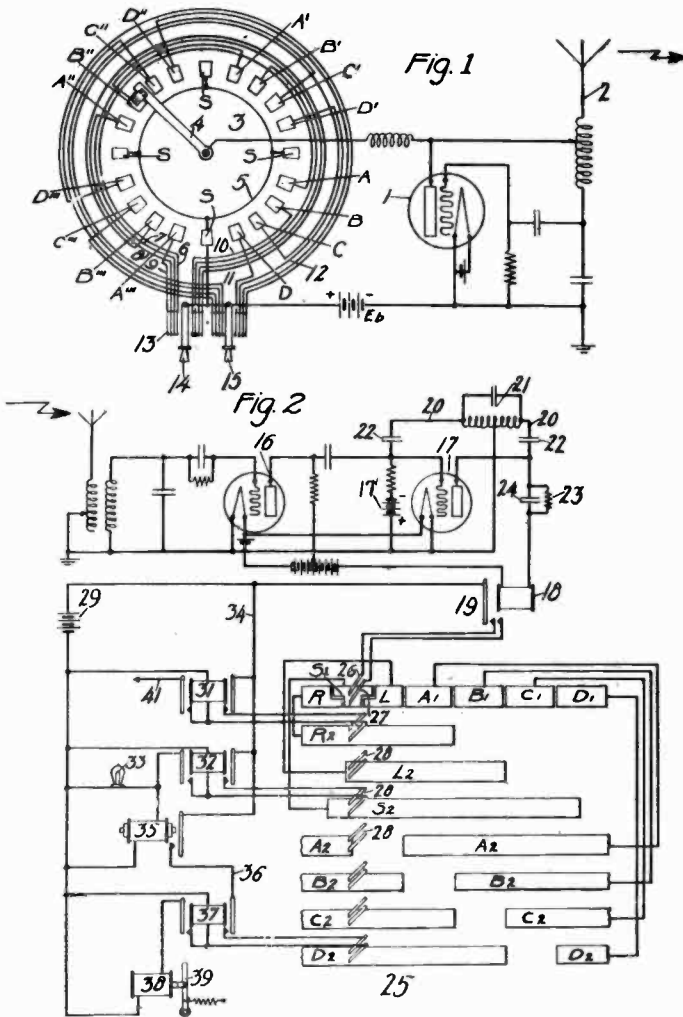
DISTANT CONTROL SYSTEM, in which a group of oscillations is transmitted at a selected period within a timing cycle to determine the particular one of a plurality of distant controls to be

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actuated. At the receiving station the apparatus is maintained in proper phase relation to the transmitter and selected controls actuated.

1,501,729—E. O. Scriven, filed July 14, 1917, issued July 15, 1924. Assigned to Western Electric Company, of New York.

Oscillating Generator having a circuit connected therewith for suppressing the undesired oscillations from the source. A resistance is connected between the source and the filter circuit which suppresses the undesired oscillations to prevent reaction between the filter and the source.

1,501,735—W. C. White, filed June 5, 1915, issued July 15, 1924. Assigned to General Electric Company, of New York.

Radio Signaling System, wherein the transmitter radiates a large amount of energy with a large variation in amplitude of the transmitted energy. An election tube circuit is provided having its cathode and anode connected across a high potential point in the antenna and the ground. Continuous wave energy is supplied to the antenna and the signals are produced by variation in the grid circuit.

1,501,830—E. F. W. Alexanderson, filed April 30, 1914, issued July 15, 1924. Assigned to General Electric Company.

Radio Signaling System, wherein a transmitting system is modulated by means of an electron tube circuit arrangement. The antenna system is continuously supplied with radio frequency energy. An electron tube has its anode and cathode serially arranged with respect to the antenna circuit and its control electrode and cathode included in circuit with a telephone transmitter or other signal controlling device. The amplitude of the radiated energy is thus varied in accordance with the signals.

1,501,831—E. F. W. Alexanderson, filed January 10, 1916, issued July 15, 1924. Assigned to General Electric Company.

Radio Signaling System, wherein modulations are produced at the transmitter by a circuit arrangement in which a variable amount of energy is diverted from the transmitting circuit during the production of signals. A circuit is provided in which inductance is neutralized by capacity through resonance. The effective resistance of the energy absorbing circuit is varied in accordance with signals to cause the radiation of suppression of energy.

1,502,063—W. Schottky, filed November 6, 1920, issued July 22, 1924. Assigned to Siemens and Halske, Aktiengesellschaft, of Siemensstadt, near Berlin, Germany.

ELECTRIC WAVE RECEIVER, wherein a local source is provided for superimposing on the received frequency a local frequency different from that received and arranged to produce a beat frequency current above the limit of audibility. A rectifier is provided which rectifies the beat frequency current. This current is transformed into an alternating current which is then rectified and observed.

1,502,343—B. MacPherson, filed May 25, 1920, issued July 22, 1924. Assigned to Wireless Specialty Apparatus Company, of Massachusetts.

PROCESS AND APPARATUS FOR MAKING ELECTRICAL CONDENSERS of the stack type consisting of alternate layers of mica and foil sheets. A receptacle for the condenser is provided which is substantially higher than the condenser which it is to enclose. The receptacle is filled with melted wax to a point substantially over the top of the condenser. Heat is supplied to the receptacle and to the upper part of the body of wax. The receptacle is then cooled to cause the wax to harden gradually from the bottom upwardly.

1,502,460—G. H. Clark, filed April 26, 1921, issued July 22, 1924. Assigned to Radio Corporation of America, Delaware

RADIO SIGNALING APPARATUS for use with a quenched spark radio transmitter where the quenched spark gap circuit delivers its energy to an intermediate circuit from which it is transferred to the antenna circuit. A mechanical wave changing device is provided for changing the characteristics of each of the circuits simultaneously.

1,502,813—L. Espenschied, filed September 30, 1919, issued July 29, 1924. Assigned to American Telephone and Telegraph Company, of New York.

HIGH FREQUENCY MULTIPLEX SIGNALING SYSTEM, wherein a terminal station is connected by line wire with a distant control station. The antenna circuit is balanced and connected to a conjugate transformer. One of the circuits is tuned to the reception frequency and the other of the circuits to the transmission frequency. A circuit is provided for modulating and demodulating the received and the transmitted currents.
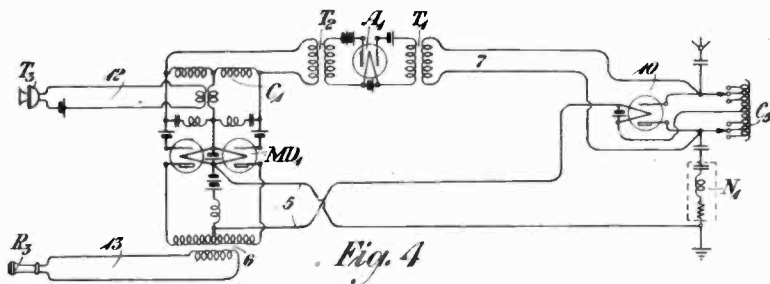
NUMBER 1,502,813—High Frequency Multiplex Signaling System

1,502,831—S. M. Kintner, filed May 11, 1921, issued July 29, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

SYSTEM OF CONTROL for an arc transmitter, wherein a continuously oscillating circuit is provided coupled with an antenna circuit. The signals are produced by reducing or interrupting the currents in the antenna circuit and simultaneously inserting a sufficient impedance in series with the oscillating circuit to render the antenna circuit ineffective.

1,502,848—F. Conrad, filed July 7, 1920, issued July 29, 1924. Assigned to Westinghouse Electric and Manufacturing Company.

TUNING SYSTEM OF ANTENNAS for radio receiving apparatus where the receiving circuit may be broadly tuned to resonance for a given wave length and next sharply tuned to resonance for increasing the intensity of the received signal. The receiving circuit is provided with a variable inductor and a variable condenser mechanically connected to be simultaneously varied so that the algebraic sum of their reactance remains approximately constant.

1,502,860—D. S. McCrum, filed November 24, 1923, issued July 29, 1924.

VARIABLE CONDENSER where the plates are in the form of cylinders arranged to telescope one within the other forming extended cylindrical capacity areas.

1,502,875—M. I. Pupin and E. H. Armstrong, filed February 10, 1916, issued July 29, 1924. Assigned to Westinghouse Electric and Manufacturing Company.
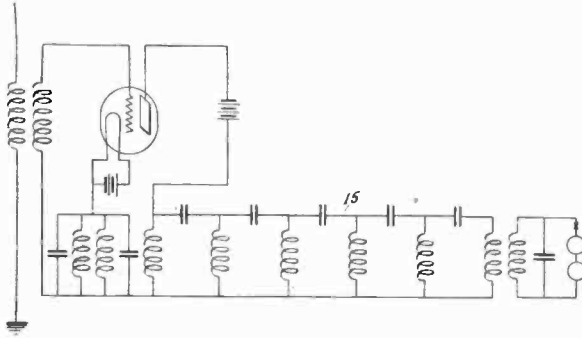
Tone-Producing Radio Receiver, wherein the receiving amplifier is arranged to repeat the incoming waves at an amplitude which varies periodically at an audible frequency so that the resultant electric waves produce a musical note in the receiving system. An electron tube circuit is provided at the receiver with a filter coupling the input and output circuits thereof, the filter including a plurality of like units, each unit comprising two reactances of opposite sign with all the reactances of one sign connected in series and all those of the other sign connected in patallel whereby electrical currents are repeated at a periodically varying amplitude.

1,502,889—H. J. Van Der Bijl, filed January 8, 1918, issued July 29, 1924. Assigned to Western Electric Company.

Method of and System for Radio Signaling, by which a large number of messages may be transmitted simultaneously without the use of a correspondingly large number of radio frequency carrier waves. At a receiving station the radio carrier frequency component of the received waves is first eliminated and each of the modulated auxiliary carrier frequency components is transmitted to a modulator from which the various signaling components may be picked out by suitable band filters.

1,503,250—P. E. Sabine, et al., filed August 7, 1920, issued July 29, 1924. Assigned to Paul E. Sabine.

Apparatus for and Method of Rapid Transmission of Telegraphic Messages by an impulse alphabet, wherein one of two kinds of impulses are transmitted while the other is suppressed. At the receiver the impulse is locally generated and supplied whereby the receiver is operated in a manner as if both kinds of impulses had been transmitted.
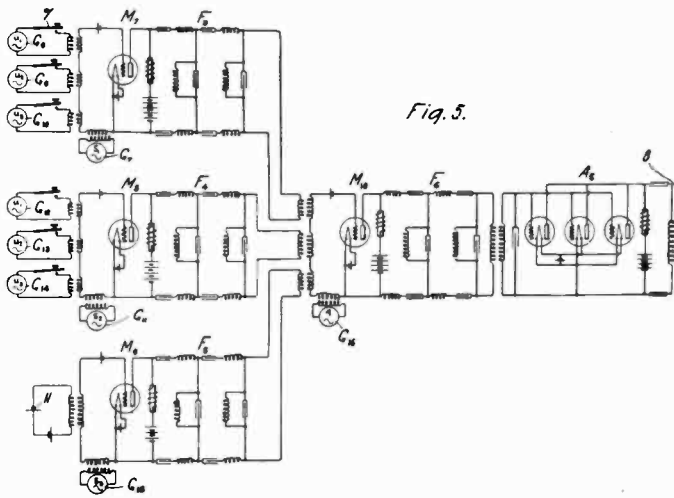
Fig. 5.
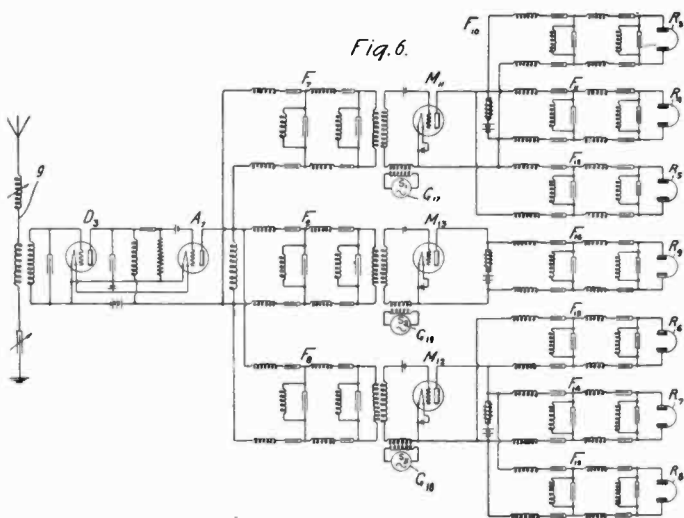
Fig. 6.

Number 1,502,889—Method of and System for Radio Signaling

1,503,308—C. D. Ehret, filed October 22, 1920, issued July 29, 1924.

METHOD OF AND APPARATUS FOR ELECTRICALLY TRANSMITTING INTELLIGENCE in the form of sustained waves without the use of the electrical beat phenomena. The signals are caused to produce a tone frequency and at a point adjacent the production of such tone frequency a magnetically produced sound wave is generated. The sound wave produced by the incoming siganl

combines with the mechanically-produced sound wave to produce a wave beat of audible frequency for observing the incoming signals.
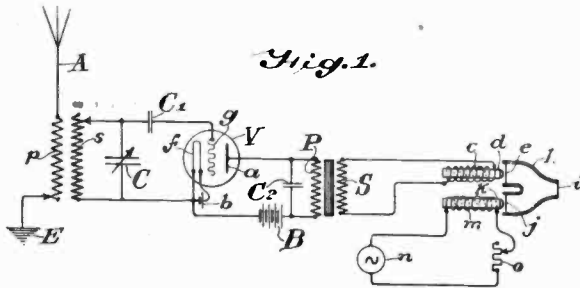
NUMBER 1,503,308—Method of and Apparatus for Electrically Transmitting Intelligence

1,503,324—R. Herzog, filed July 12, 1920, issued July 29, 1924. Assigned to C. Lorenz Aktiengesellschaft, of Lorensweg-Berlin, Tempelhof, Germany.

MEANS FOR PRODUCING ELECTRIC OSCILLATIONS BY AN ELECTRIC ARC, whereby uniform burning of the arc is secured by aid of self induction thru a circuit connected in series with the arc. A capacity sufficiently large as not to affect the tuning is inserted in a circuit including the arc and two self induction coils. This circuit does not oscillate independently and the effect of the self inductance upon the antenna is entirely or nearly neutralized, but the circuit operates to stabilize the burning of the arc.

1,503,709—H. M. Pruden, filed April 3, 1923, issued August 5, 1924. Assigned to Western Electric Company, of New York.

VACUUM TUBE CIRCUITS having automatic means for providing for the continuous flow of heating current from a common source thru a plurality of electron tube cathodes when one or more of the cathodes become broken or otherwise removed from the circuit. The invention relates to a bank of electron tubes where the circuit remains operative even tho one of the tubes may be burned out. A relay is provided which substitutes a resistance for the burned-out filament when such filament becomes open.

1,503,765—G. W. Pickard, filed October 8, 1920, issued August 5, 1924. Assigned to Wireless Specialty Apparatus Company, of Massachusetts.

ELECTRICAL CONDENSER of the stack type where a pair of stacks are arranged transversely of the bottom of the casing with a wedge disposed between the stacks for tightly compressing the stacks against the side walls of the casing.

1,504,002—E. Thomson, filed November 13, 1920, issued August 5, 1924. Assigned to General Electric Company, of New York.

ELECTROSTATIC CONDENSER for high power operation where the condenser is constructed in a stack of thin sheets of alternate conducting material and insulating material. The feature of the invention is the insertion of heat-conducting sheets between the several condenser sections for conveying away heat developed in the condenser.

1,504,303—H. A. Affel, filed June 3, 1921, issued August 12, 1924. Assigned to American Telephone and Telegraph Company, of New York.
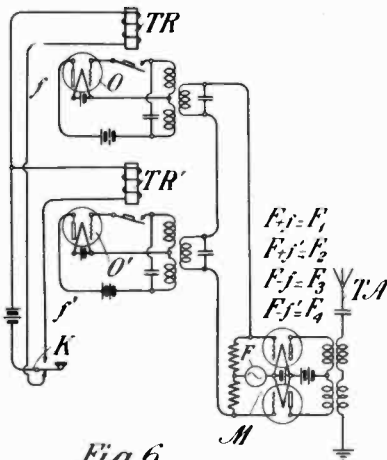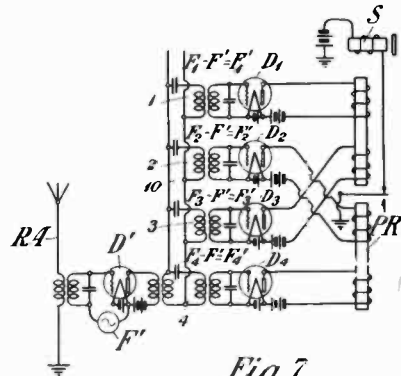
NUMBER 1,504,303—Method of and Means for Reducing Static Disturbances

METHOD OF AND MEANS FOR REDUCING STATIC DISTURBANCES, wherein the effective transmitting energy is increased in strength by transmitting the energy at several different wave lengths simultaneously, altho the total amount of energy radiated at all of the different frequencies need be no greater than that radiated as heretofore at a single frequency. The conjoint action of the several frequencies is employed to operate a receiving relay for observing the transmitted signal.

1,504,426—W. M. Bruce, Jr., filed July 7, 1919, issued August 12, 1924.

RADIO TELEGRAPHY, wherein signals are recorded by a variable ink line on a tape. A relay is connected in the output circuit of an electron tube detector with a movable coil connected with a stylus which is moved upon the receipt of signals from a zero position to a marking position by which an ink line is traced upon a moving tape in accordance with the incoming signals.

1,504,462—R. B. Woolverton, filed July 25, 1921, issued August 12, 1924.

TRANSMISSION OF RADIO SIGNALS EMPLOYING CONTINUOUS WAVES, whereby rapid fluctuation of the signal energy is permitted with the consequent production of sharp rapid signals at receiving stations. An arc generator is shown for producing the oscillatory currents which are delivered to a tuned oscillatory circuit and then impressed upon the antenna circuit. The signals are interrupted by destroying the resonance between the antenna circuit and the oscillatory circuit without interrupting the arc.

1,504,537—H. de F. Arnold, filed September 3, 1915, issued August 12, 1924. Assigned to Western Electric Company, Incorporated, of New York.

POWER-LIMITING AMPLIFYING DEVICE for use in a radio receiving apparatus whereby foreign disturbances and heavy static of large magnitude may be reduced to a value not exceeding that of the signaling energy for enabling the signals to be observed thru such interference. The principle of unilateral conductivity is employed by which to distinguish between the signaling energy and heavy static disturbances. In the preferred form of this device the unilateral conductivity is secured by causing part of the circuit to lie in the path of thermionic currents between hot cathodes and cold anodes, said thermionic currents being oppositely directed with respect to said circuit. These thermionic currents are caused to flow by impressing upon their limiting electrodes, in multiple, an electromotive force operating thru a high impedance, said high impedance performing an important function in connection with the power or current-limiting action of the device.

1,504,570—J. O. Mauborgne, et al., filed July 26, 1922, issued August 12, 1924.
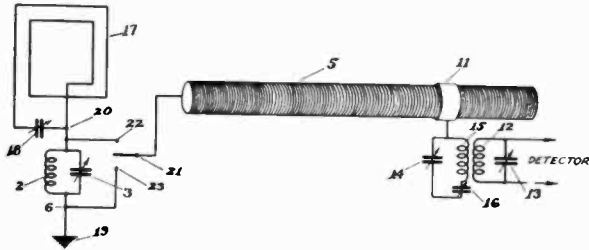
NUMBER 1,504,570—Electrical Signaling

ELECTRICAL SIGNALING, wherein radio signals may be received substantially free of interference by a combination loop antenna and wave coil receiving circuit. The loop antenna is closed thru a variable condenser and connected at one point to ground and to the terminal of a wave coil. The receiving apparatus is coupled by means of a movable ring to the wave coil which is moved along the wave coil to a position for best operation.

1,504,600—Q. A. Brackett, filed January 16, 1919, issued August 12, 1914. Assigned to Westinghouse Electric and Manufacturing Company.

MEANS FOR PROTECTING RADIO OUTFITS FROM STATIC DISTURBANCES, wherein the major portion of the energy of static disturbances is shunted around the receiving apparatus so as to be substantially ineffective in disturbing the receipt of signaling impulses. A pair of rectifying devices are connected in shunt with each other and placed directly across the receiving circuit.

1,504,604—L. W. Chubb, filed May 24, 1921, issued August 12, 1924. Assigned to Westinghouse Electric Company, of Pennsylvania.

SYSTEM OF CONTROL by signaling means of an arc without interrupting the oscillatory condition of the arc. An antenna is employed provided with parallel extended conductors, one set of conductors of which is connected with the arc signaling system, and the other set of conductors to a keying circuit by which the conductors may be included in an absorbing circuit during the production of signals. The sets of conductors in this manner operate as a huge condenser in the absorbing circuit.

1,504,940—G. W. Carpenter, and W. L. Carlson, filed January 29, 1921, issued August 12, 1924.

TELEPHONE RECEIVER CIRCUITS, wherein the telephone head-

set is electrostatically shielded from radio frequency coupling currents which might stray from other parts of a sensitive electron tube amplifier. The shielded cord as claimed in this patent is a feature of the Navy Type Brandes telephone headset. Howling in sensitive multi-stage amplifiers is avoided by using this type of headset.

FIG. 1
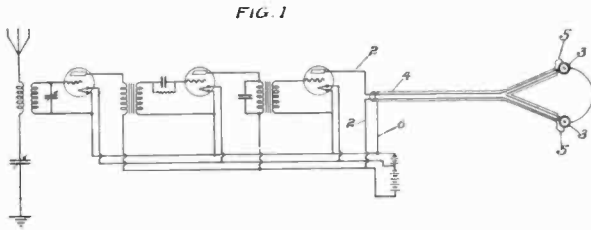
FIG. 2
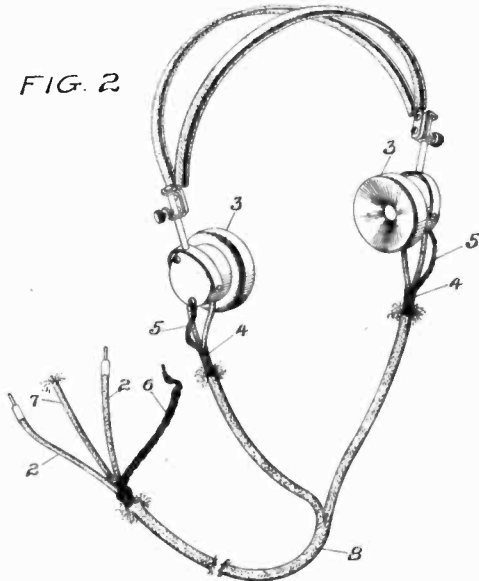
NUMBER 1,504,940—Telephone Receiver Circuits

1,504,974—C. Reno, filed March 1, 1920, issued August 12, 1924.

SYSTEM FOR TRANSMITTING ENERGY WITHOUT WIRES in a confined path in any direction. A spirally-revolving magnetic field is produced in a pair of symmetrically segmated conductors. A parabolic reflecting circuit is arranged for focusing the magnetic field in a desired direction.

1,505,158—De Loss K. Martin, filed November 23, 1921, issued

August 19, 1924. Assigned to American Telephone and Telegraph Company, New York.

FREQUENCY CONTROL SYSTEM of carrier waves from a distant radio station. A controlling frequency is transmitted to a receiving station where it is amplified and caused to operate a synchronous motor. The synchronous motor may be employed to operate any desired form of synchronous device.

1,505,600—G. W. Pickard, filed June 8, 1920, issued August 19, 1924. Assigned to Wireless Specialty Apparatus Company, Boston, Massachusetts.

PROCESS AND APPARATUS FOR THE MANUFACTURE OF ELECTRICAL CONDENSERS, where a condenser of the stack type is submerged in a bath of fluid insulating material in a casing and then, while in the bath, a permanent pressure is applied to the stack to drive out the insulating material from between the sheets. The sheets are thus brought into intimate contact with a permanent filler of insulating material around the stack without the stack being exposed to atmosphere after its submerging.

LIST OF RADIO TRADE MARKS PUBLISHED BY PATENT OFFICE PRIOR TO REGISTRATION

(The numbers given are serial numbers of pending applications)

191,177—"VOISOMETER" for radio receiving apparatus. Voisometer Mfg. Co., Mirkwood, Missouri. Claims use since January 23, 1924. Published July 8, 1924.

183,996—"HETEROFLEX" for radio apparatus. Electrical Research Laboratories, Chicago, Illinois. Claims use since November 20, 1923. Published July 8, 1924.

194,212—"POLYTRON" for radio apparatus. Polydyne Corporation, New York City, New York. Claims use since about March 8, 1914. Published July 8, 1924.

194,214—"POLYCOIL" for radio apparatus. Polydyne Corporation, New York City, N. Y. Claims use since about March 8, 1924. Published July 8, 1924.

194,215—"POLYFONE" for radio apparatus. Polydyne Corporation, New York City, N. Y. Claims use since about March 8, 1924. Published July 8, 1924.

194,440—"DYNOFORMER" for radio transformers and coupling

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coils. Electric Service Engineering Corporation, New York, N. Y. Claims use since February 21, 1924. Published July 8, 1924.

194,441—"Dynodon" for radio condensers. Electric Service Engineering Corporation, New York, N. Y. Claims use since February 21, 1924. Published July 8, 1924.

194,464—"Micro-Helix" for variable condensers. Packwood Brothers, Lincoln, Nebraska. Claims use since February 15, 1924. Published July 8, 1924.

194,631—"Trade mark consists of the picture of Lafayette, a deceased celebrity," for radio receiving apparatus. Thompson Mfg. Company, Jersey City, New Jersey. Claims use since February 7, 1924. Published July 8, 1924.

166,123—"Vocalion" for radio receiving and transmitting sets The Aeolian Co., New York, N. Y. Claims use since about June 14, 1922. Published July 22, 1924.

170,673—"Resist-O-Meter" for ohmic resistance for radio receiving sets and the like. Scholes Radio and Mfg. Co., New York, N. Y. Claims use since August 14, 1922. Published July 22, 1924.

194,211—"Poly" for radio apparatus and receiving sets. Polydyne Corporation, New York, N. Y. Claims use since about March 8, 1924. Published July 22, 1924.

194,213—"Polystat" for radio apparatus and receiving sets. Polydyne Corporation, New York, N. Y. Claims use since March 8, 1924.

194,785—"Microfarad Junior" for condensers. Chelten Electric Co., Philadelphia, Pennsylvania. Claims use since on or about November 13, 1923. Published July 22, 1924.

194,786—"Midget Vernier" for condensers. Chelten Co., Philadelphia, Pennsylvania. Claims use since on or about January 31, 1923. Published July 22, 1924.

194,794—"GR" in ornamental design for radio apparatus and receiving sets. General Radio Co., Cambridge, Massachusetts. Claims use since October 1, 1923. Published July 22, 1924.

194,830—"Radiolamp" for loud speakers. Robert E. Wheelan, New York, N. Y. Claims use since March 15, 1924. Published July 22, 1924.

193,741—"Ware" for radio sets and parts thereof. Wave Radio Corporation, New York, N. Y. Claims use since May 4, 1922. Published July 22, 1924.

193,837—"The Little Kelicoil" for tuning inductances for radio circuits. Patrick J. Kelly, Philadelphia, Pennsylvania. Claims use since November 11, 1923. Published July 22, 1924.

190,890—"The Concert Modulator" for loud speakers. Manhattan Electrical Supply Co., Inc., New York, N. Y. Claims use since August 28, 1923. Published July 22, 1924.

195,331—"Vario-Farad" for variable condensers. Thomas E. Binder doing business as Elec-Rad Co., Trenton, New Jersey. Claims use since September, 1923. Published July 22, 1924.

195,718—"Porat-Radio" for radio apparatus. Service Radio Corporation, New York, N. Y. Claims use since about April 1, 1924. Published July 22, 1924.

195,755—"Molodyne" for radio receiving sets and parts thereof. The Melodyne Co., New York, N. Y. Claims use since April 10, 1924. Published July 22, 1924.

196,258—"Grand Opera Volumeter" for radio headsets and loud speakers. Radio Industries Corporation, New York, N. Y. Claims use since about December 28, 1923. Published July 22, 1924.

196,466—"Heliotron Tube" for electron tubes. F. S. McCullough Laboratories, Wilkinsburg, Pennsylvania. Claims use since April 26, 1924. Published July 22, 1924.

196,555—"Audiotran" for radio receiving sets and parts thereof. Scientific Equipment Co., Ridgewood, New Jersey. Claims use since April 17, 1924. Published July 29, 1924.

196,814—"Rit-One" for radio apparatus, in ornamental design. Radio Sales Co., New York, N. Y. Claims use since January 1, 1924. Published July 29, 1924.

196,826—"Magna-Lion" for crystal detectors. Harry B. Schiller, Philadelphia, Pennsylvania. Claims use since May 6. 1924. Published July 29, 1924.

186,703—"The Bull Dog Grip" for plugs for radio telephones, in ornamental design. John Firth, New York, N. Y. Claims use since 1921. Published August 5, 1924.

191,580—"THE ATLANTIC & PACIFIC RADIO Co." in ornamental design, for radio apparatus. Atlantic & Pacific Radio Co. Claims use since June, 1921. Published August 5, 1924.

193,830—"AMPLIOLA" for loud speakers. Alfred Graham & Co., of London, England. Claims use since April 1, 1923. Published August 5, 1924.

193,842—"MANHATTAN" for radio apparatus. Manhattan Electrical Supply Co., New York, N. Y. Claims use since 1889. Published August 5, 1924.

194,532—"SINGLE-TROL" for radio apparatus. G. H. Fischer & Co., Glendale, New York, N. Y. Claims use since March 3, 1924. Published August 5, 1924.

194,891—"UNIDYNE" for radio receiving sets. United Mfg. & Distributing Co., of Chicago, Illinois. Claims use since January 1, 1924. Published August 5, 1924.

195,583—"LINRAD" for radio receiving sets and parts thereof. Lininger Radio Co., Cleveland, Ohio. Claims use since February 11, 1924. Published August 5, 1924.

195,685—"HARCO" for radio receiving sets and parts thereof. Harris & Birdseye, Inc., New York, N. Y. Claims use since October, 1923. Published August 5, 1924.

166,125—"AEOLIAN" for radio transmitting and receiving sets. The Aeolian Co., New York, N. Y. Claims use since June 14, 1922. Published August 12, 1924.

193,527—"LYTTON" for radio set and parts thereof. Walter Lytton, Inc., Chicago, Illinois. Claims use since January 1, 1924. Published August 12, 1924.

193,556—"TRESCO" for radio set and parts thereof. Tri-City Radio Electric Supply Co. Claims use since 1914. Published August 12, 1924.

193,557—"TRESCO" in ornamental design, for radio sets and parts therefore. Tri-City Radio Electric Supply Co. Claims use since 1914. Published August 12, 1924.

194,592—"D" for radio sets and supplies. De Forest Radio Telephone & Telegraph Co. Claims use since 1919. Published August 12, 1924.

194,730—"TRAYDIO" for radio apparatus. Musical Products Distributing Co., Inc. Claims use since March 17, 1924. Published August 12, 1924.

194,982—"Teledyne" for radio receiving sets and parts thereof. Cutting & Washington Radio Corporation. Claims use since February 11, 1924. Published August 12, 1924.

195,118—"Accuratune" for radio apparatus. The Mydar Radio Co., Newark, New Jersey. Claims use since June 15, 1922. Published August 12, 1924.

196,330—"Crystalstat" for crystal detectors and rectifiers. The Breckenridge Electric Co., Detroit, Michigan. Claims use since March 1, 1924. Published August 12, 1924.

195,981—"Minute Man" for radio sets and parts therefor. Pathe Phonograph & Radio Corp., Brooklyn, New York. Claims use since February 6, 1924. Published August 12, 1924.

195,982—"Deputy" for radio set and parts therefor. Pathe Phonograph & Radio Corp., Brooklyn, New York. Claims use since January 30, 1914. Published August 12, 1924.

196,420—"Neutradial" for radio sets and parts therefor. Broadcast Manufacturers, Inc. Claims use since March 31, 1924. Published August 12, 1924.

196,441—"Telomonic" for radio set and parts therefor. Danzer-Jones, Inc., New York, N. Y. Claims use since April 2, 1924. Published August 12, 1924.

196,524—"K-E" for radio apparatus. Kirkman Engineering Corporation. Claims use since February 19, 1924. Published August 12, 1924.

187,998—"Woodehorn" for radio sets and parts therefor. Inter-Ocean Radio Corporation. Claims use since September 1, 1922. Published August 12, 1924.

XI

ALPHABETICAL INDEX TO ADVERTISEMENTS

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