VOLUME 23 SEPTEMBER, 1935 NUMBER 9

PROCEEDINGS
of
The Institute of Radio Engineers

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PROCEEDINGS OF

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GENERAL INFORMATION

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AIMS AND OBJECTS. The Institute functions solely to advance the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Among the methods of accomplishing this is the publication of papers, discussions, and communications of interest to the membership.

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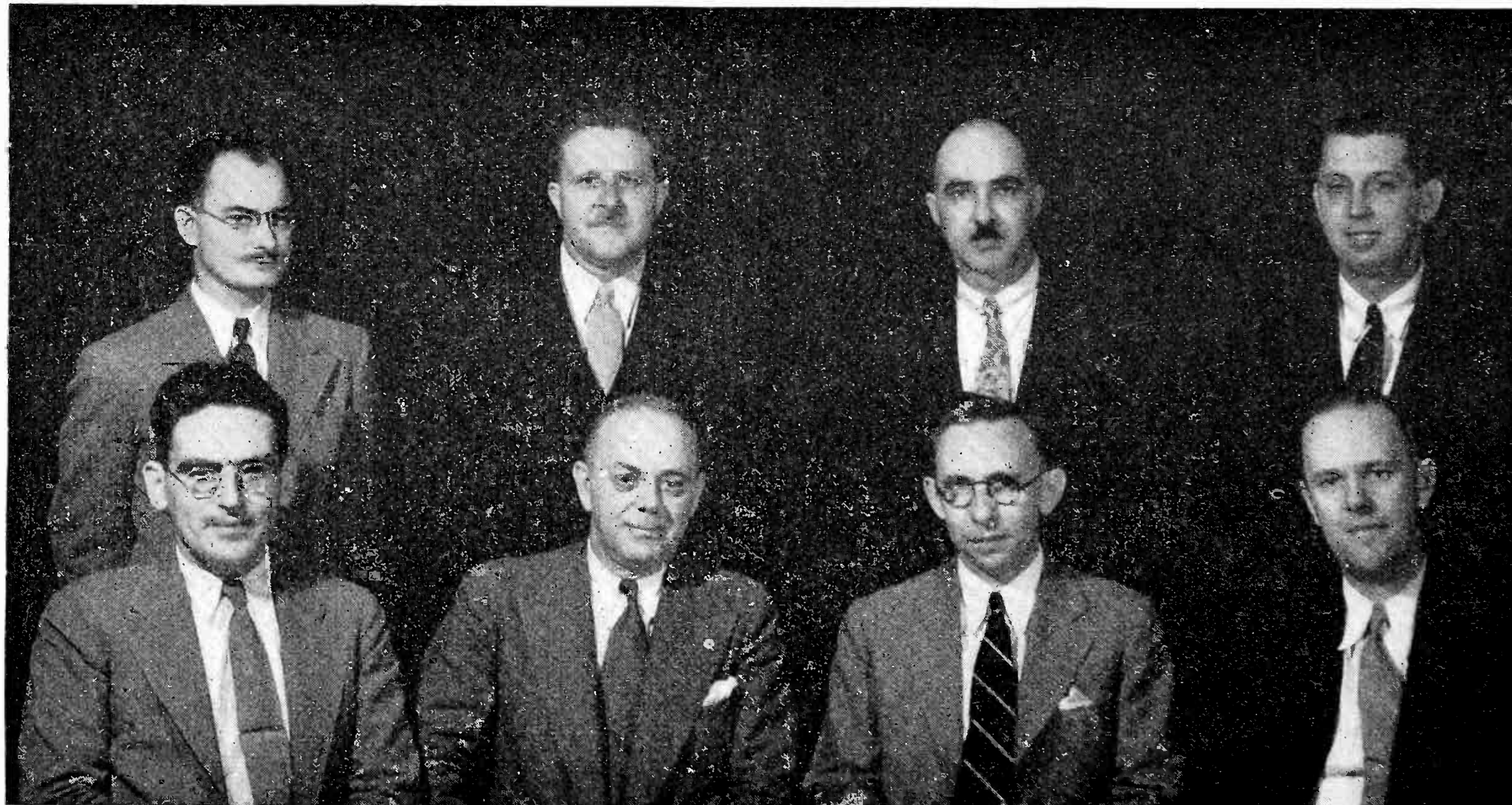

TENTH ANNUAL CONVENTION COMMITTEE

Top row, left to right—F. H. Roush, S. D. Le Barre, E. H. I. Lee, and E. C. Denstaedt.
Bottom row, left to right—Samuel Firestone, G. W. Carter, H. L. Byerlay, and E. W. Sears.

INSTITUTE NEWS AND RADIO NOTES

Tenth Annual Convention

The Tenth Annual Convention of the Institute which was held in Detroit, Michigan, was attended by four hundred and ninety men and ninety-six women and was held in the Hotel Statler during the first three days in July. The banquet, on the evening of July 2, was attended by four hundred and twelve. There were thirty-four exhibitors who filled all of the room available for that purpose with samples of their products and demonstrations of equipments.

Twenty-one papers were presented at the seven technical sessions which were held. Visits were made to General Motors Research Laboratory, Greenfield Village, and the Ford Motor Plant.

The Institute awards were presented to the recipients during the banquet by President Ballantine. W. G. Bryant, the Netherlands consul in Detroit accepted the Medal of Honor in behalf of Dr. van der Pol and the Morris Liebmann Memorial Prize was handed to Dr. Llewellyn.

Those who were present appreciated the effective arrangements made by H. L. Byerlay and his convention committee who deserve the sincere thanks of all who participated in the meeting.

Personal Mention

C. B. Aiken formerly with Bell Telephone Laboratories is now Associate Professor of Electrical Engineering in charge of communication at Purdue University.

R. C. Ballard previously with General Household Utilities Company has become a television engineer for RCA Victor in Camden, N.J.

L. M. Clement has been named vice president in charge of engineering of the RCA Manufacturing Company at Camden having previously been with the International Standard Electric Company at London, England.

F. R. Dodge, Lieutenant Commander, U.S.N., has been transferred from the U.S.S. *Detroit* to the U.S.S. *Raleigh* basing at San Diego, Calif.

J. B. Hawkins has left Emerson Television and Radio Corporation to become general superintendent of the radio receiving set division of RCA Manufacturing Company at Camden.

B. L. Hubbard is now an industrial engineer for Philco Radio and Television Company at Philadelphia having previously been with the RCA Victor Company.

I. H. Loucks in the Federal Communications Commission inspection service has been transferred from Grand Island, Neb., to Philadelphia, Pa.

M. C. Partello, Lieutenant Commander, U.S.N., has been transferred from the U.S.S. *Biddle* to the Brooklyn Navy Yard.

V. S. Roddy has left Wired Radio, Inc., to join the engineering staff of the United States Signal Corps at Wright Field, Dayton, Ohio.

E. W. Sanders formerly with the Juniata Radio Laboratory is now with the RCA Manufacturing Company at Camden.

R. A. Swan, Jr. of the Hygrade Sylvania Corporation has been transferred from Salem, Mass., to Emporium, Pa.

H. C. Tittle previously with Stewart-Warner Corporation has become chief engineer of the Radio Division of General Household Utilities Company in Chicago.

Proceedings of the Institute of Radio Engineers
Volume 23, Number 9 *September, 1935*

TECHNICAL PAPERS

AIRCRAFT RADIO EQUIPMENT FOR USE ON EUROPEAN AIR LINES*

By

A. D. Hodgson

(The Plessey Company, Ltd., Ilford, Essex, England)

Summary—*The apparatus to be described has been developed to provide a transmitting and receiving equipment for small and medium commercial and private aircraft which will be extremely simple to operate and as foolproof as possible.*

Many features are introduced which, although they have been in use for broadcast and other commercial radio apparatus, have not been applied to aircraft radio equipments.

For example, an electrical remote tuning system is used for the receiver, and automatic volume control on both telephony and telegraphy is provided. The equipment provides a telephonic range of 180–200 miles, and a telegraphic range of 350–400 miles, and weighs only sixty pounds complete.

FOR communication on the European air routes frequencies between 340–350 kilocycles are in general reserved for telephony, while telegraphic working is confined to the band of 325–336 kilocycles. The majority of large aircraft on regular international routes employ an operator and carry out all their communication by telegraphy, while the smaller aircraft on shorter routes and machines used for private charter employ telephony and the equipment is operated by the pilot, as the extra weight and accommodation for an operator cannot be allowed.

In some cases telephony and telegraphy are carried out on the same wavelength, for example, at Le Bourget, where the quantity of telephony communication is relatively small, both telegraphy and telephony are carried out on 333 kilocycles. In addition to this, on other routes direction finding facilities are only available on 325 kilocycles. The aircraft equipment must therefore be capable of working on 350, 333, and 325 kilocycles, and furthermore a quick change from one to the other without any following adjustments to obtain accurate setting is necessary. Again, where aircraft are operating over a long sea crossing the frequency of 500 kilocycles is also necessary for emergency work. The equipment must also be capable of transmitting continuous waves, interrupted continuous waves, or telephony on any

* Decimal classification: R520. Original manuscript received by the Institute, September 10, 1934.

of these four wavelengths, and the type of transmission must be capable of instantaneous change.[1]

To fix the operating frequencies exactly the transmitter employs a master oscillator in which a fixed subdivided condenser with semi-variable trimming condenser is employed, separate sections being switched in for each wavelength. The master oscillator drives two amplifier valves in parallel, and to avoid the possibility of overloading the amplifier valves when the aerial circuit is untuned, the anodes of the amplifiers are coupled to the aerial circuit via a high impedance transformer.

This output circuit arrangement has the added advantage that the operating frequency of the transmitter may be changed while it is in operation, further if the aerial is lost or if the transmitter is switched on before the aerial is run out, no damage can occur to the amplifier valves. The transmitter is therefore extremely foolproof, and no amount of misuse of the controls will cause any damage to the valves or circuit.

Modulation for telephony is performed by the valve grid leak method, and interrupted-continuous-wave telegraphy is obtained from a high note buzzer having a secondary winding connected to the microphone transformer primary. Keying for telegraphy is performed in the grid circuit of the amplifiers, a heavy negative bias being applied during the spacing period.

The receiver is a superheterodyne having a detector-oscillator of the pentagrid type with two coupled preselector circuits, one intermediate-frequency amplifier working at 117 kilocycles, a double diode triode, and a double diode pentode. Delayed automatic volume control is provided by the double diodes, and for continuous-wave reception the triode section of the double diode triode is switched on to an oscillating circuit whose frequency and amplitude are adjusted to operate in conjunction with the automatic volume control circuits and to provide a clear beat note free of key clicks from ten microvolts total input to two millivolts.

On telephony the triode section of the double diode triode forms an additional low-frequency stage and drives the double diode pentode for maximum output. With this arrangement a maximum of 0.6 watt can be obtained across the telephones for a total signal input of thirty-five microvolts.

The oscillator-detector portion of the circuit is built into a separate unit, and the output from this frequency converter is passed via a line

[1] *Note:* Since this article was written additional operating frequencies have been called for, and now the transmitter must operate on 500 kilocycles, 365, 345, 333, 326 at least.

to the main receiver unit which contains the intermediate amplifier and the remainder of the circuits. By this means the frequency converter which contains the tuning circuits can be placed near the pilot of the aircraft and thus enable direct tuning of the receiver to be obtained without the use of mechanical remote controls with the inevitable trouble of backlash.

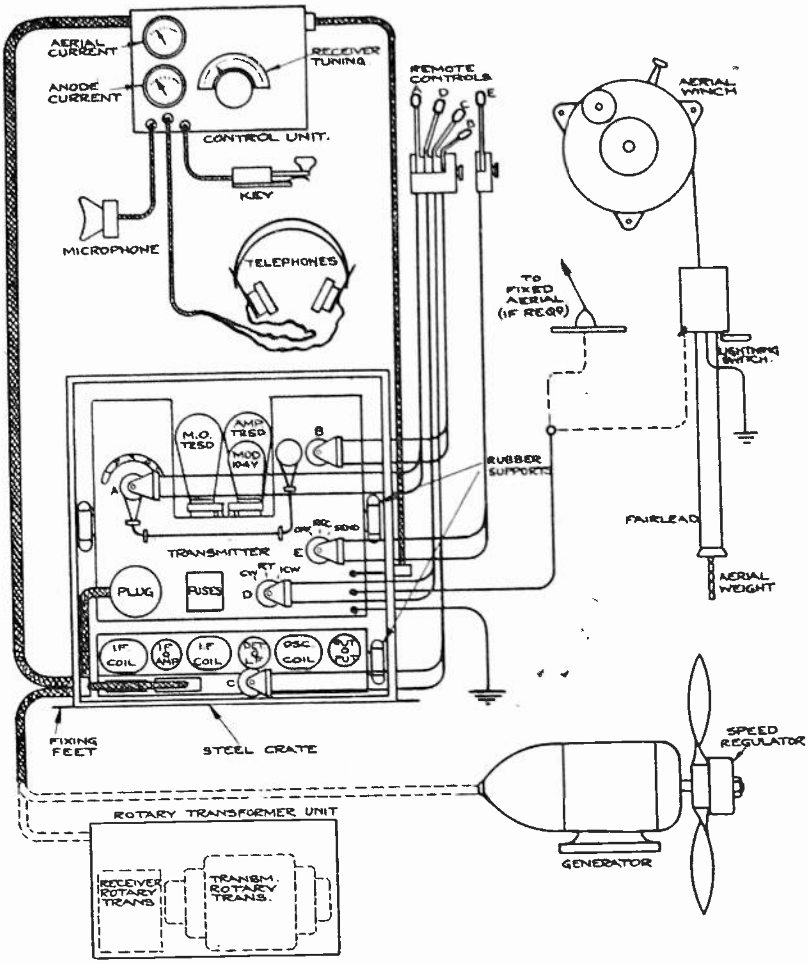

Fig. 1—Schematic layout of type AC 44 equipment.

Owing to the automatic volume control circuit no reaction control is necessary, and the receiver therefore has only one accurately calibrated control and is extremely simple to use.

Incorporated in the frequency converter unit are two meters reading the transmitter aerial current and the total direct-current feed to the equipment; these meters are remotely operated. Plugs for microphone, telephones, and key are also provided. This unit thus forms in effect the main control unit, and is mounted near the pilot together with the remote controls for the transmitter switches, while the remainder of the equipment may be placed in any convenient part of the

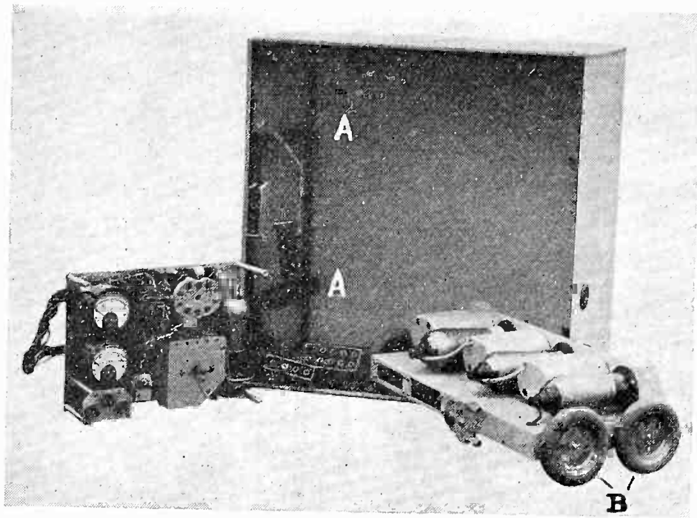

Fig. 2

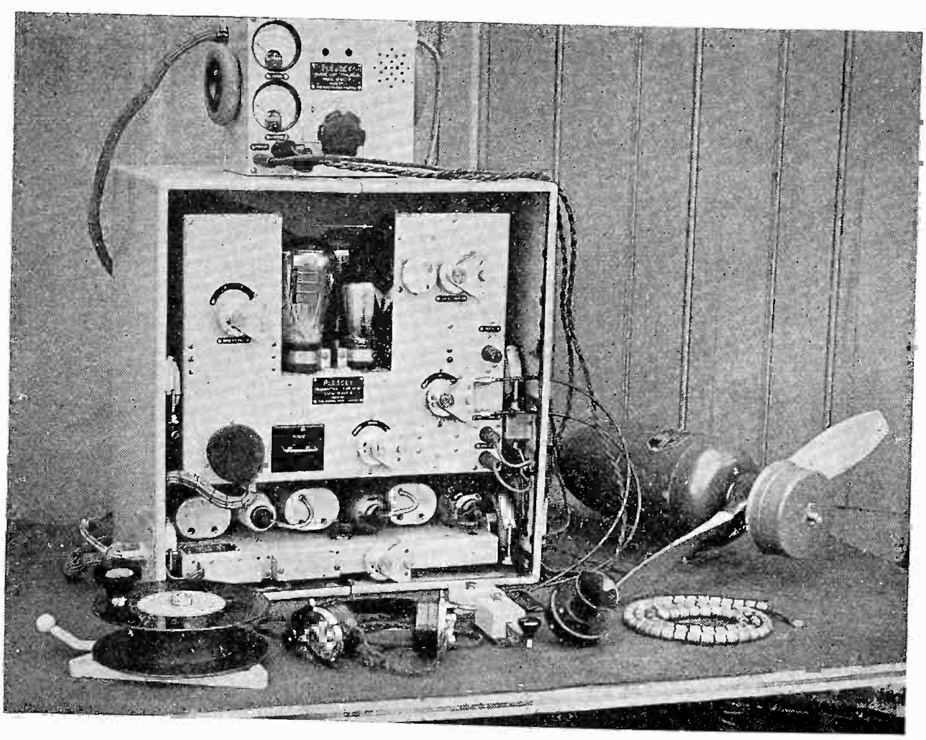

Fig. 3

machine. Fig. 1 shows the schematic layout of the complete installation.

The transmitter and receiver are both built on an anodized and enameled aluminium framework. On two sides of these frameworks steel disks carrying sponge rubber rings are fixed. These rings engage in supports in a tubular steel crate; the transmitter and receiver therefore sit one above the other on these rubber rings in the crate which is provided with light removable aluminium sides for protection. The crate may therefore be bolted direct into the aircraft without any further rubber supports. In Fig. 2 the crate, receiver, and control unit with cover removed are shown. The supports for the rubber rings are shown at A, while the steel disks and rubber rings are marked B. The two meters and microphone, telephone, and key plugs are plainly visible on the control unit, also the frequency changing valve.

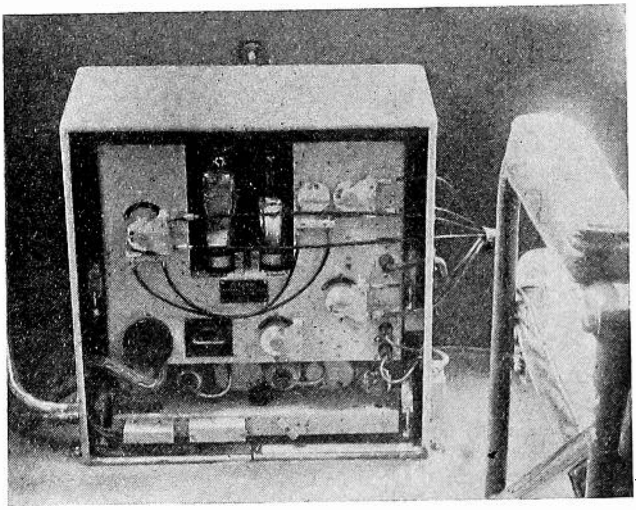

Fig. 4

The power supply for the equipment may be obtained from a wind-driven generator or rotary transformer running from the main lighting battery. Power at nine volts low tension and 600 volts high tension is required, and on telephony the transmitter has an output power of eighteen watts with seventy per cent modulation; on telegraphy thirty-five watts output is obtained. With this power from medium size eight- to ten-seater machines a range of 180–200 miles telephony and 350–400 telegraphy is obtainable between the aircraft and the normal airdrome stations.

The entire equipment consists of a transmitter and receiver in crate, control unit, generator with windmill and speed regulating mechanism, aerial winch with two hundred feet of wire and two pounds flexible weight, aerial fair-lead, remote controls for transmitter switches, telephones, microphone, and key.

The total weight of these items is sixty pounds. The equipment therefore detracts but little from the pay load of the aircraft, and yet is solidly built, reliable, and simple to operate.

Fig. 3 shows the entire equipment with the transmitter and receiver mounted in the crate. In Fig. 4 the equipment is seen mounted behind the pilot's cockpit in a seven-seater aircraft.

ACKNOWLEDGMENT

In conclusion I wish to thank the Plessey Company of Ilford, England, for permission to publish these details.

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PARASITES AND INSTABILITY IN RADIO TRANSMITTERS*

By
G. W. Fyler
(General Electric Company, Schenectady, N. Y.)

Summary—*Several types of parasites or spurious oscillations and other forms of radio transmitter instability are described. Methods of locating and eliminating parasitic circuits are discussed.*

A PARASITE in radio work is any spurious oscillation taking place in a vacuum tube circuit other than the normal oscillation for which the circuit is designed.

Parasites occur as any normal oscillation when the necessary conditions for oscillation exist. In a great many instances troubles which are attributable to other causes are actually due to parasites. They may cause additional carriers and side bands to be radiated, voltage flashover, instability, loss of efficiency, and short life or failure of vacuum tubes and other circuit elements. Parasites or spurious oscillations cannot be fully anticipated in building radio transmitters of a new type. If the best results are desired it is usually necessary to eliminate any existing parasites by special tests after a transmitter has been constructed. In many cases the determination of the parasitic circuit may require considerable study and the use of cut-and-try methods. Detuning and damping of the offending circuit to stop the oscillation is then quite simple. It is the purpose of this paper to describe methods of determining and eliminating common parasites and a few other forms of transmitter instability.

Ten or fifteen years ago most vacuum tube transmitters consisted of an oscillator tube feeding the antenna directly, without much attention being paid to frequency stability or efficiency. If the oscillator did not function properly it was common practice to reconstruct the outfit with one of the other oscillator circuits. Actually, the trouble may have been due to parasitic oscillations which were not fully recognized at the time.

We still have parasites in radio transmitters and with high power tubes and complex circuits the trouble is increased.

During the installation of the WLW 500-kilowatt transmitter using twenty 100-kilowatt tubes, it was necessary to eliminate approximately a dozen parasitic oscillations before stable operation and high efficiency

* Decimal classification: R133×R355. Original manuscript received by the Institute, March 25, 1935. Presented before Washington Section, U.R.S.I., February 11, 1935.

were obtained. This was not indicative of poor design; it was the condition to be expected when first testing a very high power transmitter. Oscillations were found all the way from an audio frequency which could be heard up to an ultra-high frequency of 200 megacycles, or one and one-half meters wavelength. The latter oscillation was very unusual considering the large coils and condensers used in this transmitter. It is desirable to consider the way in which it was found and studied, in order to illustrate the peculiar conditions which may exist with parasites.

Another parasite at a frequency of twenty megacycles was being studied in one of the three large amplifiers with a wavemeter tuned to the frequency of oscillation. The plate voltage of 12,000 volts was removed to make a circuit change and the wavemeter current reading did not drop to zero. The meter reading dropped to zero only when the bias was removed from the amplifier. Also the oscillation started up again when the bias was applied. During these tests the plate ammeters read zero current. It was noticed that the wavemeter current reading could not be changed appreciably by tuning the wavemeter, nor could any resonant frequency be found throughout the available wavemeter range of twelve to 20,000 meters. However, the greatest meter reading was obtained with the smallest wavemeter coils and with the tuning condenser at its minimum position. Therefore, midget condensers and coils were connected with a 100-milliampere radio-frequency meter in an improvised wavemeter to find the approximate frequency of oscillation. Finally, with a two-plate midget condenser connected directly across the terminals of the meter with leads about one inch long a resonant point was found and a full scale reading of the meter obtained with the improvised wavemeter located near one of the large amplifier tubes. The oscillation wavelength was found to be about one and one-half meters by measurements with Lecher wires. Since positive bias was being used to obtain this oscillation it was apparent that the oscillation was of the Barkhausen type. This was verified by noting the frequency change with variations of the grid potential. Knowing the bias voltage and the diameter of the tube anode a good check on the observed frequency was obtained.

This parasite is probably present in most class B and C radio-frequency amplifiers and oscillators during negative peaks of the radio-frequency plate voltage. Apparently the oscillation is not detrimental because of its limited conditions of oscillation and inherent poor efficiency. No high voltages appear to result from the oscillation even in 100-kilowatt tubes.

Probably the most detrimental parasites are those which cause flashovers or which materially lower the amplifier efficiency. If flashovers occur at the seal of a costly tube or if the efficiency is only half the expected value, the parasite is most undesirable.

The tubes and associated circuits in a transmitter may have damped and undamped parasites depending upon the circuit losses, the feed-back coupling, the grid and plate potentials, and the reactance or tuning of the parasitic circuits. The damped oscillations or "trigger" parasites as they are sometimes called occur as the result of a shock or transient with peak modulation, keying transients, or flash arcs in vacuum tubes. It is well known that oscillations start more easily with high plate voltage and with zero, or even positive, grid-bias potentials, due to increased mutual conductance. Hence parasites may oscillate only during part of a modulation cycle. However, it is advisable and usually possible to obtain steady-state or undamped oscillations of all parasites in order to study the essential circuit conditions. The possibility of an oscillation existing may be predicted mathematically or by previous experience under certain conditions, but with existing parasites it is usually a question of studying the circuit when oscillation exists and changing or damping the circuit so as to eliminate the parasite.

There are certain parasites which stop when the grids are saturated as in a class C telegraph or plate modulated stage, but some of the parasitic oscillations may still exist even when the grids are well saturated. This is particularly true of ultra-high-frequency parasites where the parasitic frequency is much higher than the normal amplifier frequency and the parasite may oscillate for many cycles during each radio cycle. It has also been found that low-frequency parasites may be present during part of a modulation cycle in a class B radio-frequency amplifier as described later.

A capacity must have connections made to it and at a very high frequency the capacity circuit may actually be inductive with negligible series resistance. Also a coil has distributed capacity and at certain frequencies the inductance coil may be, in effect, a capacity with negligible series resistance. Hence, one should visualize a capacity being an inductance, and an inductance being a capacity at certain frequencies. Also it is necessary to consider the interelectrode capacities in vacuum tubes as tuning and coupling elements with high-frequency parasites. Shunt tuned parasitic tank circuits are more common than series tuned circuits, but abnormally high radio-frequency voltages indicate the presence of series resonant circuits and these are frequently very difficult to locate.

For determining whether parasites may occur in a given amplifier, the normal radio-frequency grid excitation or oscillation should be removed without upsetting the normal circuit conditions. Reduced plate voltage is advisable for the first test and the bias should be reduced to a low value without causing the plate dissipation rating to be exceeded except possibly for momentary tests. It is desirable to try positive grid bias with low plate voltage for dynatronic grid parasites. Most oscillations have been found to start best in certain parts of the tube characteristic and, therefore, the filament, grid, and plate voltages should be varied over wide limits, including the emission limiting region which seems especially important.

On large tubes with tungsten filaments the filament cannot be damaged by running at low voltage, and it is easy to control the amplitude of a parasitic oscillation within reasonable limits by lowering the filament voltage. The parasitic frequency and circuit can then be determined by a wavemeter coupled to the circuit and by finding the high voltage points. When these facts have been determined, it is generally possible to deduce the type of oscillation and the circuit. When one parasite is eliminated it is quite possible that a different type of parasite may start. Although vacuum tubes have been found to oscillate simultaneously on more than one frequency, it is quite possible that one oscillation may prevent one or more other oscillations.

A small neon lamp is commonly used to indicate whether an oscillation exists and for finding the high voltage points in the circuit. The lamp may be fastened on the end of an insulated rod and, for best results, the glass part of the neon lamp should be brought close to the various conductors, coils, and capacitors where parasitic voltage may exist. The plate and grid connections have the highest parasitic potentials in most cases. For studying a low-frequency parasite it is sometimes desirable to add a piece of wire to a terminal of the lamp to increase the voltage sensitivity of the lamp.

If it is not possible to obtain steady-state oscillations, the various transmitter circuits may be checked without power being applied for resonant frequencies by inserting small radio-frequency meters at various points and coupling a radio-frequency oscillator to the circuits. If resonant conditions are found which indicate possible parasitic circuits, the necessary changes may be made without applying power. This method is useful for locating the circuits of highly damped "trigger" parasites.

An all-wave receiver is very useful for checking the transmitter output for spurious frequencies in addition to the normal harmonics. One must be careful not to misinterpret image frequency response if the receiver is located near the transmitter. An oscillating detector or auxiliary beat oscillator further increases the value of this method of checking parasites. A pure tone should result in listening to an unmodulated carrier from a transmitter. A rough tone usually indicates the presence of parasites.

The cathode-ray oscillograph is useful for studying low-frequency or weak parasites in the absence of the normal transmitter oscillation. Little coupling is usually required for obtaining sufficient deflection, and the oscillation frequency is easily obtained by comparison with a known sweep frequency.

Following is a list of common oscillator circuits (Fig. 1) in the order of their importance for analyzing the various parasitic circuits:

(A) The tuned-grid—tuned-plate oscillator circuit is found to be the basic circuit for the most common types of parasitic oscillations. There must be grid and plate circuits approximately tuned to the same frequency with capacitive feed-back through the grid-to-plate capacity of the vacuum tube. The oscillation may be stopped by excessive damping or by detuning of the circuits. The grid circuit is detuned preferably to a much higher frequency than the plate circuit for parasitic elimination.

(B) Grid dynatron oscillator circuit. In large tubes a negative slope of the grid-current characteristic corresponding to a few hundred ohms negative resistance (due to secondary emission from the grid) may exist with positive grid potentials and high positive plate potentials. The negative resistance of the grid may permit spurious oscillations in several grid circuits. In general the plate-circuit constants are not important in determining the oscillation frequency.

(C) Relaxation oscillation circuit. This type of oscillation usually occurs as a modifying feature of other types of parasites. Certain parasitic oscillations, as for instance, the tuned-grid—tuned-plate variety, may start and stop at a rate determined by the periodic charge and discharge of a grid-blocking capacitor through a resistance, similar to the operation of a relaxation oscillator. This type of oscillation is used in superregenerative receivers of the self-quenching type. An amplifier may have two different modes of high-frequency parasitic oscillation and change from one to the other at a low-frequency rate due to a similar relaxation effect.

(D) Hartley circuit.

(E) Colpitts circuit. Both of these oscillating circuits have a common tank circuit with inductive feedback predominating between the grid and plate portions of the circuit in contrast to the capacitive feedback in the tuned-grid—tuned-plate oscillator.

(F) Meissner oscillator. This is somewhat similar to the Hartley and Colpitts circuits in that it has a single tuned tank although the grid and plate circuits are inductively coupled.

(G) Electronic oscillators. These include the Barkhausen, Gill-Morrell, dwarf oscillations, etc. These internal tube oscillations require positive grid and approximately zero plate potentials.

The following diagrams show several types of parasites and these are simplified where possible by omission of neutralizing circuits and other circuit elements. Two very common high-frequency parasites will be considered first. A series parasite is arbitrarily defined as a push-

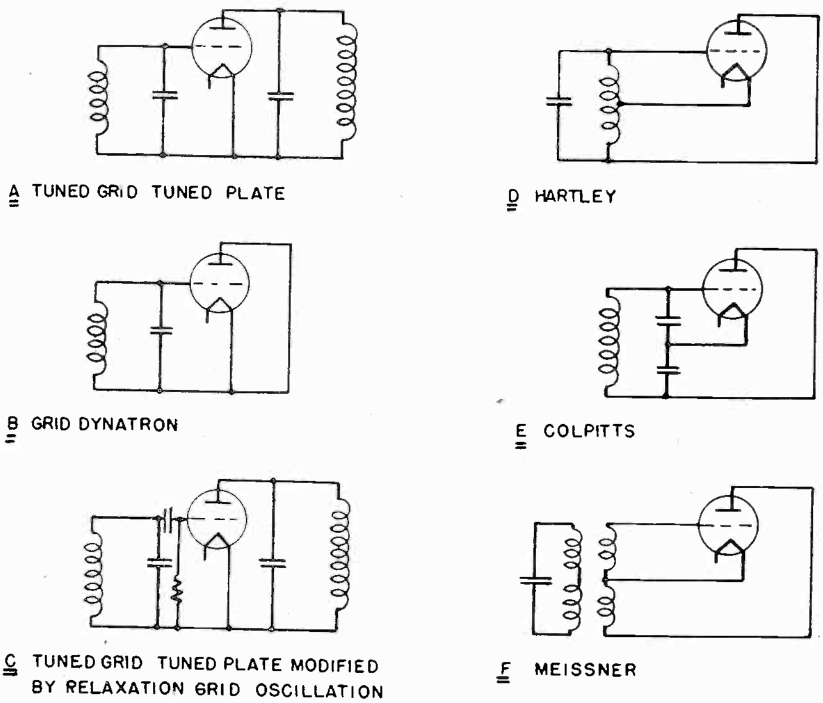


Fig. 1—Common oscillator circuits for parasitic analysis.

pull oscillation with the grid potential 180 degree out of phase. Similarly, a parallel parasite indicates the paralleling of tubes with the radio-frequency grid voltages in phase. The oscillation is seen to take place through the capacitor circuits and at high frequency we may consider the inductances of these circuits as being the grid- and plate-circuit inductances of a tuned-grid—tuned-plate oscillator. The interelectrode tube capacities are the coupling and circuit capacities. Due to the inductance of the neutralizing leads the circuit is not neutralized for the ultra-high-frequency parasite. These short-wave oscillations may be quite vicious and produce high voltages and currents. Fig. 23 shows a high-frequency parasitic arc standing on an insulator. If the parasite occurs in but one tube, the push-pull circuit or the tubes may be unbalanced and this is easily determined by checking the tubes and circuit constants. If the oscillation is a parallel parasite the grids or the plates may be connected together without appreciably affecting the oscillation. The neutralizing capacitors may prevent the series or push-pull oscillation if the neutralizing bridge happens to be balanced at very high frequency although this is a very remote possibility. It is obvious that the neutralizing circuit is of little use in preventing

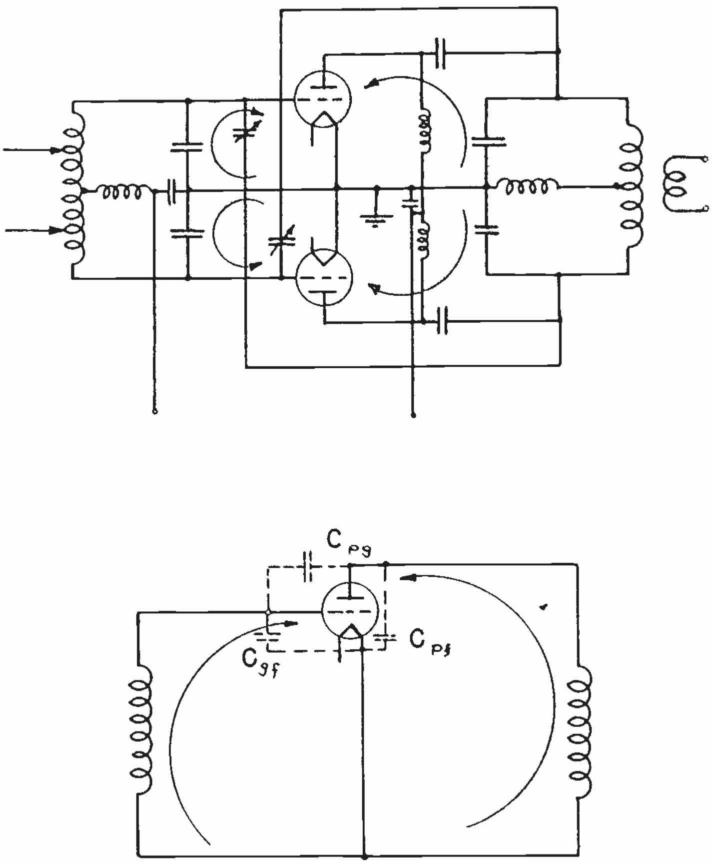


Fig. 2—High-frequency parasites.

paralleled tube oscillations since the neutralizing capacity adds to the grid-to-plate tube capacity in supplying coupling between grid and plate circuits. This will be shown in more detail below.

The high-frequency parasite may be eliminated in a number of ways. For class C amplifier circuits, resistors in the order of one to twenty-five ohms may be added close to the grid or plate connections of the tubes for damping the circuits. The resistors for damping the grid circuits should be noninductive wire-wound or preferably carbon-stick resistors. In class B linear amplifiers, and also with large tubes, it is not desirable to add very much series resistance in the grid circuit due to possible limiting on the positive peaks of modulation caused by grid current flowing through the resistor. In this case low resistant chokes may be added in parallel with the resistors to carry the direct current. It is also possible to eliminate the oscillation by detuning the grid and plate circuits so that the grid circuit is tuned to a much higher frequency than the plate circuit. This is accomplished by making the grid to filament circuit as short as possible, or in other words, by mounting the grid tank capacitor close to the tube. This lowers the inductance of the parasitic grid circuit. Small choke coils with one or two microhenries of inductance may then be added next to the plate connection so as to add inductance in the plate circuit. This method of eliminating high-frequency parasites is particularly applicable with tubes having the grid and filament connections brought out at one end. Also the plate circuit dimensions in practice are much larger than those of the grid circuit. Resistors may be added in parallel with the small parasitic plate chokes or the chokes may be wound with resistive wire if necessary in order to prevent "trigger" oscillations. However, sufficient detuning of the parasitic circuits usually accomplishes the same result. Simple choke coils next to the plate tend to improve the amplifier efficiency and reduce the harmonic components of the plate current. The parasitic plate choke may form a series resonant circuit with the plate-blocking capacitor at the fundamental frequency for improving the plate efficiency. Hence it is desirable not to dampen the parasitic plate chokes unless a trigger oscillation takes place in this circuit. In push-pull amplifiers the neutralizing circuit commonly used is shown in Fig. 2. If the neutralizing capacitor leads have appreciable inductive reactance the amplifier may become sufficiently unneutralized as to oscillate at a frequency near the normal operating frequency. Complex resonance conditions due to tapped circuit elements, coupled circuits, or chokes may further complicate the difficulty. In high power short-wave transmitters the large size of the apparatus and length of connections also tend to aggravate the parasitic trouble. The usual representation for the bridge neutralized circuit is given in Fig. 3A. This may be modified by considering inductance in the neutralizing capacitor branch of the circuit Figs. 3B and C, and it is evident that the bridge is not balanced except at one frequency where the net capacitive reactance in the neutralizing circuit equals the plate-to-grid capacitive reactance. If the grid tank circuit or excitation is connected at the mid-points S as shown in Fig. 3C instead of directly on the grid connections, and if the capacitor $C_n = C_{gp}$, the circuit will remain neutralized over a wide band of frequencies. Similarly if parasitic plate chokes are connected next to the anodes a balanced bridge may be obtained by making $C_n = C_{gp}$ and connecting the plate tank circuit to the points marked T in Figs. 3D and 3E so that the inductance of the parasitic plate choke between the plate of the tube and the point T equals the inductance of the circuit from the point T to the grid of the opposite tube. Care must be taken that the inductive reactance in either circuit

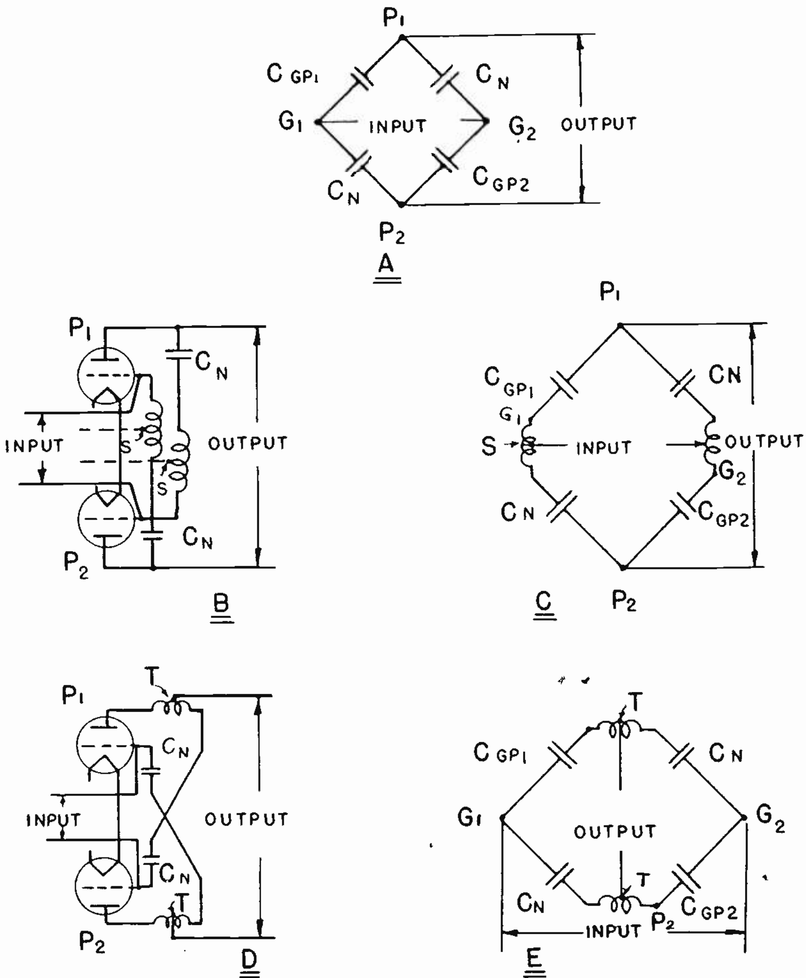


Fig. 3—Wide frequency band neutralization.

is small compared with the C_{gp} or C_n reactance and in very high-frequency transmitters this is accomplished by using short wide neutralizing leads.

The typical amplifier circuit of Fig. 2 is redrawn for illustrating two forms of low-frequency parasites where the tubes are in effect paralleled. Figs. 4A and 4B show the parasitic circuit with series-fed plate-voltage and Figs. 4C and 4D show the parasitic circuit with shunt fed plate voltage.

With a parallel parasite it is possible to connect both grids or both plates together without influencing the oscillation since these points are at the same potential. It should be noted that the grid circuit consists of an inductive branch through the leakage inductance of the grid tank coil and the grid choke to the cathode. The grid tank capacitors are in parallel. Also the stage is not neutralized for parallel parasites. In fact, the total feed-back capacity between the grid and plate circuits is the sum of the grid-to-plate capacity in the two tubes plus

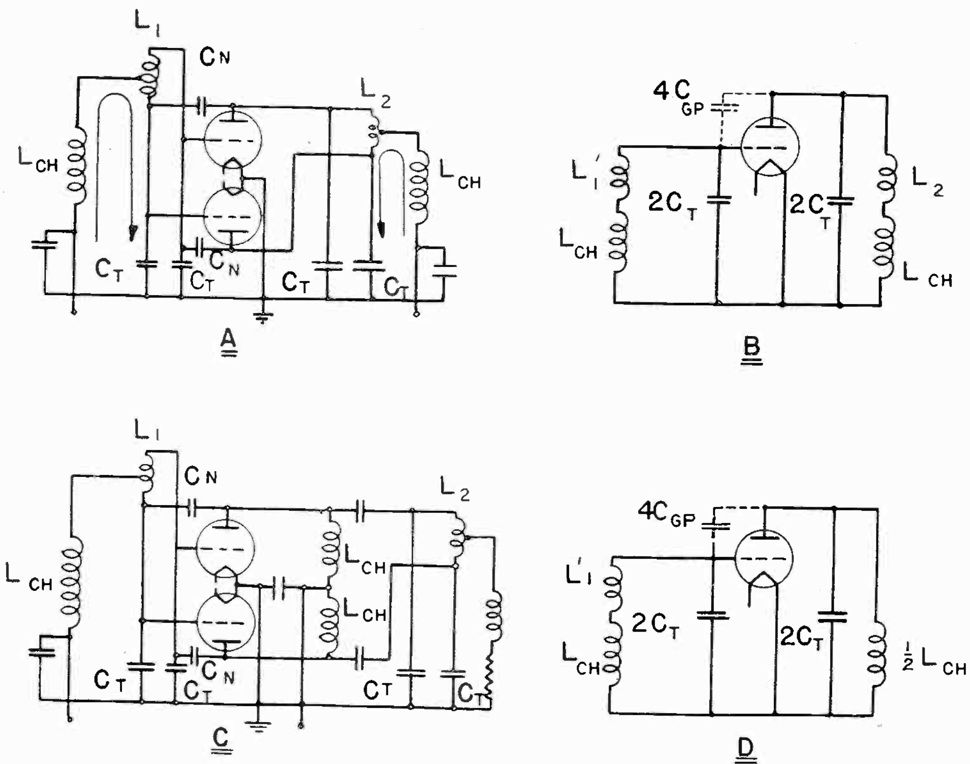


Fig. 4—Low-frequency parasites.

the two neutralizing capacitors. This large amount of feed-back capacity makes possible very low-frequency oscillations.

In a class B radio amplifier this parasite has been found to exist even with normal carrier excitation on the amplifier and to vary in amplitude with modulation.

Modulated carrier waves may be found on each side of the normal carrier at a spacing from the carrier frequency of plus and minus n times the parasitic frequency. The modulation envelope may be distorted and flashovers may occur in unloaded circuits due to high parasitic voltages generated at high percentages of modulation. These flashovers have been found to be most prevalent with low-frequency audio modulation which allows sufficient time for build-up of the oscillation. In a 50-kilowatt broadcast transmitter having a 30-kilocycle parasite, flashovers occurred at 200-cycle, 100 per cent modulation, but as the audio frequency was lowered, flashovers started at lower percentages of modulation. At thirty cycles the modulation was only about fifty per cent for flashovers. Apparently the build-up time of the low-frequency oscillation was appreciable.

The parasite may be eliminated by reducing the size of the grid

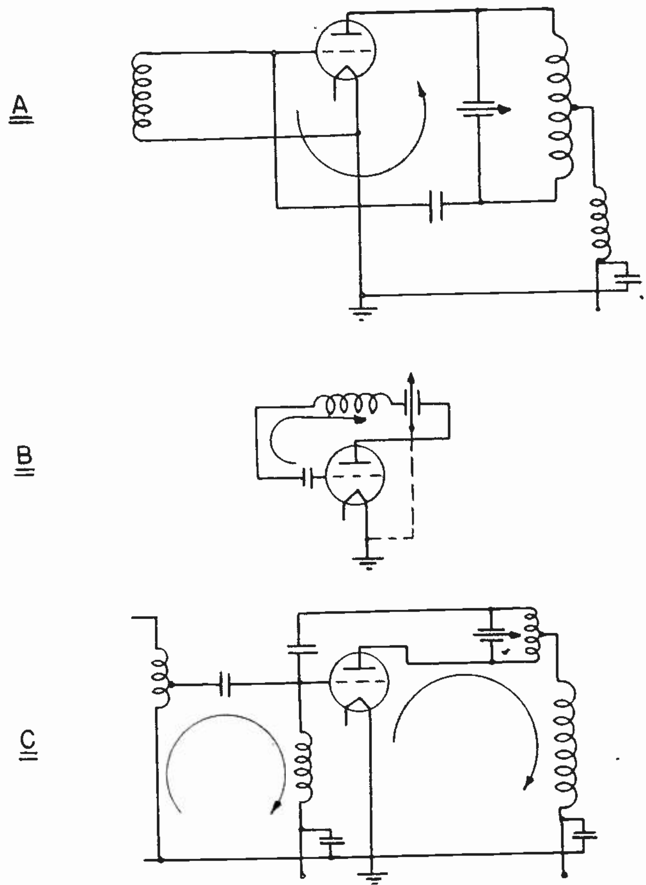


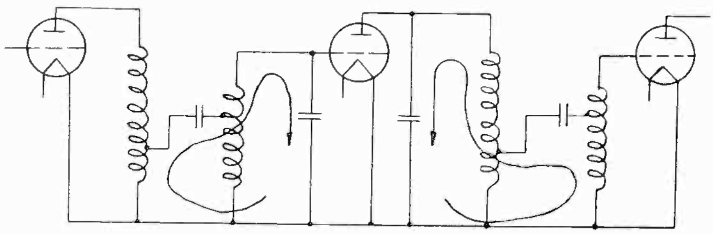
Fig. 5—Parasites due to ungrounded split-tank capacitors.

choke and by increasing the size of the plate choke. This follows the general procedure of tuning the plate circuit to a lower frequency than the grid circuit. Resistance may also be added in parallel with either choke, although in most cases this does not appear necessary.

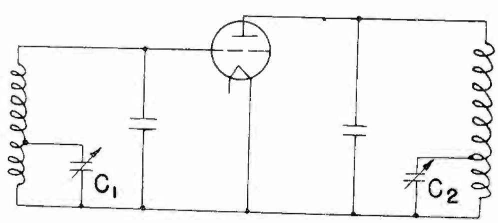
It is common practice to use a split capacitor with the center point grounded when a push-pull radio-frequency circuit is used or when a single tube is loaded into a double-ended tank circuit. If the capacitor is not grounded for radio-frequency potentials, a parasite may result in the circuit of Fig. 5A. The circuit as simplified is given in Fig. 5B and this is similar to the Colpitts circuit as modified for ultra-high-frequency oscillators. The current flows through the neutralizing lead and the plate tank capacitor. By-passing the center point of the tank capacitor to the cathode effectively shorts the plate circuit of the tube and this should stop this type of oscillation.

A low-frequency parasite may also occur in the single ended circuit particularly if chokes are used in grid and plate circuits. One possible circuit is shown in Fig. 5C.

This oscillation may be stopped by using a split stator capacitor with the rotor grounded, or by detuning and damping of the parasitic circuit as explained previously for low-frequency parasites.



A. CIRCUIT WITH EXCITATION AND LOADING TAPS



B. CIRCUIT WITH TAPPED TUNING ELEMENTS

Fig. 6—Parasites due to tapped coils and capacitors.

When taps for loading, Fig. 6A, or tuning, Fig. 6B, are used, additional circuits for parasitic currents are found. If the parasite is caused by the use of tapped coils for loading or excitation, it is apparent that detuning of the coupling circuits by the addition of reactance is indicated, or else a change to inductive coupling is necessary. When tuning capacitors C_1, Fig. 6B, are tapped across part of the tuning inductances for obtaining a vernier adjustment in tuning or to reduce the capacitor potential, a complex circuit is formed which offers resonance at more than one frequency. In an oscillator this may aggravate the drag loop tuning effect. In an amplifier a parasite may occur at a frequency for which the neutralization is ineffective. In general this method of obtaining a vernier control of tuning is undesirable, particularly if the capacitor C_1 is tapped across a relatively small part of the inductance.

If grid and plate transmission lines, or either, are connected directly to the tank coils, a parasite may be found at a frequency close to the normal operating frequency due to the fact that the transmission line offers multiresonance conditions and the amplifier oscillates at a frequency for which it is not neutralized, as described above. This parasite may sometimes be stopped by changing to inductive coupling or by detuning of the transmission line circuits so that the grid and

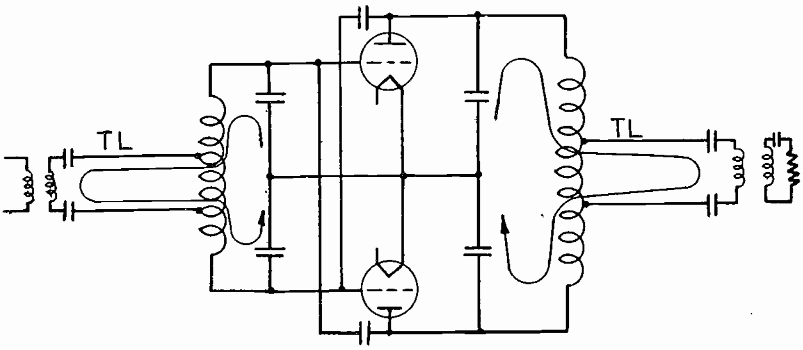


Fig. 7—Transmission-line parasites.

plate parasitic circuits are widely detuned. It may also be stopped by adding damping resistors at a point of high parasitic current in the transmission lines. In the case of a parallel parasite through both transmission lines to ground, the damping resistance can be added so as not to cause any loss at the normal carrier frequency.

If "shunt feed" of both bias and plate voltage supplies is used, considerable trouble may result from complex resonance conditions.

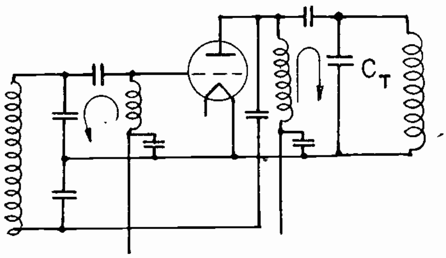


Fig. 8—Choke-coil parasites with "shunt feed."

The choke coils tend to resonate at various frequencies with the tank elements causing parallel or series parasites of the tuned-grid tuned-plate variety. After studying choke parasites, one is tempted to eliminate all choke coils, due to the fact that sufficient trouble can be had without tolerating choke parasites. If chokes are to be used in a transmitter for a specific purpose, as for instance, eliminating the direct plate voltage from the neutralizing and plate tank capacitors, then it is desirable to use the shunt feed in either the grid or the plate circuit but not in both. In the 500-kilowatt WLW power amplifiers (push-pull circuit) the shunt method of feeding plate voltage is used for these particular reasons and the plate chokes were very carefully resonated at the assigned frequency. Even then it was found that very high radio-frequency voltages existed at certain points along the choke coils and arcovers occurred to near-by objects. It was also noticed that considerable fourth harmonic of the fundamental radio frequency was being radiated from the antenna. A wavemeter near the transmitter showed almost as much deflection on the fourth harmonic as on the fundamental. All of these results were traced to a resonance condition in the grid transmission line and the plate choke coils at the fourth harmonic of the operating frequency. It was necessary to change carefully the number of turns on the coil and add resistance to the grid line in order to eliminate the difficulty.

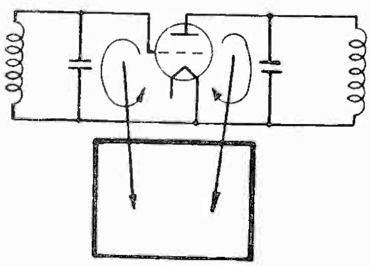


Fig. 9—External link circuit parasite.

Considering the Meissner oscillator circuit, consisting of an external tank circuit coupled inductively to a grid coil and to a plate coil, a parasite may occur in transmitters having additional circuits as a framework link coupled to the normal circuit elements. There has been found one case, where the internal framework of a set link coupled the grid and plate circuits to cause a parasite. This was eliminated by opening the framework circuit to prevent the link coupling.

With large water-cooled tubes operating at high plate voltage, a large amount of secondary emission from the grid may occur with positive grid potentials. The negative slope of the grid-current characteristic indicates negative grid resistance which will produce oscillations in conjunction with a tank circuit. The grid-current characteristic shown in Fig. 10 indicates a negative grid resistance which will produce oscillations over the range of A to B. The oscillation frequencies which occur are largely determined by the grid-circuit constants, and, therefore, the cure used for the tuned-grid—tuned-plate oscillation does not apply. In connection with dynatronic grid parasites, it should be noted that the grid-current meter will usually read negative grid current. This fact often helps in determining this type of parasite. There are several grid circuits or grid paths to ground, especially in a push-pull circuit, any one of which may cause a dynatronic grid parasite. Parallel and series oscillating circuits have been found. The inductance of the armature of a bias motor generator and its associated filter capacitor may control the frequency of a very low-frequency oscilla-

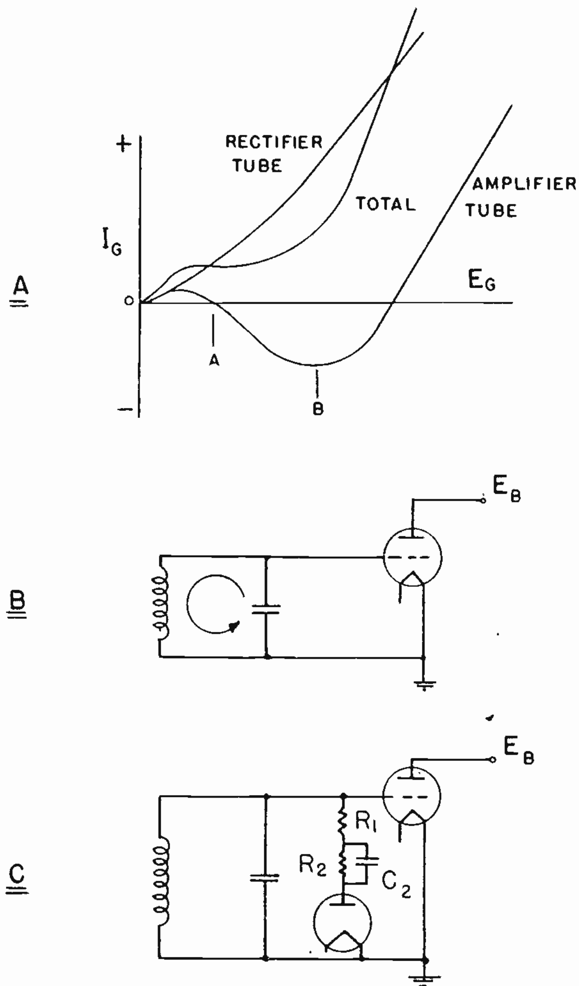


Fig. 10—Dynatronic grid parasite and dynatron rectifier.

tion of the radio-frequency carrier amplitude in a class B radio-frequency amplifier. In addition to all of the oscillations due to the various grid circuits on a tube, it should be emphasized that the amplifier may oscillate at the fundamental frequency even though it is perfectly neutralized, since the grid circuit oscillates at the grid tank circuit frequency and the oscillation is amplified as usual in the plate circuit.

Some success has been obtained in damping the oscillation circuit with resistance, but for complete protection it is most practicable to add a dynatron rectifier to the grid circuit which will draw positive grid current when the amplifier grid draws negative current. Series grid resistance is undesirable in class B radio-frequency amplifiers, especially with wide variation in the value of positive and negative grid current. A runaway effect in the operation of the tube may result from the use of series grid or biasing resistance when incipient gas flashes occur in large tubes, since the grid may suddenly assume a high positive potential.

With 100-kilowatt vacuum tubes, the dynatron rectifier, Fig. 10C, may be a two-element 50-watt size kenotron tube having resistance R_1 in series with the plate circuit in order to control the current drawn

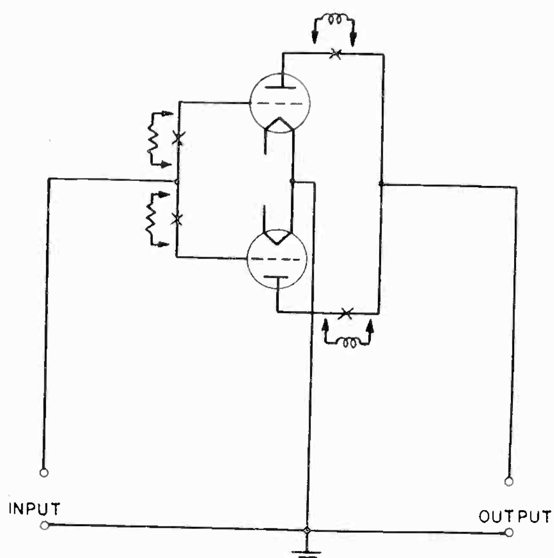


Fig. 11—Parallel connected amplifier tubes.

by the tube. The rectifier tube and resistor are connected between grid and filament of the amplifier tube, so that current flows only with positive radio-frequency grid potentials.

A series resistor R_2 shorted by a capacitor C_2 may be used to bias further the dynatron rectifier so that it does not draw current until the negative resistance region A-B is reached. The resulting over-all grid-current characteristic over this range may be made approximately zero or slightly positive as shown in the upper curve of Fig. 10A. The rectifier tube current may be emission limited for high positive values of grid voltage, and this is desirable to limit the peak grid load in a class B radio-frequency amplifier for minimizing audio distortion.

When tubes are paralleled it is well known that intertube parasites may exist. The oscillation is at very high frequency and of the series type. The frequency is determined by the inductance of the conductor between the grids, and the input capacity of the tubes. The plate circuit similarly has a conductor between the plates of the tubes, and, of course, a tube output capacity. It may be eliminated by connecting small resistors in the order of twenty-five or fifty ohms in series with each grid or the grids may be tied together with as short leads as possible, and small choke coils may be connected in series with each plate.

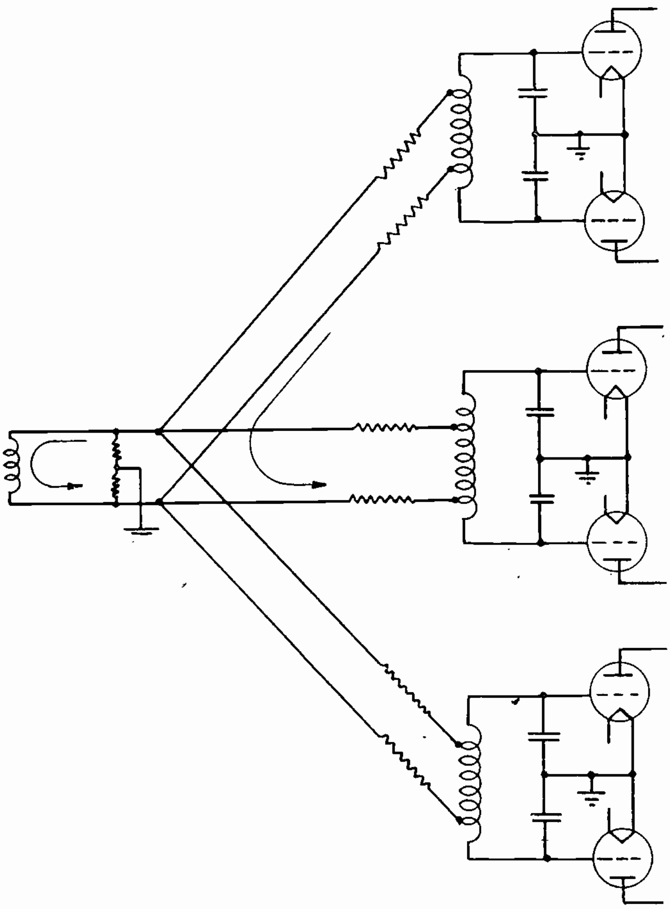


Fig. 12—Paralleled amplifiers.

In the WLW 500-kilowatt transmitter there are three identical amplifiers, each of which will deliver 170 kilowatts. The grid circuits are excited in parallel and the inductively coupled output coils are connected in series.

Additional parasitic circuits with the paralleled amplifiers were largely due to the common grid transmission line. A parallel parasite of all tubes was stopped by grounding the mid-point of the grid loading resistor. In addition to the common grid line circuit there are parasitic circuits between amplifiers. The plate chokes and tank capacitors formed the plate circuit for the tuned-grid—tuned-plate type oscillations. Damping resistors were added in the transmission lines to each amplifier as shown and these effectively prevented oscillations between amplifiers. There were two resistors in series in this circuit between amplifiers, yet for the excitation current from the driver stage the resistors are connected in parallel. This minimizes the power lost in the resistors at the fundamental frequency.

In a push-pull amplifier with transformers in the grid and plate circuits, high-frequency oscillations above the audio-frequency range may occur on peaks of audio excitation. In the high power class B audio modulator at WLW this oscillation was found at a frequency of 550 kilocycles.

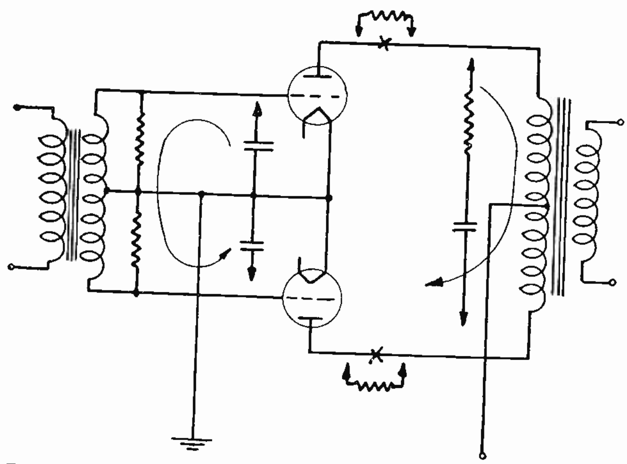
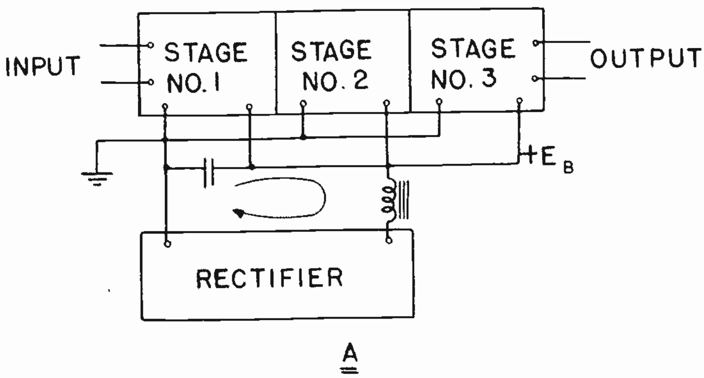


Fig. 13—Class A and B audio amplifier parasites.

The oscillation is commonly of the push-pull tuned-grid—tuned-plate type. The interelectrode tube capacities become significant at radio frequencies and the audio amplifiers are not neutralized in general practice. In general, audio transformers have natural resonant frequencies above the audio-frequency range where the impedance is very high or very low corresponding to shunt and series resonance. The high-frequency resonant points are in approximate harmonic relation as modified by the fixed stray capacity of the terminal bushing insulators and winding. Above the resonant frequency range the transformer impedance may be low and capacitive at a power factor determined by the characteristics of the insulation. Transformer impedances vary over wide limits with different input and output transformers in a class B stage and the oscillation frequency will be determined by the optimum circuit conditions.

There are many ways to stop the oscillation. The grid transformer secondary is usually loaded into separate resistances, but this does not necessarily prevent the oscillation if the grid leads are long and include a large loop area. The grid loop area may be reduced, or, better still, by-pass capacitors may be connected in parallel with the grid and filament connections of the tubes close to the tube connections for circuit detuning. Small resistors may be connected in series with each plate lead. Also a resistor may be connected across the primary of the output transformer close to the tubes in series with a capacitor which has high reactance at audio frequencies and low reactance at the parasitic fre-



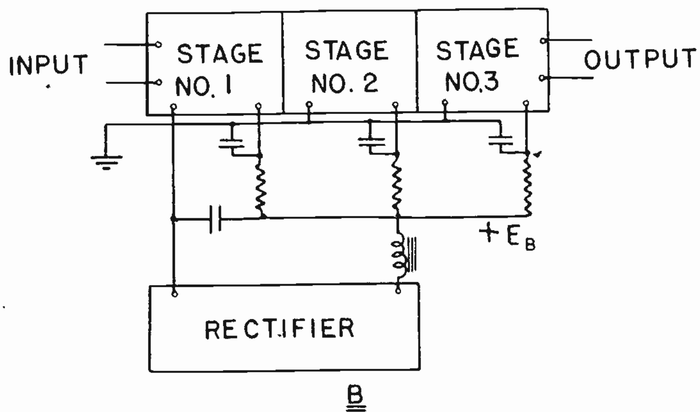


Fig. 14—Cascade amplifier instability.

quencies so that the resistor is effective in damping the output transformer primary at radio frequencies.

The input transformer in a class B amplifier should have low leakage reactance in order to minimize audio distortion and parasitic oscillations which may be aggravated by periodic commutation of the grid current.

In large transformer coupled class A amplifiers it is possible to have parasites similar to those described in the class B audio amplifier and the method of prevention is the same.

Cascade amplifier instability is a function of the amplification and interstage coupling, particularly between the output and input circuits. Frequently "motor-boating" or low-frequency oscillations may be caused by resonance conditions in filter circuits which are common to more than one stage. In resistance coupled class A amplifiers with a common plate voltage supply additional complications result, in proportion to the size of the interstage coupling capacitors and the grid resistors. It has been found that the RC product of these two values is limited to about 100,000 (ohms$\times\mu$f) in typical amplifier circuits for stable operation.

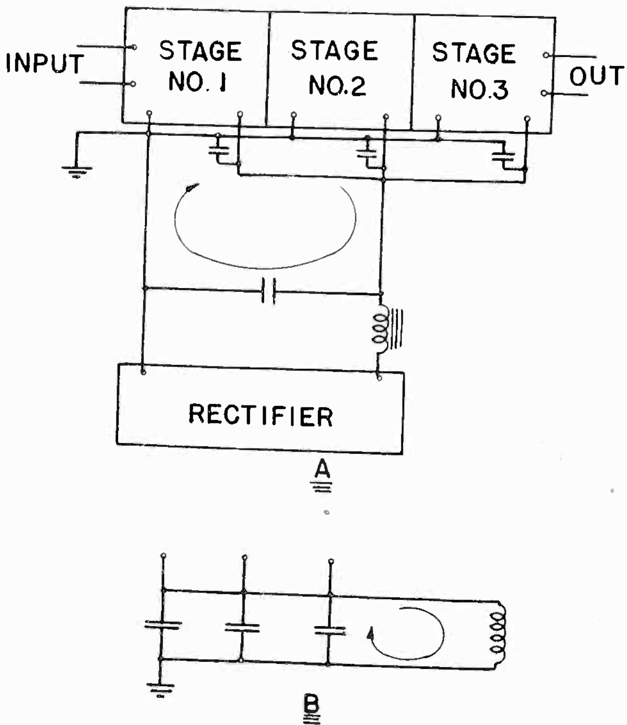


Fig. 15—Trigger parasites.

Decoupling circuits, consisting of resistors and capacitors, Fig. 14B, or separate filters, are effective in preventing instability between amplifiers having nominal gain Separate resistors are connected in series with the plate supply to each amplifier and each lead is by-passed on the load side in each amplifier. Separate power supplies may be required to prevent feedback from a high level amplifier into a low level amplifier.

If capacitors alone are used in by-passing common grid or plate power supplies, Fig. 15A, interstage coupling is possible at very high frequencies where the by-pass capacitors form parallel resonant circuits with the common connections, Fig. 15B. A small change in the length of a lead may produce stability. Nevertheless the decoupling circuits are most satisfactory for preventing instability.

Input circuits usually require careful shielding and filtering to prevent stray magnetic or electrostatic coupling, particularly from the radio-frequency output of a transmitter. Electrostatic shields between audio transformer windings are sometimes required.

In high power amplifiers requiring high plate voltage, an intermittent parasite may exist due to shock excitation by a transient of some circuit in the amplifier. The parasite may be a damped oscillation of any of the types mentioned.

It is conceivable that steady-state parasites may be eliminated but a high voltage oscillation may exist for a few cycles during a transient. If parasites are eliminated by circuit damping rather than by circuit detuning the possibility of transient or trigger parasites is reduced.

Very spectacular flashovers may result when a parasite excites a series resonant circuit. It is well known that the voltage across both of the series elements may rise to one hundred times or possibly one thousand times the applied voltage depending upon the ratio of reactance to resistance of the circuit. If a series resonant circuit exists in the plate circuit of a high power amplifier where the series inductance happens to be the plate lead to a tube, and the stray capacity of a coupled tank circuit to ground forms the capacitor element, it is possible to obtain extremely high voltage.

During testing work with a ten-kilowatt oscillator, a momentary flashover of two feet occurred from a secondary circuit to ground when the plate voltage of seven kilovolts was applied to the oscillator. The arc was thin and blue in color and only occurred for a fraction of a second on the starting transient. Apparently this type of parasite may be stopped by circuit detuning or damping as previously described. In the test oscillator a small capacitor from grid to cathode eliminated the trouble.

There is another type of surge in radio transmitters which simulates the phenomena of a "trigger" parasite. This is the internal flash arcs of large vacuum tubes. The flash arc has been called "gas flash," "Rocky Point effect," and "arcback," although the latter designation is commonly reserved for mercury-vapor tubes. The internal tube flash is primarily a function of potential gradient and gas pressure in high vacuum tubes. Potential gradient is a function of the potential between elements and of the curvature of the tube elements. The random occurrence of the flash arc in vacuum tubes, frequently not concurrent with modulation peak potentials, indicate spasmodic gas-pressure variations. The intensity of the visual flash arc varies from a barely noticeable incipient breakdown in the tube producing a momentary audible ping, to an intense power arc resulting in a virtual short circuit between tube anode and cathode or grid accompanied by a loud metallic sound from the anode. It is possible to burn out the filament and grid wires with a sustained arc. The amplitude of the short-circuit plate current in large tubes must be controlled by a series plate resistor, Fig. 16B, and the high voltage power supply to the tubes should be opened or discon-

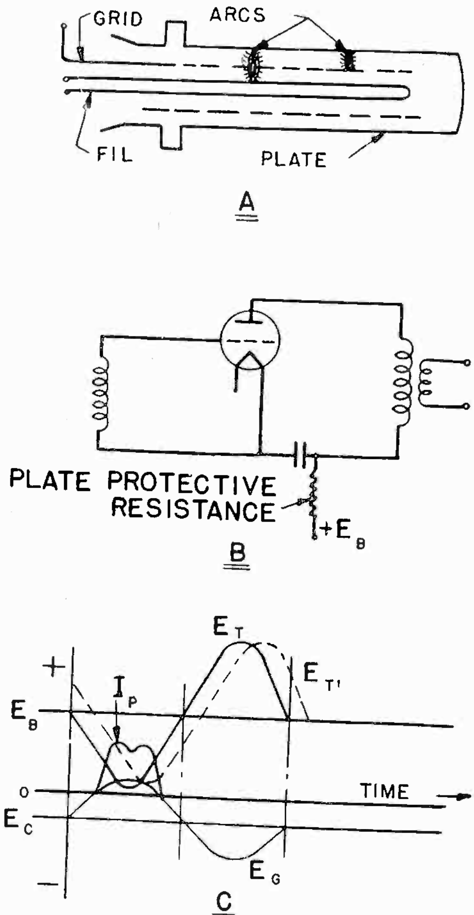


Fig. 16—Flash arcs in vacuum tubes.

nected very quickly if the energy dissipated within the tube is to be kept within safe limits. Protective plate resistance in the order of 100 to 200 ohms for a 10-kilowatt tube and 20 to 40 ohms for a 100-kilowatt tube has been found desirable. The tungsten filaments of large tubes permit gradual clean-up of gas and a circuit tuning condition has been found advantageous, especially with new tubes. If the plate circuit is tuned to a slightly lower frequency than the normal operating frequency to produce a leading power factor of plate current and a five to ten per cent increase in the direct plate current in a class B or C radio-frequency amplifier, the clean-up action of the filament is apparently improved. The plate tank inductance or capacity is increased slightly in order to achieve the leading power factor. (Fig. 16C.)

Frequently a screen-grid tube amplifier will oscillate when connected in the usual circuits even with the screen grid suitably by-passed to the filament. The tube may oscillate at the fundamental frequency due to stray coupling between the grid and plate circuits, Fig. 17A. Grid and plate choke coils or other circuit elements may be mounted

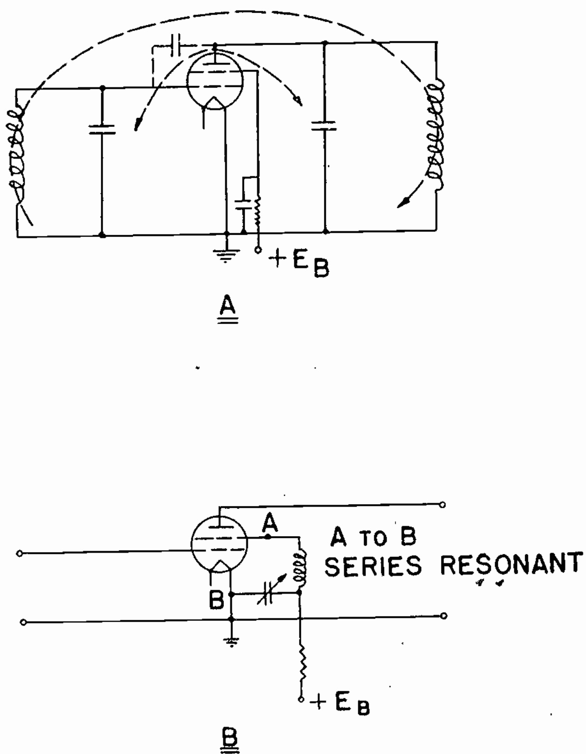


Fig. 17—Screen-grid tube oscillations.

too close together or the grid and plate circuits may be coupled through paralleled leads. If the screen-grid by-pass capacitor is too small, or if the lead connecting it to the cathode is too long, also, if the frequency is too high, the screen grid may not be effective in preventing feed-back coupling through the control-grid-to-plate capacity in the tube. It has been found that a variable capacity may be substituted for the fixed screen-grid by-pass capacitor as a possible means of preventing this instability at the higher frequencies of operation. The screen-grid radio-frequency potential may then be reduced to a low value by series resonance of the external circuit to the cathode. This does not mean series resonance of the circuit including screen-grid-to-cathode capacity as this would give high radio-frequency screen-grid voltage, and permit a tuned-grid—tuned-plate oscillation with an ineffective screening element.

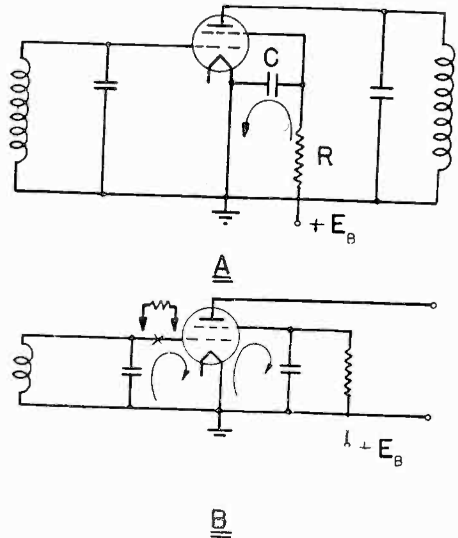


Fig. 18—Screen-grid tube parasites.

If the screen grid is supplied with voltage through a high value of series resistance from the plate-voltage supply, and if the excitation is low in a class C amplifier, an unstable oscillating condition may result which is similar to a relaxation effect. The frequency is determined by

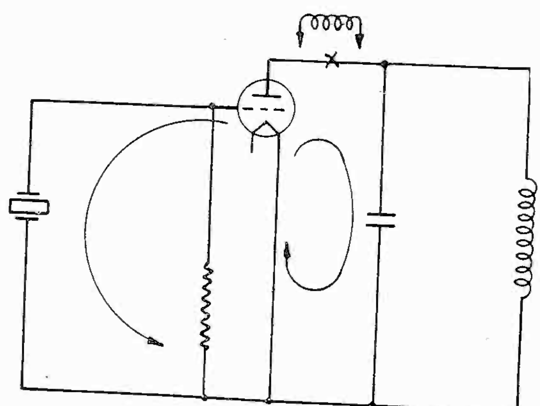


Fig. 19—Crystal oscillator instability.

the *RC* time constant of the screen-grid circuit. This condition may be corrected by using a voltage divider for the screen-grid voltage supply, so as to prevent excessive changes in the screen-grid voltage. Another type of parasite, Fig. 18B, consists of an ultra-high-frequency oscillation where the screen grid becomes an anode and the oscillation is of the tuned-grid—tuned-plate type as shown. This is stopped by adding inductance or resistance to the grid circuit either to detune or dampen the circuit.

When a very high-frequency crystal is used, the spacing between the plates becomes small and the shunt capacity, therefore, increases. If the leads between grid and filament of a crystal oscillator tube are too long, a high-frequency parasite may exist between the grid and plate circuits as a tuned-grid—tuned-plate oscillation where the grid inductance consists of the long grid lead. The plate-circuit inductance consists of the plate-tuning capacitor lead. The tube interlectrode capacities form the capacitive elements of the tank circuit. It is desirable to keep the grid circuit short and lengthen the plate circuit, possibly by adding a choke coil next to the plate of the tube. Inductive feedback is reduced by separation of the grid and plate leads and by

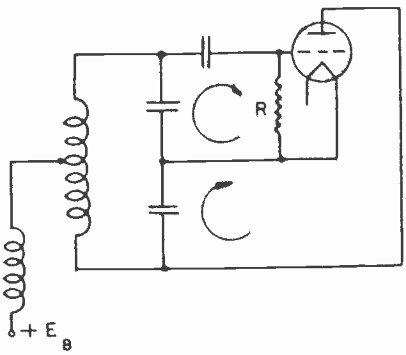


Fig. 20—Oscillator parasites.

decrease of the area of the grid-circuit loop. Certain crystals have abrupt frequency changes due to a slight change in temperature. One cure is to reduce the crystal current to a low value to prevent heating by the use of a pentode oscillator. Another is to change to the new low temperature coefficient crystals which permit even greater output from the crystal oscillator.

One form of superregeneration detection depends for its action upon the use of a high resistance grid leak so that the oscillations in the tube start and stop at a superaudible rate as determined by the value of the grid leak and capacitor. This type of composite oscillation might be considered a form of relaxation oscillation. The interruption frequency is determined by the values of the grid leak and blocking capacitor. An oscillator may break into this superregeneration or intermittent blocking, due to the grid-leak resistance being too high but it may also be due to the fact that the ratio of grid excitation to plate tank voltage is too high, or in other words, because of overexcitation.

Oscillators of the tuned-grid—tuned-plate, Hartley, and Colpitts types have parasitic circuits through capacitor leads and chokes as described above for amplifiers, and similar methods of prevention apply. The Meissner oscillator parasites easily at a frequency determined by the grid and plate coils and tube capacities, especially with loose coupling to the external tank circuit.

In receivers, the loud speaker, together with variable condensers, tubes and other elements, may cause an objectionable howl due to

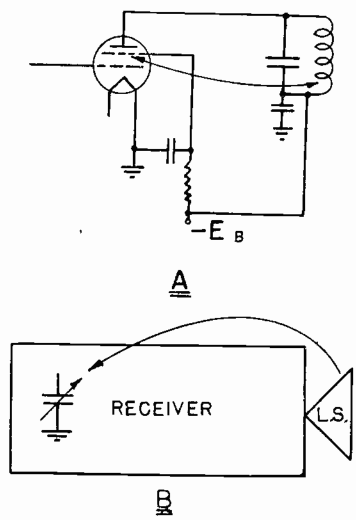


Fig. 21—Mechanical resonance oscillation.

mechanical coupling. Similarly, the tube elements in transmitting tubes, and the coils in large transmitters may sometimes mechanically control the frequency of an audio oscillation. This oscillation may also

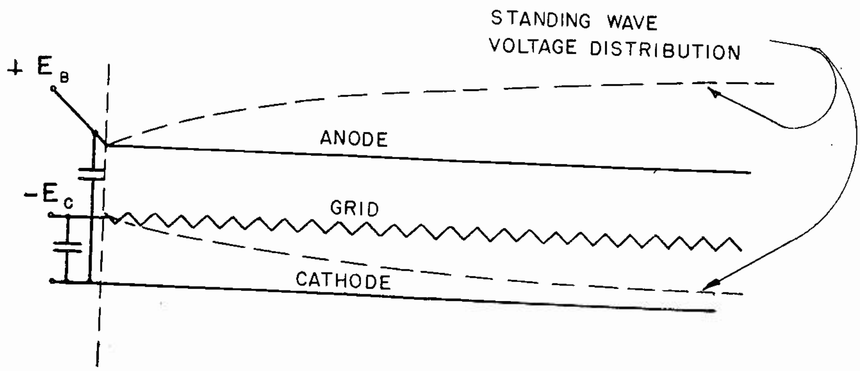


Fig. 22—Standing-wave parasite on plate of tube.

depend upon the transmission of sound through air or other medium for coupling between mechanically vibrating elements. In any case the solution to the problem is quite simple because of the fact that the sound may be heard. The mechanical coupling or damping may be changed sufficiently to prevent the oscillation by the use of shock mounting or more rigid construction. In the diagram shown, (Fig. 21A) vibration of the screen-grid tube was found to cause an audio oscillation due to mechanical coupling between the tank coil and the tube. The oscillation was easily stopped by more rigid mounting of the coil.

About six years ago an interesting parasite was discovered at WGY with developmental 100-kilowatt tubes. If both grid and plate leads are by-passed to the filament, standing waves of voltage at an

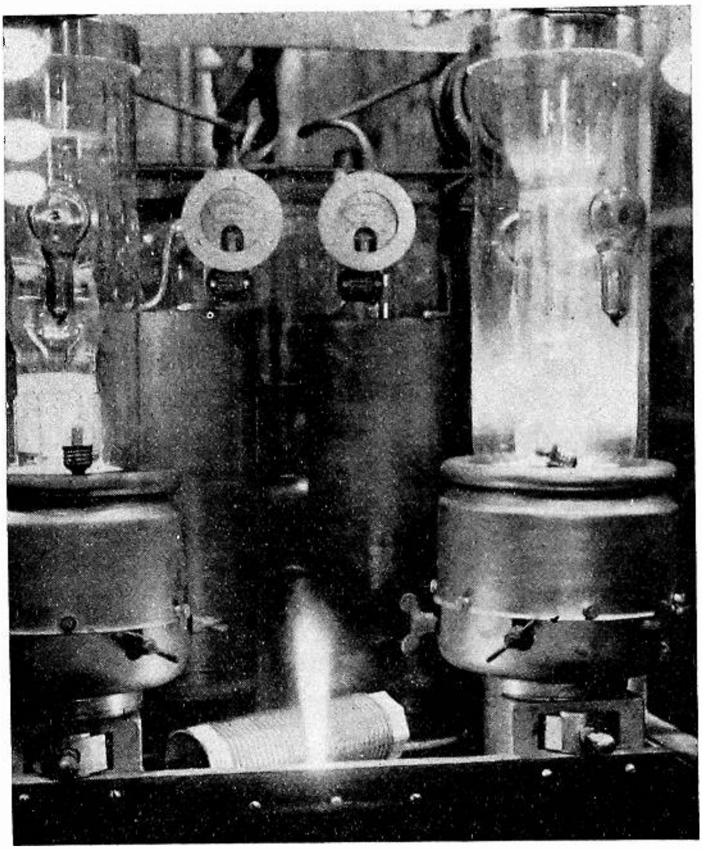


Fig. 23—High-frequency arc standing on an insulator.

ultra-high frequency may be found on the tube elements. The voltage is, of course, low at the by-passed end and high at the free end of the tube (Fig. 22). The frequency of oscillation was about 50 megacycles. Many kilowatts can be developed in this way although the circuit is rather inefficient because of the fixed loss in the lower voltage end of the tube. This parasite may be cured by removal of either by-pass capacitor and by inserting an inductance in the plate lead as described above for elimination of high-frequency parasites.

Precautionary Note

The study and elimination of parasites require close adherence to safety precautions because of the attendant exposure to high power circuits while making test changes. High-frequency parasitic currents have been found to break down the insulation in safety interlocks and in start-stop switches. The moral is to use adequate safety devices in working on any circuit with voltages over three hundred volts. A red light which shows when the high voltage is on, gives a certain protection but is an indicator and not a safety device. All high voltage filter condensers should have bleeder resistances and best of all a safety switch should be used to short-circuit high voltage filter condensers when working on the transmitter.

Conclusions

In the elimination of parasites from a transmitter, the circuits should be kept as simple as possible to prevent complex resonance conditions. Radio-frequency choke coils and shunt-feed circuits should be kept at a minimum. Wide band neutralization circuits are desirable. The grids of vacuum tubes should be effectively by-passed capacitively to the cathode through a capacity, and inductance added next to the plates of the tubes to eliminate short-wave parasites. If necessary, the plate or grid parasitic circuits should be damped with resistance. Inductively coupled rather than capacitively coupled input and output circuits should be used wherever possible. The mechanical layout of a short-wave transmitter should be well planned with short leads and compact tank circuits to keep the current where it is supposed to be and to minimize stray coupling between circuits.

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DEVELOPMENT OF TRANSMITTERS FOR FREQUENCIES ABOVE 300 MEGACYCLES*

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Summary—*The fundamental functions of the electrons and their work cycle, in the interelectrode space of high vacuum tubes, are discussed. It is shown how the triode feed-back circuit becomes inoperable at very high frequencies due to space-time and reactance characteristics. It is further shown how the space-time conditions can be organized to benefit the maintenance of oscillation instead of becoming a detriment. Some of the more familiar arrangements, such as the Barkhausen and the magnetron circuits, which are based on this principle, are discussed in some detail. With these illustrations as a background the author describes a new method of frequency multiplication at very high frequencies. This method yields much greater power outputs than hitherto possible and promises to become very useful.*

Various means for frequency stabilization are referred to and the merits of frequency controlling devices, such as crystals and low power factor circuits, are compared.

Special problems encountered in the application of modulation at very high frequences are described and reference is made to methods developed to meet these problems.

Practical considerations of circuit arrangements are described in some detail. Several examples of transmitter design are given. These sections are illustrated with photographs.

Important points in connection with antennas and transmission lines are discussed and the results of some measurements are given.

The paper ends with a brief reference to some propagation results obtained by RCA Communications engineers and others.

Introduction

THE purpose of this paper is to report progress in theoretical conception as well as in practices pertaining to the application of the frequency band between 300 and 1000 megacycles to radio communication.

The Electron Performance in High Vacuum

Since electrons are negative electric charges, their presence in the interelectrode space of a vacuum tube causes, by virtue of electrostatic induction, positive charges to be distributed over the electrodes which will vary in accordance with variation in position of the electrons. If the external circuit consists of a resistance it can be seen that the cur-

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rent produced in this resistance, by virtue of the electron motion in the interelectrode space, always causes a voltage drop on the electrodes which retards the electron motion and decreases the rate at which electrons enter the interelectrode space.

When the electrons land on the electrodes, the positive charges in the electrode meet the corresponding electron charges and are cancelled. The current caused by the electron motion therefore ceases at the moment the electrons land. Whatever kinetic energy the electron possesses at the time it lands is lost in the form of heat.

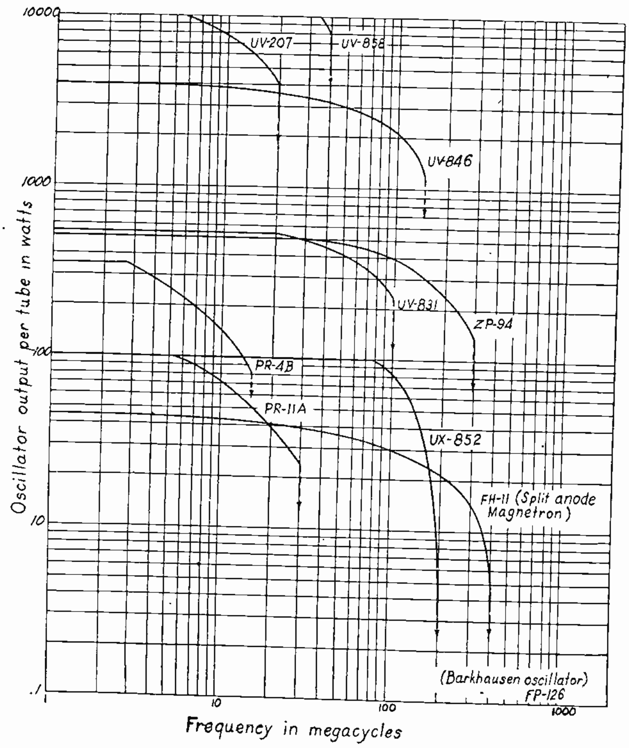


Fig. 1—Frequency versus power output at rated anode dissipation of some American transmitting tubes. Curves not otherwise indicated refer to performance in triode feed-back oscillators. These curves were originally published by W. C. White in the *General Electric Review*, September, (1933).

The Electron Performance in High Vacuum at High Frequencies

When the well-known, triode feed-back oscillator is adjusted for higher and higher frequencies, a frequency will eventually be reached beyond which the device fails to perform. (Fig. 1.) It is usually assumed that the increased circulating currents necessary to maintain proper electrode potential across the decreasing capacitive reactance of the interelectrode space cause prohibitive losses. This is, however, not the major factor in well-designed circuits and the chief limitations are instead to be found in the vacuum tube itself.

Hitherto it has been possible to neglect the interelectrode transit time of the electrons in vacuum tube phenomena. The finite velocities of the electrons introduce phase lags in the electron motion which are unsuitable to the triode feed-back method and results in a reduction of efficiency. Accompanying this phase lag electrons are trapped in the interelectrode space. As the transit period terminates, the electrons in the grid-cathode space come to a stop. The grid potential is rapidly becoming very negative and assumes a controlling effect upon the field in the grid-cathode space. This field therefore changes direction and the electrons in the grid-cathode space are thrown back into the cathode at high velocity, resulting in high kinetic loss. Since the plate becomes more positive, while the grid grows negative, the field in the grid-anode space is increased. The electrons in this space therefore receive additional impetus in the direction of the anode and arrive there at high velocity and thus with a high kinetic loss. At the lower frequencies the interelectrode spaces are "cleaned out" before the potentials have had time to reach excessive values. The higher the frequency the greater becomes the number of electrons which fail to accomplish the transfer or which transfer under field conditions which cause excessive kinetic losses. The existence of these losses has previously been referred to but not explained.[1] Since a great portion of the grid input energy, due to trapped electrons, appears at the cathode, its value can be observed by noticing the increase in cathode resistance from the increased cathode temperature. Estimates of the loss obtained in this way indicate that it is a major source of frequency limitation in conventional transmitting tubes.

In order to reduce the losses during the "cleaning-out" period, the interelectrode space must be made small in volume so that it contains a small number of electrons in transit. If the power output is to be reasonably retained and since the cathode emission at the present cannot be increased, the cross section of the interelectrode space cannot be excessively reduced. The only way to obtain substantial decrease of volume is thus to reduce the length of the interelectrode space. While this on one hand results in increased capacity with the handicap of higher circulating currents it improves the phase relation between the moving electrons and the electrode potentials. Thompson and Rose[2]

[1] F. B. Llewellyn, "Vacuum tube electronics at ultra-high frequencies," Proc. I.R.E., vol. 21, pp. 1532–1574; November, (1933).

[2] B. J. Thompson and G. M. Rose, Jr., "Vacuum tubes of small dimensions for use at extremely high frequencies," Proc. I.R.E., vol. 21, pp. 1707–1722; December, (1933).

have had rather outstanding success in compromising these factors in the design of receiving tubes for very high frequencies.

Generation of Higher Frequencies

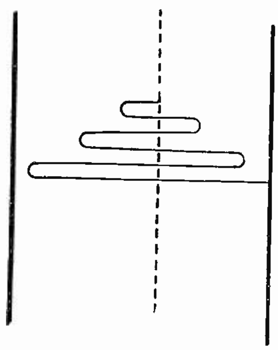
In the triode feed-back circuit the electron flow takes place during a very short favorable portion of the oscillation cycle. Due to the finite velocity of the electrons it has been shown that it is not possible to confine the existence of electrons in the interelectrode space to such short portions when the frequency is increased. As a result, and because of the nature of the circuit, the oscillating power created in the external circuit is returned to the electrons and lost in the form of heat. It is, therefore, necessary to choose methods in which the electrons may perform usefully during more prolonged portions of the oscillation cycle; in which the time of travel of the electrons and the electric fields produced by the electrons themselves contribute to the condition of oscillation. In other words, the electrons themselves and their motions constitute the whole oscillator. All such oscillators require electric, and often also magnetic, field conditions such that the electrons as a group are subjected to unstable conditions which can produce whistle effects in the interelectrode space. The periodic pressure effects in an air whistle would thus correspond to the potential effects set up by the periodic formation of concentrations in space charge.

These conditions may be obtained if the electrons are given an opportunity to miss an anode as they are accelerated toward it. After missing the anode and as the electrons are thus carried away from it by their own momentum, they will be subject to a retarding force instead of an accelerating force from the anode. They will eventually come to a stop and again be accelerated toward the anode. The positive direct-current potential on the anode thus makes the electrons describe a pendulum motion. Since there are many electrons and thus many such pendulums they cannot be permitted to oscillate at random phase since they will then cancel each other's effect upon the external circuit. The pendulum motions must, therefore, be organized to operate in unison. By comparison with the traffic congestion on a highway which occurs at points where the traffic speed is reduced, it is easily seen that congestion of electrons will arise in the regions of the interelectrode space where the electrons turn around. As these accumulations form, the resultant electric field in the interelectrode space is gradually being altered. This alteration influences the motion of the electrons. Also, electrons arriving later at the turning region, retard the turning around of the earlier ones, while, on the other hand, the earlier electrons speed up the turning around of the later ones. This condition is therefore conducive to synchronization of the oscillating electrons so that they form into groups. When the oscillating electrons form into groups they will thus no longer cancel each other's influence upon the external circuit. The losses from the currents induced in the external circuit by the group motion will, therefore, introduce a load upon this motion, and the motion of the individual members of the oscillating electron group will become attenuated. This attenuation takes place in the direction toward the prime mover, the anode, where the electrons are ultimately deposited. Since they are continually being replaced by newly emitted electrons of high momentum, the motion of the group as a whole is not attenuated but may be represented by an average of the motion of its continually changing members. It should also be clear that many of the electrons will be subject to accidents so that they will prematurely collide with the anode. There is no electrode system known where this can be prevented. The fact should, however, be noted that while such collisions represent very great losses the kinetic energy so spent is not taken from the oscillating energy in the external circuit as the case happened to be at the end of transit in the triode feed-back oscillator. Practically all of the momentum possessed by the electrons has been derived from acceleration by the interelectrode, direct-current field. Since the cathode is inherently located in a region where the electrons reverse their motion and since the space charges, there forming, appear periodically, it is clear that the emission from the cathode will also be subject to periodic fluctuations. This phenomenon and the synchronizing tendency between the individual electron pendulums as they approach the turning regions amplify one another and establish a reasonably substantial tendency for the electrons to form a whistle effect in the interelectrode space.

The best known methods utilizing these principles are the Barkhausen and the magnetron methods. In both these arrangements the electrons are made to turn around before reaching the plate. In the Barkhausen method an ordinary three-element vacuum tube may be used. The grid is highly positive and acts alternatingly as accelerator and decelerator for the electrons which pass through. The plate is mostly operated at a slightly negative potential to facilitate the reversing of the electron motion before the electrons reach the plate. The path of an individual electron is shown in Fig. 2.

The magnetron, as is well known, consists in its most common form of a centrally located cathode which constitutes the axis of a cylindrical anode. The anode has a high positive potential and the electrons are made to miss it by virtue of the deflecting properties of an axial magnetic field. The path of an individual electron is shown in Fig. 3.

One very interesting phenomena in common for both the Barkhausen and the magnetron methods is the difference in the behavior of the oscillations when the tuning of the external circuit is approached from a state of lower or higher circuit tuning. This phenomenon is due to the fact that the voltage inducing effect from the electrons is twofold. The electrons accumulating near the plate or near the cathode cause voltages to be produced on these elements by electrostatic induction. The voltage drop across the resistive component of the external circuit is, however, a maximum when the electrons are in a state of highest velocity. This voltage is therefore ninety degrees in phase lead of the voltage set up by the electron accumulations. The combined voltage on the electrodes therefore tends to make the electrons turn



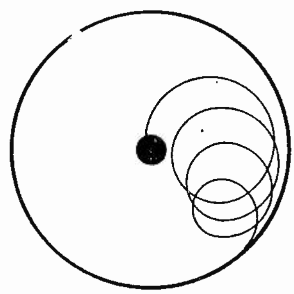


Fig. 2—Attenuated path of the individual electron when partaking in organized group motion in a triode pendulum oscillator.
Fig. 3—Attenuated orbit of the individual electron when partaking in organized group motion in a magnetron oscillator.

around sooner. The frequency is increased. Since the frequency of oscillation thus increases as the resistance between the electrodes is increased, a peculiar effect occurs when the tuning of the interelectrode circuit is varied. If the circuit resonance is made to approach the electron oscillation frequency from a lower value, the electron oscillation in the tube will recede upward and will thus have to be trailed by the circuit tuning. If the circuit tuning is adjusted above the electron oscillation frequency and then lowered, the oscillation frequency moves up to meet it. In the Barkhausen case this phenomenon has been called Gill and Morell oscillations.

Effects in the Vicinity of Electrodes

In addition to the effects of the electron motion so far considered it may also be of interest to consider the local effects in the close vicinity of an electrode.

The electric field from the electron, which at a distance covers the electrode fairly uniformly and causes a current of no definite origin to flow through the electrode, as the electron moves, becomes more concentrated as the electron nears the electrode. The current origin in the electrode becomes more and more defined into a spot under the electron where it becomes very concentrated. The direction of this current is toward the spot if the electron is in an approaching state and away from the spot if the electron is in a receding state. These considerations, of course, do not apply to electrons moving parallel with the electrode surface. This phenomena is naturally extremely rapid in that such concentrations do not become noteworthy until the electron is fairly near the electrode. It takes place during a very small fraction of the total transit time of the electron. Its period is therefore greatly in excess of that represented by the transit time and represents real ultra-high frequencies. These "surface oscillations" are independent of the frequency at which the device operates and depend only on number and velocity of the electrons.

Carrying the discussion a little further it becomes increasingly difficult to see where to draw a line between these "spot impulses" and heat quanta. It depends largely upon the size of the area under consideration if the period belongs to the radio-frequency region or the heat region. If the electron is headed for a landing on the electrode the spot becomes smaller and smaller until we reach the molecular and atomic structure of the electrode where the remaining kinetic energy is interchanged.

If the electron does not approach the electrode quite so close, like for instance when an electron passes through a grid structure the frequency produced, while high, is definitely one far below that of heat.

Not being organized these impulses cannot be shown in the external circuit under ordinary conditions. It may, however, be possible, by special methods to set up conditions, by using a very restricted number of electrons, whereby these oscillations may be shown.

Negative Resistance

The expression "negative resistance" is very commonly used in explaining oscillatory phenomena. Reference to it in the previous discussion has been avoided until sufficient background could be established for a clear understanding of its nature. It has been seen how in the triode feed-back circuit maximum electron transit is obtained when the anode potential is at a minimum. As the anode voltage decreases the current through the tube increases. In an electron pendulum device the electrode toward which electrons are moving has its highest negative tendency as the electrons possess their greatest radial velocity.

Therefore, as the electrode voltage decreases the current through the tube increases. In the so-called dynatron method, similar current-voltage conditions are obtained with the aid of secondary emission. The electrons are usually made to pass through a grid of high positive potential toward a plate of less positive potential. The electrons will, therefore, land on the plate with considerable velocity, causing other electrons in the plate to bounce off and be attracted by the grid. As the positive potential on the plate is increased the oncoming electrons will have a higher velocity and thus cause more electrons to bounce off and travel in the opposite direction. The total current will decrease because the ratio between electrons coming to the plate and leaving it has been decreased. All these methods, therefore, have the one characteristic in common that current and voltage vary inversely. This is characteristic of negative resistance. Since the operation of the dynatron is not based on time delay in electron transit this method, in its fundamental principle, is not adaptable to generators of extremely high frequencies. More or less developed dynatron action is, however, often obtained in conjunction with other methods whenever electrodes are bombarded by high velocity electrons.

Frequency Multipliers

Since the electron pendulum methods are critical to fields and space charges, each type of tube has a fairly well established maximum output at a particular frequency. The outputs from conventional sizes of tubes are limited to a very small portion of the power output of which they are capable when operating at frequencies where the time of electron transit is of no significance. It was, therefore, considered that if the frequency of the greater amounts of power possible to produce at the lower frequencies could be multiplied efficiently, greater power may also be realized at the higher frequencies.

The commonly known vacuum tube frequency multipliers, which are widely used for various purposes, consist of a triode, a circuit connected to the grid which is tuned to the fundamental frequency and a circuit connected to the plate which is tuned to a harmonic frequency. The grid has a negative bias and the plate has a positive potential. The frequencies used are usually well below the values at which the time of electron transit assumes significance. This shortness of the transit time is in fact an asset since the production of harmonics depends on the immediate establishment and discontinuation of electron current as the grid potential passes above and below the cut-off value. By virtue of the plate power the device also operates as an amplifier.

As the input frequency is increased, the electrons will remain in the process of transit while the electrode potentials may vary considerably. Like in the triode feed-back circuit this is very detrimental. As the grid potential goes below cut-off value, it first retards the electrons moving toward it from the cathode. The electrons therefore deliver the energy they chiefly received from the direct-current source into the external oscillating circuit and the grid receives an impetus in accord with its negative swing. As the grid continues to become more negative the electrons will finally come to a standstill whereafter they will be accelerated back toward the cathode. They will then draw power from the external circuit. The acceleration received in this direction is greater since the grid is now very negative. The total effect of the motion from the cathode and back again is, therefore, a loss. The electrons in the grid-anode space received added acceleration in their original direction as the anode potential swings in a more positive direction. This accelerating power is thus also obtained from the oscillating energy in the external circuit. Both the fundamental frequency circuit and the harmonic circuit, therefore, lose more and more of their power as the frequency is increased.

Due to the increase in potential gradient toward the central cathode in cylindrical element tubes the greatest effects upon the motion of the electrons are, of course, obtained in this region. Electrons which emerge from this region have a more predetermined motion for the rest of their path than electrons in flat element tubes. This condition makes the electrons passing through the interelectrode space of cylindrical tubes less sensitive to variation in electrode potentials, during a great portion of their journey, than in flat electrode tubes with uniform fields. This fact, of course, holds regardless of the methods of oscillation employed.

At the lower frequencies the grid simply acts as a throttle which lets electrons through to the anode during a short portion of the cycle. The electron accelerating power required for the production of this motion and the subsequent harmonic it induces is thus chiefly taken from the direct-current plate supply. As the frequency is increased the number of electrons which do not complete their transit from cathode to anode becomes a greater and greater portion of the total number of electrons which enter the interelectrode space. The power stored in these electrons, as they are pushed toward the electrodes, is taken from the alternating-current component of the external circuit. It is, therefore, clear that as the frequency is increased, the frequency multiplier becomes less of an amplifier. The fundamental frequency input, instead of being only a guide to the electron performance, becomes a driving force. Since the frequency multiplier outputs at higher frequencies must derive the driving power from the fundamental frequency input it was necessary to find a way by which the motion of the electrons, thus driven, could be organized to converge, so that they would, during a certain phase of their motion, fall through an appreciable portion of the interelectrode space in the short time corresponding to a harmonic current. It was found possible to accomplish this by introducing a magnetic field which is perpendicular to the electric field. In a cylindrical tube this then becomes an axial magnetic field. The effect of the magnetic field may be most easily understood by referring to the curves in Figs. 4, 5, and 6. For the sake of simplicity a tube with

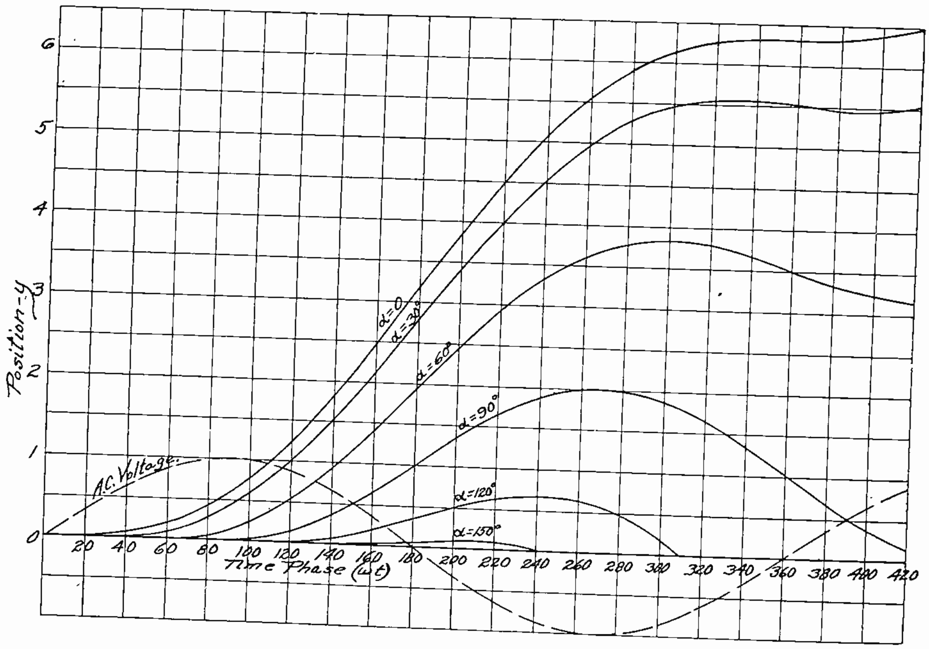


Fig. 4—Position versus time of electrons in an alternating-current field between flat electrodes. Each curve represents electrons which have left the cathode during a certain phase of the alternating-current cycle.

two parallel plane electrodes is being considered. Alternating voltage is applied of such frequency that the time of transit is too slow for crossing the interelectrode space. The positions of the electrons leaving the cathode at different times, under the influence of the positive half cycle of the voltage on the other electrode, will be distributed throughout the space as shown in Fig. 4. From these curves it can be seen that there is a tendency for the electron positions to become more and more divergent with time. Even a rectifying action can be observed in the motion of the electrons which leave the cathode during the first half of the positive half cycle. If a magnetic field, perpendicular to the electric field, is introduced it can be given such a strength that the electrons which leave the cathode successively, during the generous time period of a half cycle of the fundamental frequency, will all complete their return journey to the cathode within a very definite time limit. (See Fig. 5.) By integrating the velocities, the curve shown in Fig. 6 is obtained. It represents the shape of the resulting electron current curve. A transformation of the fundamental frequency power directly into a harmonic component thus becomes possible. The electron stream exhibits a positive resistance to the fundamental frequency current

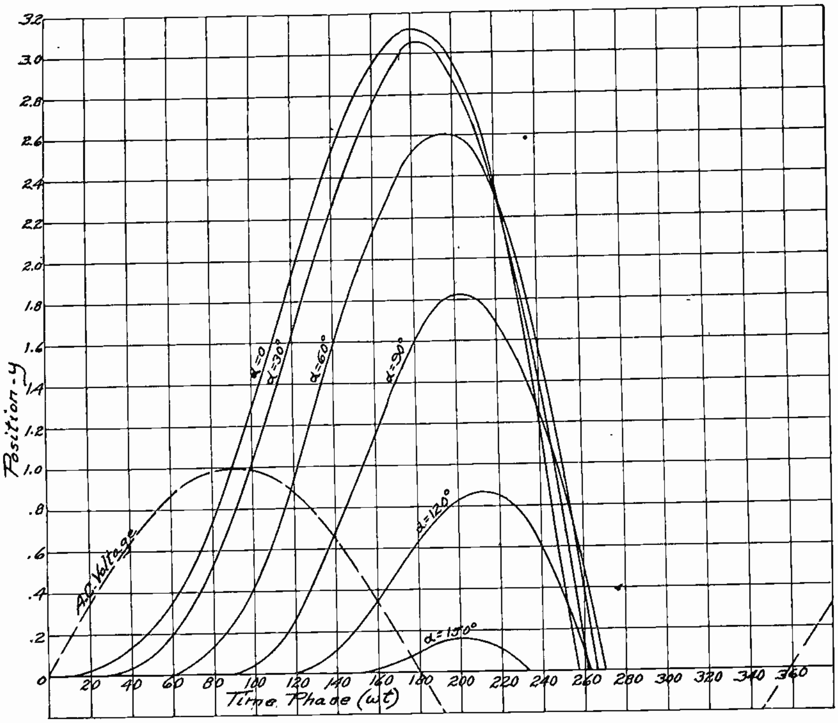


Fig. 5—Organization of positions shown in Fig. 4 when introducing a steady, transverse magnetic field. Note how the electrons, returning to the cathode, converge in respect to time.

and a negative resistance to the harmonic. As the electrons return to the cathode they still possess considerable kinetic energy. If it is attempted to lower this kinetic energy by further circuit loading, the disorganizing effect from the load prevents further gain in efficiency. The efficiency also depends largely upon which harmonic is chosen. While the efficiency obtained when producing the second harmonic is high, the efficiency of the generator of the fundamental frequency oscillations is low since it has to operate nearer the border limits of the triode feed-back circuit than when a higher harmonic ratio is used. The multiplier efficiency, however, drops very rapidly as the harmonic ratio is increased. Fewer electrons become subject to complete organization. When odd harmonics are produced, two tubes can be operated in push-pull fashion. Very simple circuit arrangements are then obtained. From these considerations, the third harmonic has often been chosen as the most satisfactory compromise.

Among the standard tubes available, some three-element tubes happened to be the most satisfactory for frequency multiplication. Examples of such tubes are shown in Fig. 7. Although it has been found most efficient to let all the elements become part of the circuit either for bias purposes alone, or for tuning purposes as well, the cathode and the grid are, of course, the essential elements between which the major

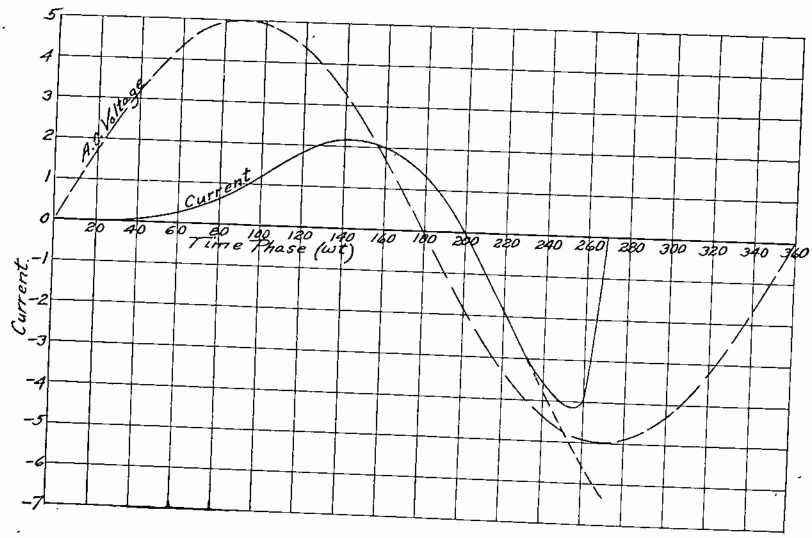


Fig. 6—Electron current curve corresponding to conditions shown in Fig. 5.

electron performance takes place. The grid is negatively biased and the plate is given zero or negative potential. The input and the output circuits can be applied to any one of the elements as long as the combined circuit tuning produces potentials in coördination with the desired electron motion. Different values of magnetic field strength are sometimes necessary when choosing various modes of connection.

The efficiency obtained with standard, cylindrical element triodes as triplers is about ten per cent. The efficiency gain by using a magnetic field varies but has so far been found to be at least three times that obtained without magnetic fields.

Two advantages result from the use of frequency multipliers. The power obtainable with a tube used as a frequency multiplier is many times greater than that which it would deliver as a pendulum oscillator. Frequency control circuits of great accuracy may be employed.

Frequency Control

One of the advantages of very high frequencies for communication purposes is the insignificant width of the modulation band as compared with the frequency of the carrier. This makes it possible to consider a great number of communication channels within frequency limits only a few per cent apart. In order that full advantage may be taken of this situation it is, however, clear that as the carrier frequency is increased its relative stability must also be increased. Aside from this consideration, very accurate frequency control is required to permit maximum selectivity in the receiver in order to obtain optimum signal-to-noise ratio. The transmitters have been required to provide stability sufficient for reception by means of very selective superheterodyne receivers, without loss of signal quality. Special efforts have been made to produce oscillator outputs as free as possible from undesirable amplitude and frequency variations. Since the difficulties of eliminating these disturbances increase in proportion to the oscillator frequency, the utmost of every available means for frequency stabilization had to be mobilized. Vital circuits must be well shielded against electric or magnetic influences of variable character. It is also very important that the frequency determining parts of the circuit be mechanically separated from their surroundings in order to avoid disturbance from vibrations. In coupling up with output circuits such as transmission lines and antennas, great care must be taken that a minimum of coupling is being used. The reaction from variation in the constants of such circuits, due

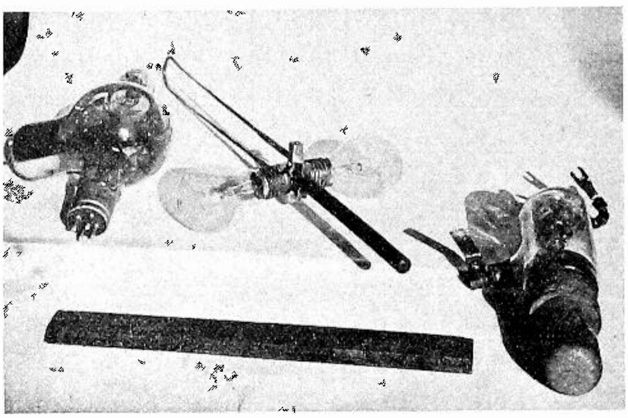


Fig. 7—Left: UX-852 transmitting tube; very useful all-around tube for triode feed-back circuits (Fig. 1), triode pendulum, and frequency multiplier circuits.
Middle: 300–600 megacycle wavemeter and output indicator. Note the two 50-watt load lamps.
Right: RCA-846 (UV-846) water-cooled tube, useful for triode feed-back circuits (Fig. 1) and for frequency multipliers.

to wind, precipitation, etc., will otherwise become very disturbing. Self-bias, by means of leak resistances and self-rectification often results in a tendency to maintain constant tube impedance if introduced in correct porportions in various branches of the tube circuit.

It has often been found necessary to provide smoother anode potentials than normally required at the lower frequencies. Variations in the direct-current supply have been eliminated by means of a regulator consisting of a combination of resistors with a vacuum tube, Fig. 16. The grid of the regulator tube is connected to a potentiometer inserted between the direct-current power supply and ground through a source of constant negative bias such as a battery. The function of this bias source is to overcome the positive voltage drop in the potentiometer so that the grid may be maintained at proper operating potential. The anode of the regulator tube draws current from the direct-current power supply through a resistor. The oscillator power is supplied through the same resistor. As an example, an increase in voltage across the direct-current power supply makes the grid of the regulator tube less negative. This increases the current through the common regulator resistor through which both the anode of the regulator tube as well as the anode of the oscillator are supplied. An increase in the voltage drop across the regulator resistor thus takes place. If the device is properly adjusted, the rise in voltage is completely compensated by the drop across the regulator resistor. A front view of such a regulator device is shown at the right in Fig. 14.

In addition to such measures it is very desirable to incorporate special frequency controlling devices such as crystals and low power factor tank circuits. Since crystals, of reliable performance above a few megacycles, are difficult to obtain it is necessary to use many stages of frequency multiplication in conjunction with this type of master oscillator. Low power factor tank circuits can be applied more directly and, since they otherwise compare favorably with crystals, they are more practicable as frequency stabilizers at very high frequencies.

Modulation

In modulating an oscillator it is sometimes quite difficult to obtain pure amplitude modulation which is not distorted by variations in phase or frequency of the oscillation. This difficulty is rather outstanding with the electron pendulum type oscillators.

Higher degrees of purity may, however, be obtained by resorting to compound modulation, i.e., by applying the modulation energy in at least two different ways and in such a manner that components representing undesirable modulation effects are cancelled. In a Bark-

Lausen oscillator it is, for instance, possible to obtain amplitude modulation substantially free from frequency variations by modulating the plate and grid cophasially with correct proportions of modulating voltage. An increase in the positive grid potential will tend to increase the electron velocity while a simultaneously applied decreased negative potential on the plate will increase the length of path. The frequency of the electron oscillation may thus be kept constant while the flow of the electron current is being modulated. In order to obtain linear amplitude modulation over as large a range as possible, the power output of the nonmodulated carrier must, of course, be reduced since the maximum output of electron oscillators is very definite.

In master oscillator frequency multiplier types of transmitters, undesired frequency modulation is less prevalent, but may sometimes still be desirable to eliminate inasmuch as any variation in the load on the master oscillator will influence its frequency stability. Compound modulation, of various types, may thus be used to advantage.

One very effective way of producing amplitude modulation is to control the power output by interposing a modulator between the transmitter and the antenna. For producing a steady tone, such as required for interrupted continuous wave telegraphy, a commutator across the transmission line is very effective. For more universal purposes a vacuum tube circuit is, however, required. At frequencies above 200 megacycles most vacuum tubes now available are not capable of furnishing efficient conditions for electron transit. In the description of electron pendulum oscillators it was pointed out that the orbital motions of the individual electrons are subject to decay as they deliver their energy to the external circuit. In the case of an absorber the electrons must instead absorb energy from the external circuit and deposit the so-acquired kinetic energy in the form of heat on the electrodes or the electrons must be so organized that they will serve efficiently as intermediary links by which the energy is efficiently directed into resistive load circuits. If the electrons themselves are to be made to absorb energy, conditions must be set up which give to the electrons a natural period, or the tendencies for such a period, near the period of the energy to be absorbed. In such a case, the orbital motion of the electrons will be of increasing amplitude so that its kinetic energy will increase. This principle has also been verified in the development of apparatus for the production of high speed ions for the purpose of bombarding atomic nuclei. Periodic oscillations of increasing amplitude are then given to ions instead of electrons.[3]

[3] E. O. Lawrence and M. S. Livingston, "The production of high speed light ions without the use of high voltages," *Phys. Rev.*, April 1, (1932).

Circuit Considerations at Ultra-High Frequencies

If the electromagnetic field, established around an electrical circuit carrying oscillating energy, has dimensions comparable to the wavelength, energy will be radiated.

Radiation may be reduced by using a balanced or symmetrical circuit, one whose adjacent, corresponding parts give rise to electromagnetic fields of equal amplitudes at opposite phase. The circuit can actually be built symmetrical as is done in push-pull circuits or as in two-wire transmission lines. An orginally nonsymmetrical circuit may be located near a conductive surface, Fig. 8.

In the ultra-high-frequency technique, linear conductors are often used for tuning instead of circuits with lumped constants. The relative merits of various combinations of tuning circuits may sometimes be judged by the total number of resonance points obtained. The fewer

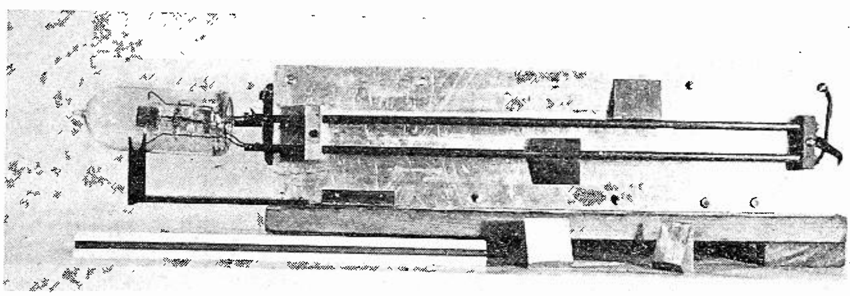


Fig. 8—FP-126, single tube, triode pendulum oscillator. Note the arrangement of the nonsymmetrical circuit near a shield and the wedge-shaped blocking capacitors which serve as tuning sliders. They are wedged between the conductor and the shield. The surface facing the shield is lined with mica.

such degrees of freedom, the better is the circuit. These resonance points arise from many conditions. At very high frequencies, the tube elements and the internal leads of a vacuum tube form an appreciable portion of the total tuned circuit. When continuing these elements and their leads with linear conductors, of dimensions giving constants which match, no additional degree of freedom is obtained. When using external circuits with lumped constants, the total circuit does, so to speak, more definitely consist of several distinctly separated reactance components. These components may neutralize each other in various combinations, creating several resonance points. An additional source of multiple degrees of freedom is also provided by the coupling phenomena created between the various circuits by the capacity combinations between the vacuum tube elements. This effect adds resonance points to linear tuning systems as well as to tuning systems with lumped constants. The resonance points so far referred to, are usually quite close in frequency. The linear type of tuning circuit may, in addition to these, respond to octaves or harmonics of the frequency to which they are fundamentally tuned.

The choice of tuning method is thus determined by many factors. For low frequencies, the linear tuning method becomes impracticable on account of the large physical dimensions required. As the frequency is increased and it becomes possible to use linear tuning, a great number of harmonics may be obtained if the oscillator is still capable of producing an output at the harmonic frequency. By using circuits with lumped constants the harmonic feature may be more easily avoided. The trouble with coupling frequencies may be minimized, or avoided, by so adjusting the circuit that only one of these resonant

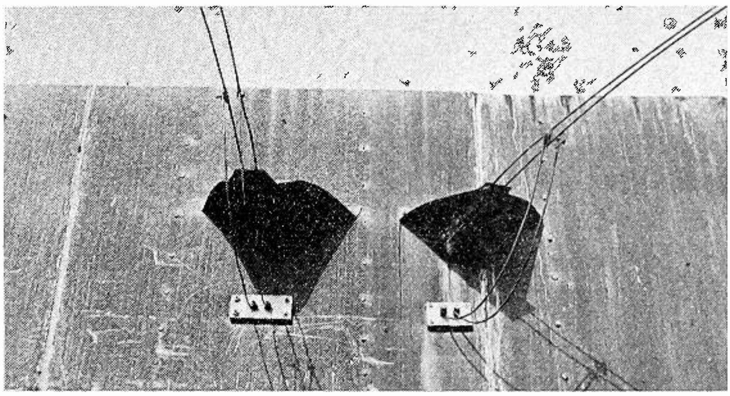


Fig. 9—Transmission line lead-in detail, showing high impedance metallic suspension links sometimes called "metallic insulators."

frequencies conditions, such as voltage and phase, become of controlling influence. At still higher frequencies, where special methods are used and in which the direct-current potential of the electrodes, in a larger measure, controls the frequency, the choice of tuning methods becomes more arbitrary. It may then be determined by other factors such as mechanical features, cost, space required, etc. When the frequency is so high that the first voltage nodal point appears on the lead inside the glass envelope of a vacuum tube, or in some other inaccessible place, the external circuit must be given characteristics equivalent to the addition of a conductor extending half a wave, or a multiple thereof, from this nodal point. Such arrangements are especially successful in master controlled circuits which are void of the ability to self-oscillate.

Linear conductors may, at very high frequencies, replace insulating supports. Such conductors must then be given a certain length so that their impedance is very high at the point of support. For this reason they must also be arranged to be nonradiating in the same manner as linear tuning circuits. A quarter-wave conductor with one end grounded will have high impedance at the other end. For the support of a push-pull circuit a pair of such conductors may be used. Such a pair then corresponds to a U-shaped conductor half a wave long from end to end, with an electrically neutral point at the bend. The losses incurred from the use of these metallic insulators are usually much lower than obtainable with insulating material. Their mechanical strength is superior and their disturbing effect upon the electrical circuit is smaller.

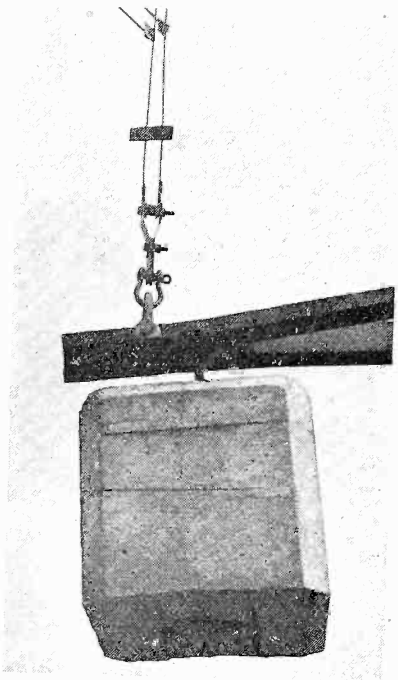


Fig. 10—Bend in a transmission line showing application of metallic insulator.

They have been found particularly useful in transmission-line work down to frequencies as low as fifty megacycles. Applications are illustrated in Figs. 9 and 10.

It has been found that stranded wires cause very high losses. The reason is that the twisting increases the inductance of the individual strands. At very high frequencies a considerable portion of the current will, therefore, force its way over the shorter and lower reactance path from strand to strand either capacitively or conductively. Since such conductors are seldom clean, the losses at such crossovers are very high. This point requires special consideration in connection with some of the standard transmitting tubes which are equipped with stranded connecting leads. It is best to eliminate these leads whenever possible.

For obtaining a direct grounding effect on the filaments of a tube it is sometimes necessary, for reasons already mentioned, to extend the length of these leads to make them one-half wave long. In less conventional circuits certain improvements in the phase and voltage conditions between the electrodes may be obtained if it is possible to tune the filament circuit. Whether linear conductors or circuits with lumped

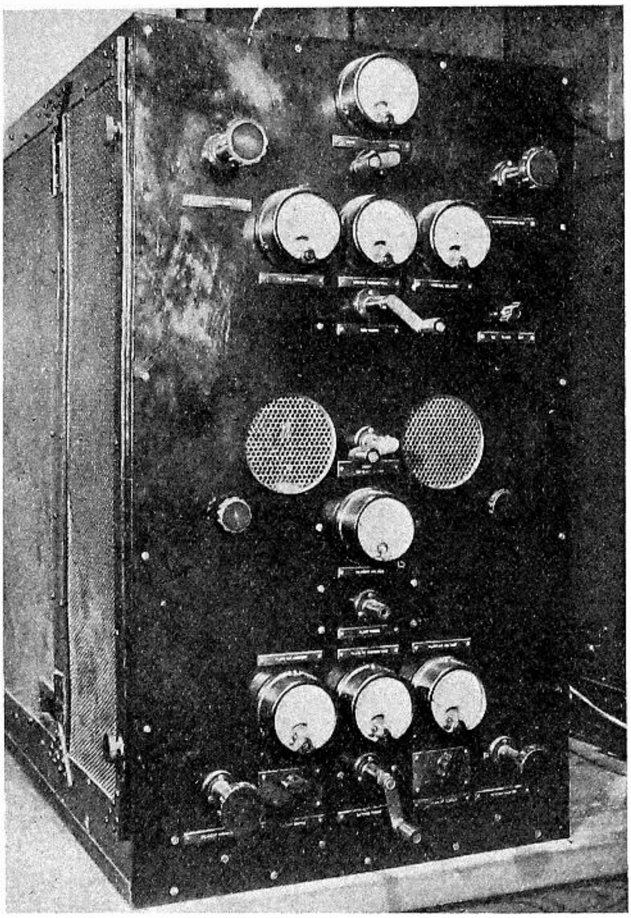


Fig. 11—Front View of 462-megacycle, 6-watt output, UX-852 tube, push-pull transmitter of the triode pendulum oscillator type.

constants are used, the easiest way of supplying the filament heating current is to make the radio-frequency conductor of tubular material and to locate the leads for the heating current inside this conductor. Small blocking condensers should then be used at the filament end to prevent radio-frequency currents from entering the heating circuit. Such arrangements are illustrated in Fig. 18.

When using larger tubes, of the water-cooled type, the water can be supplied to the tube jackets through metal tubing which also serves as tuning conductor. The supply and return water may be handled either through concentric or independent tubings. The latter is sometimes convenient in that the two parallel tuned circuits thus obtained allow independent branches for tuning and coupling. (See Figs. 17 and 18.) If these leads also carry high direct-current potential the water circuit is continued through a rubber hose of adequate length. This hose is usually attached at a point neutral to radio-frequency voltages.

During experimental work it may often be found difficult to get a new circuit going. Since it is much easier to study the sources of trouble

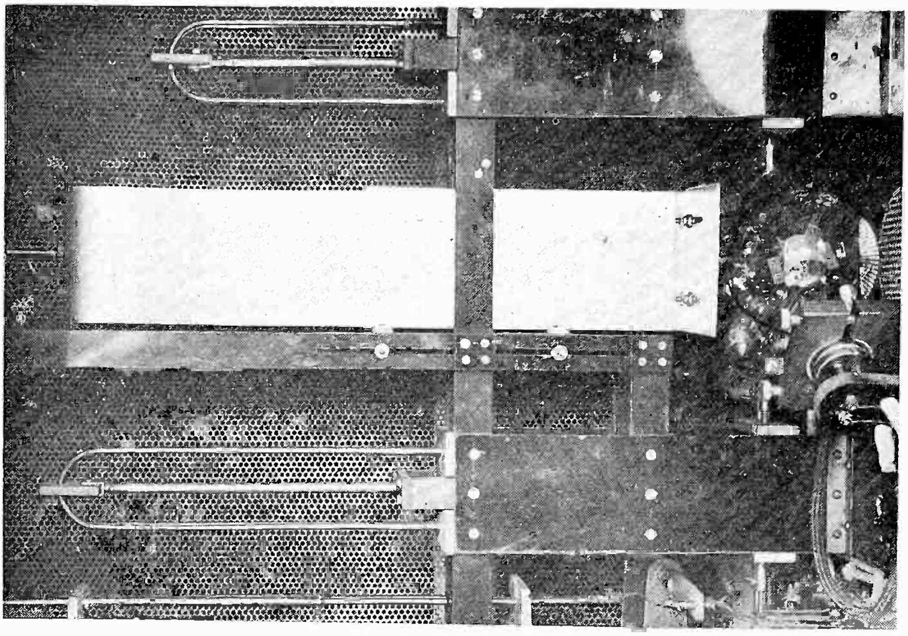


Fig. 12—Side view of 462-megacycle, 6-watt output, UX-852 tube, push-pull transmitter of the triode pendulum oscillator type, showing "catalyst" circuit.

if a circuit is oscillating, it is sometimes valuable to use an auxiliary trigger or a "catalyst" circuit. The functions of such a circuit are manifold. The conductive losses of the oscillator circuit may be too high. There may be an excessive radiation resistance or some phase and voltage condition may need correction. A catalyst circuit has, therefore no definite form but must be made up to suit the suspected condition. A wide strip of copper, a half wave long, may be bent around in different ways and placed in various positions near the tubes or the associated circuits, Fig. 12. If, for instance, the trouble consists of too high resistance in the flexible tube leads, some of the external circuit tuning may be taken over by the catalyst circuit since it may be placed in such a position that it is capacitively coupled to the tube elements directly through the glass envelope without the aid of the leads. Phase conditions may sometimes be corrected by means of tuned loops. Most forms of catalyst circuits are applicable in reducing radiation. The location of shields at certain distances from the oscillator may also sometimes produce desired results.

Example of Transmitting Equipment

In order to obtain as complete design data as possible the experimental models have in some cases been developed to such a point that

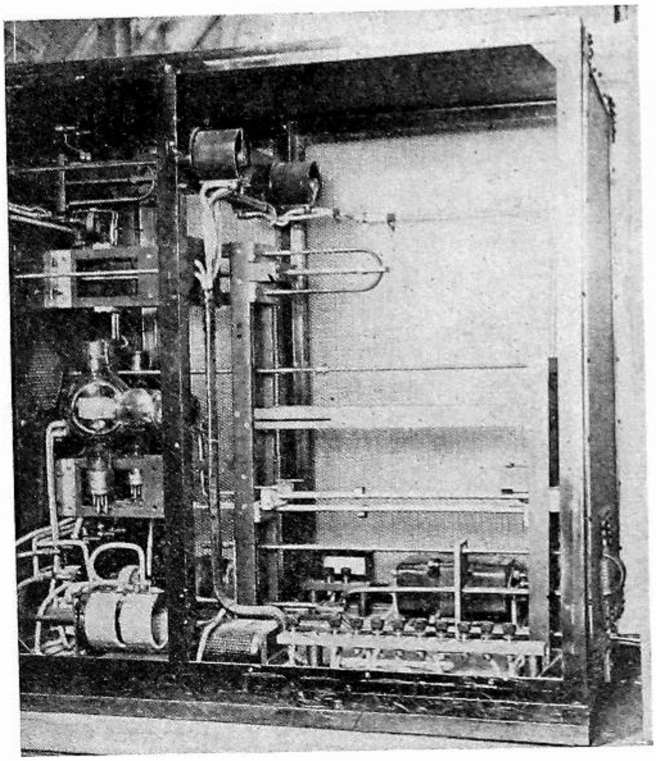


Fig. 13—Side view of 462-megacycle, 6-watt output, UX-852 tube, push-pull transmitter of the triode pendulum oscillator type, cathode circuit tuning is here being used.

they would, as nearly as possible, conform with the exacting requirements of commercial operation. Many of the considerations in such designs have already been discussed in a general way. Examples of such equipment need, therefore, only a brief description.

1. UX-852 Barkhausen Transmitters

The first two transmitter units were built to operate at a frequency of 462 megacycles and a wavelength of 65 centimeters. For about a year they were used for two-way telephony between Rocky Point and Riverhead. These transmitters, which are shown in Figs. 11, 12, and

13, are of the triode electron-pendulum type. Each transmitter consists of a pair of UX-852 tubes, operated in push-pull. The tuning circuits are of the linear conductor type and of the trombone variety. In one transmitter only the plate and the grid circuits were tuned. The filament heating current was supplied through choke coils. The oscillator was equipped with a catalyst circuit, Fig. 12. This circuit consisted of sections of three-inch wide metal strips forming a U of variable length similar to a trombone circuit. At the end of this U were flanges, each facing one of the oscillator tubes. As an alternative the other transmitter was equipped with a tuned cathode circuit, Fig. 13. On

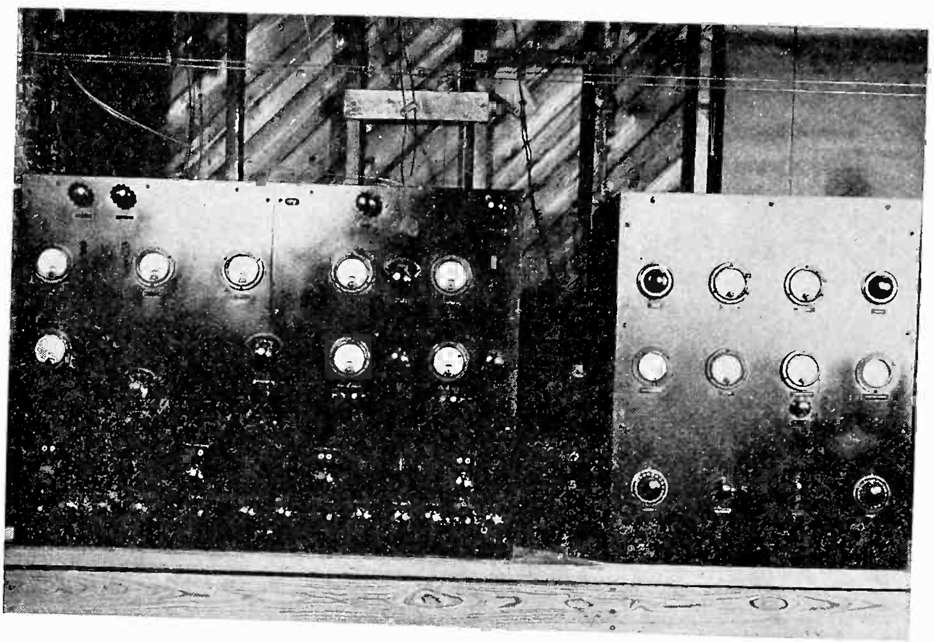


Fig. 14—Left: Front view of 432-megacycle, 15-watt output transmitter, consisting of a 144-megacycle oscillator, a buffer amplifier, a frequency tripler, and an amplitude modulator of the absorber type. All stages are of the push-pull type, using UX-852 tubes.
Right: Front view of plate voltage regulator for the plate supply of the oscillator stage and the buffer stage.

account of the heating current leads inside, the filament tuning conductors cannot conveniently be of the trombone type but must instead be tuned by a sliding connector. It made little difference which circuit was coupled to the load. The power was taken from the plate circuit in this case. Potentiometers for individual control of the direct-current electrode potentials of the two tubes were also introduced. All tuning, coupling, and potential regulation was performed from the front of the panel. Radio-frequency ammeters were inserted at points neutral to radio-frequency voltages. It was due to the desirability of keeping these meters in a fixed position and on the front of the panel, that the circuits had to be looped over and be tuned by trombones.

This transmitter was compound modulated. A special modulator, shown in Fig. 23, was built for this purpose.

The positive grid voltage required at 462 megacycles was 500 volts. The negative plate voltage was 125 volts. The normal filament voltage is ten volts for the UX-852 tubes. In order to obtain suitable emission of 250 milliamperes per tube it was necessary to cut the filament voltage down to between seven and nine volts. Due to the great heating gradient from the grid, the filament emission would at times become unstable. Some tubes would eventually become stable at about nine volts. This phenomena was credited to the sensitivity of the thoriated

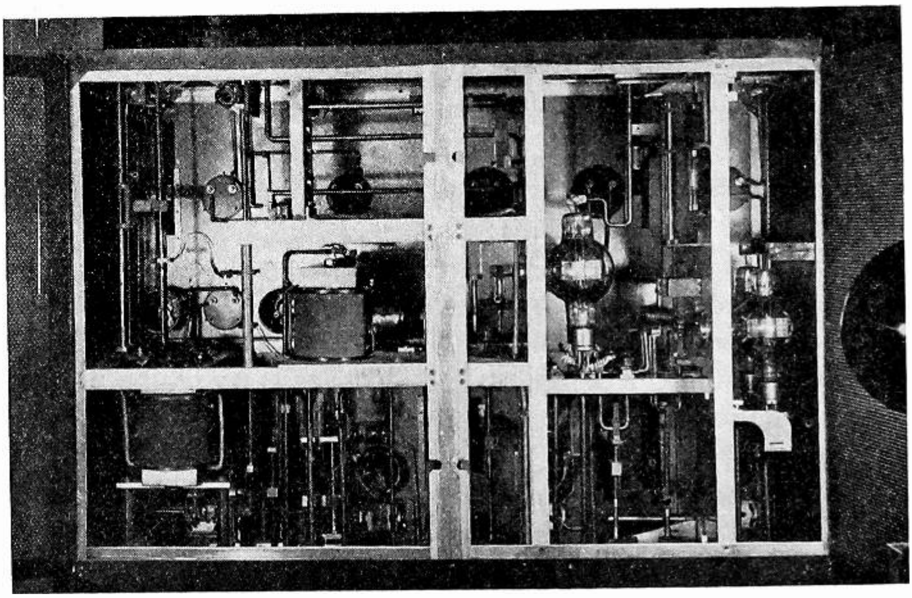


Fig. 15—Rear view of the 432-megacycle, 15-watt output, multiplier type transmitter shown on the left in Fig. 14.

filaments used in this type of tube. A power output of six watts was obtained from these transmitters.

2. UX-852 Frequency Multiplier

As a result of the experiments with frequency multipliers, a transmitter as shown in Figs. 14, 15, and 16 was built. It is of the push-pull type throughout. The first stage may either self-oscillate or be driven at about 144 megacycles. The output from this circuit drives an amplifier, or buffer stage, which gives a power amplification of two to one. This is the best amplification that could be obtained with the UX-852 tubes at this frequency. The power from this buffer is fed into a tripler circuit which delivers 432 megacycles output at about ten per cent efficiency. The tripler is equipped with means for producing an axial

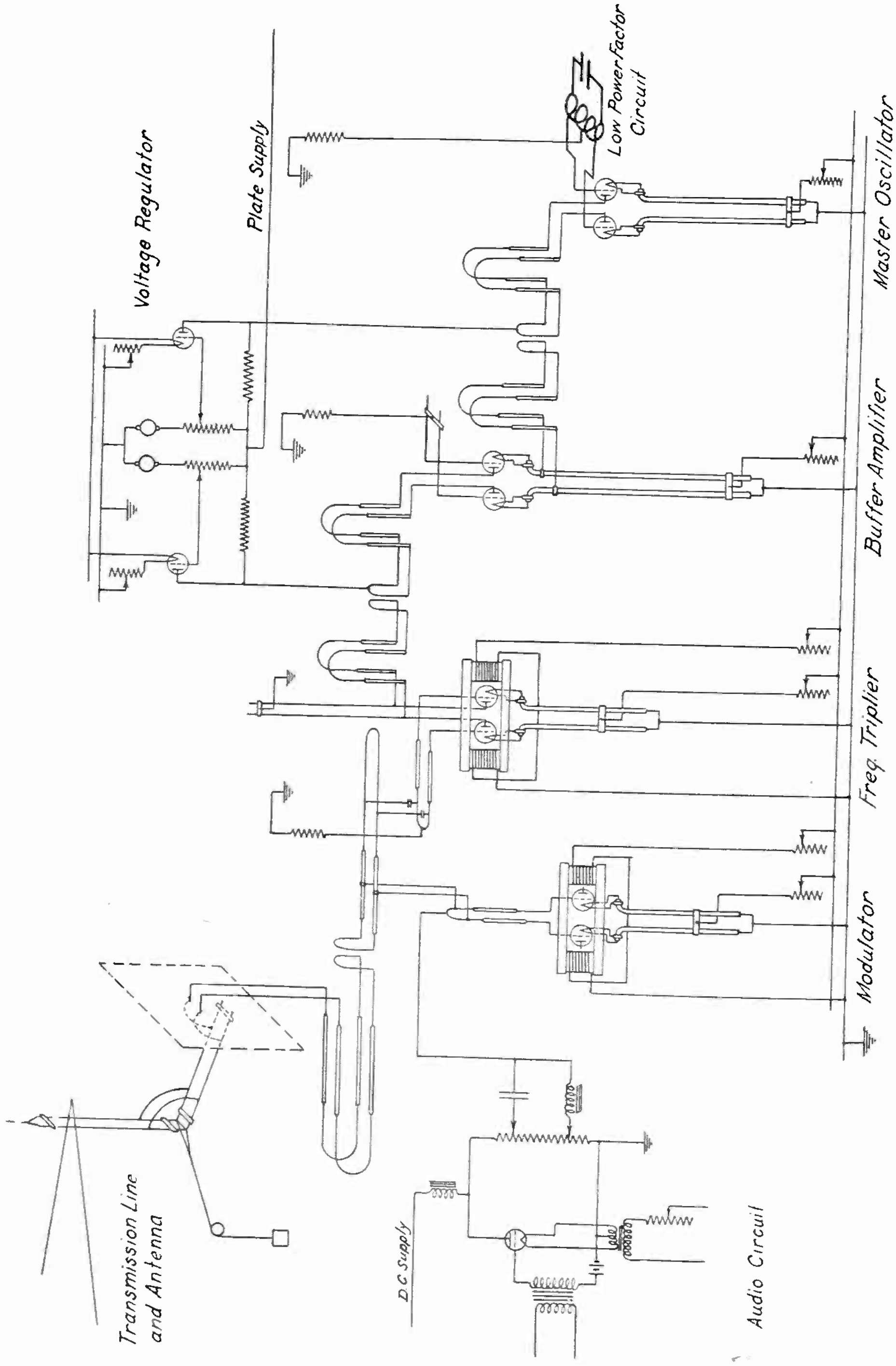


Fig. 16—Schematic wiring diagram of the 432-megacycle, 15-watt output, multiplier type transmitter shown in Figs. 14 and 15.

magnetic field of about 150 gauss. The power output obtainable from this transmitter is fifteen watts. All the tube elements are tuned and the tuning circuits are of the linear conductor type and of both the trombone and slider variety, depending upon the need.

One of the reasons for studying frequency multipliers has been the need of greater frequency stability and the particular adaptability of this method to frequency control. Crystal control was one of the first methods to be tried. An auxiliary control circuit was built which consisted of a crystal oscillator operating at about two megacycles followed by six stages of frequency multipliers. The crystal circuit and the first four stages were single sided circuits employing UX-210 tubes. The last of these stages fed into the combination of a single UX-860 buffer and a single UX-852 doubler driving a push-pull UX-852 tripler circuit, raising the frequency from 48 to 144 megacycles. It was found beneficial, even in this frequency region, to introduce an axial magnetic field. The field strength required was less than 100 gauss. The output from this auxiliary frequency control circuit was then used to drive the first 144-megacycle stage of the transmitter.

For modulation, a push-pull UX-852 absorber was used which was connected across the antenna transmission line as shown in Fig. 16. The length of the connecting leads to the absorber were such that optimum modulation effect would be obtained. These leads were connected to the grids of the absorber which act as modulation terminals. The plates may be left floating since they have no part in the circuit. The grid and the filament circuits are tuned. The grids are given a positive bias of about sixty volts, which causes them to draw a current of about five milliamperes each at an axial magnetic field of 200 gauss. The modulation voltage as well as the radio-frequency carrier to be modulated were then superimposed upon this grid condition. In this way it was possible to obtain linear amplitude modulation of about 35 per cent.

Numerous arrangements for replacing the complicated crystal drive, with its many multiplier circuits, have been tried. These arrangements consisted of various forms of low power factor circuits connected to the grids of the 144-megacycle oscillator. Some of these arrangements have proved very successful in that stability approaching that of crystals was obtained. Data on such circuits have already been published and further information may be published later.

3. *RCA-846 Frequency Multiplier Transmitter*

Although the design of a semicommercial model of this apparatus has not been finished, at the time of writing this paper, it appears to have sufficient points of interest to warrant being included among the transmitter examples in its experimental form, Figs. 17 and 18. It consists of a push-pull oscillator operating at 137 megacycles furnishing power to a push-pull frequency tripler. The tripler is provided with an axial magnetic field. All the tubes are RCA-846 water-cooled tubes. The tuning circuits are of the linear conductor type and arranged similarly to the circuits for the UX-852 multiplier. The tubular plate tuning conductors also carry the water supply to the plate jackets.

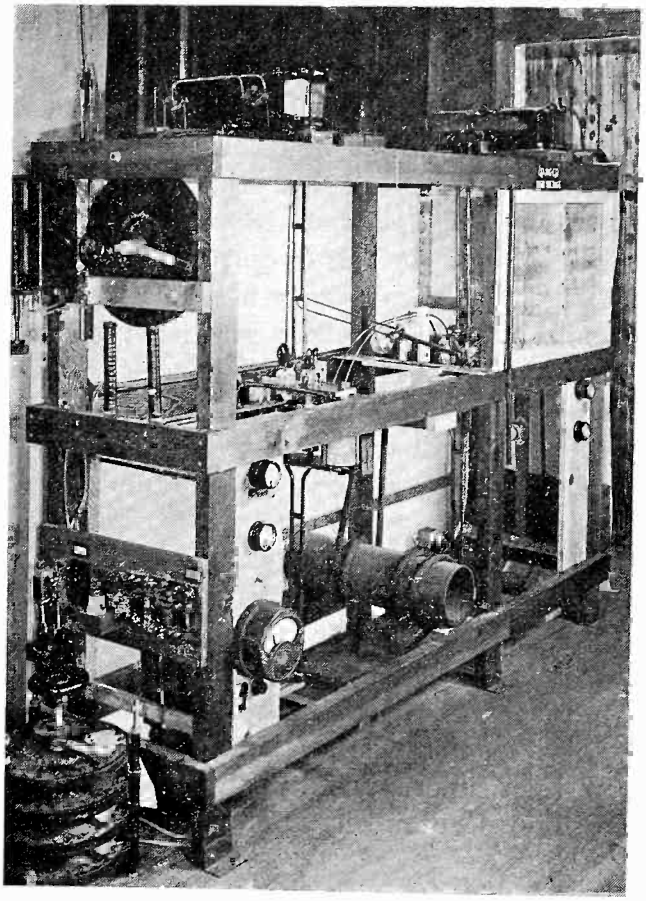


Fig. 17—Frequency multiplier transmitter, consisting of a 137-megacycle, 1500-watt output, triode feed-back oscillator (in the background) and a tripler, with organizing magnetic field, giving an output of 115 watts at 411 megacycles. Both stages use RCA-846, water-cooled tubes in push-pull formation. Note lamp loaded wavemeter (same as in Fig. 7).

Several optimum circuit combinations, each requiring a different magnetic field, could be otained. With an input to the tripler of 1200 watts the best output obtained when using a magnetic field was 115 watts at an output frequency of 411 megacycles. The magnetic field required for various circuit arrangements varied between 100 and 300 gauss. When using no magnetic field the best circuit arrangement would give an output of 35 watts.

4. Triode Feed-Back Circuits

Great progress has been made recently in the development of triodes for operation, according to the conventional feed-back method, at very high frequencies. Transmitting tubes have been developed which will give good output up to as high frequency limits as 300 megacycles. (ZP-94 curve in Fig. 1.) It is therefore no longer necessary to resort to special methods for frequencies of this magnitude. These developments have, however, also resulted in a proportional increase in the frequency borders of the frequency multiplier method and it is

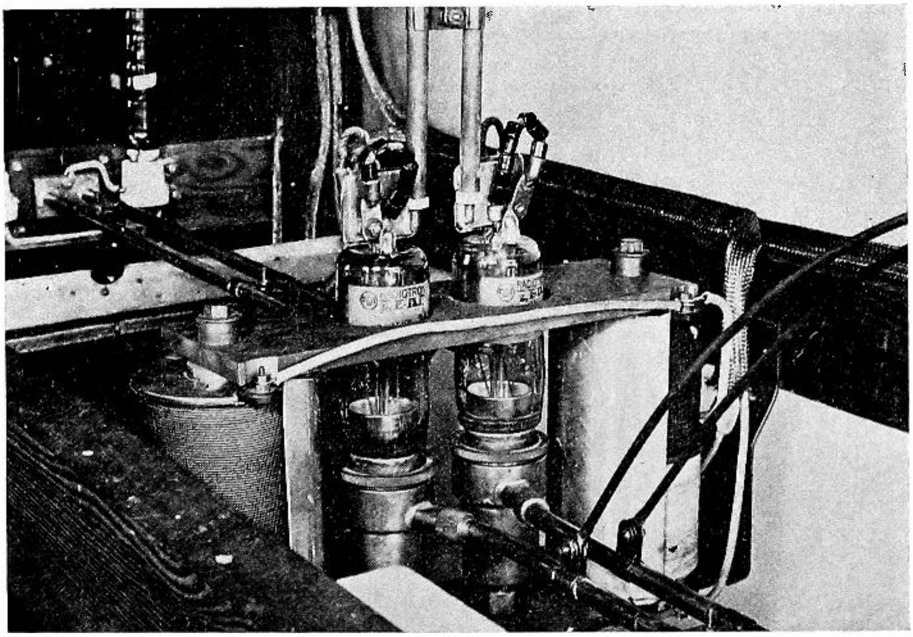


Fig. 18—411-megacycle, 115-watt output, RCA-846 water-cooled tube, push-pull tripler with parallel coil system magnet for producing organizing magnetic field.

now possible to consider this very stable method in the 1000-megacycle region.

Antennas and Transmission Lines

In designing directional antenna systems, choice can be made between two general principles. The first one, which may be called the Hertzian method, consists in redirecting the radiated energy from a single nondirective radiator into one direction. This is done by means of a parabolic, metallic reflecting surface or by means of conductors arranged in a parabolic array. The perfection of the method depends upon the size of the reflector system, measured in wavelengths.

In the second method no reflecting surfaces need be employed. The origin of radiation is instead distributed. A number of radiating elements, such as dipole antennas may be so spaced and phased that they add up very efficiently only in diametrically opposite directions. By further combination, radiation in one of these directions may also be eliminated.

Several years ago the method of combining straight harmonic radiators instead of dipoles was introduced by RCA Communications, Inc.[4] This principle has greatly simplified the details of directive antenna design. Detailed descriptions of such systems have been published previously in the PROCEEDINGS.[5] As the name indicates, the harmonic wire antennas are based on the standing wave principle.

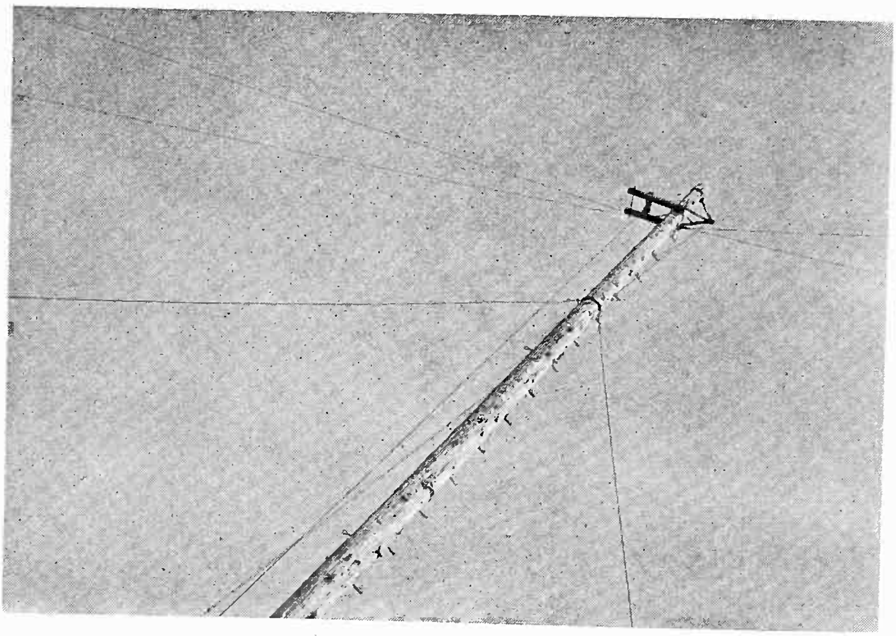


Fig. 19—Feed end of the V antennas.

When the frequency is sufficiently high so that it becomes possible to make the length of the radiators on the order of 50 to 100 wavelengths, the radiation attenuation becomes so great that very little energy reaches the far end of the wires. The antenna then becomes unidirectional and aperiodic. This is a great simplification since it makes reflector systems and tuning arrangements unnecessary. An antenna of this type is shown in Fig. 19. It is commonly known as a V antenna. A V antenna 100 wavelengths long should have an angle of 9.8 degrees between the wires. The power gain, over a doublet, is more than 100.

It may at times be desirable to provide an antenna entirely free

[4] Lindenblad, U. S. Patents No. 1,884,006 and No. 1,927,522.

[5] P. S. Carter, C. W. Hansell, and N. E. Lindenblad, "Development of directive transmitting antennas by R.C.A. Communications, Inc.," PROC. I.R.E., vol. 19, pp. 1773–1843; October, (1931).

from back radiation. Such an occasion may arise when the transmitting and receiving antennas have to be closely located and operate on near-by frequencies. An antenna having a metallic sheet reflector is then desirable since it is more perfectly free from back radiation than other antenna types. In order to avoid curved reflector surfaces, antennas as shown in Fig. 20 were designed. This antenna consists of rectangularly bent wires with half-wave distances between the bends. By combining several such wires an arrangement is obtained in which all the vertical elements are free to radiate and are of the same phase, whereas all the horizontal ones have their radiation cancelled by a

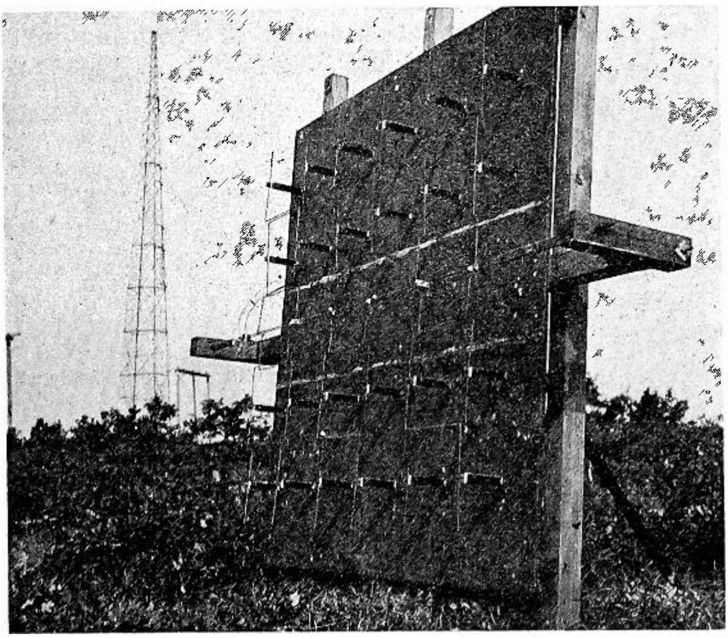


Fig. 20—"Billboard" type, directive antenna.

horizontal member of another wire in juxtaposition. The various bent wires are attached to a common feeder line in pairs, at half-wave intervals. The wires are supported by means of insulators or metal columns at the voltage nodal points. The whole system is mounted in front of either a metal screen or a solid metal sheet at a distance of about a quarter of a wave or an odd multiple thereof.

Many workers have followed the practice of building the transmitter and antenna system together as a single unit. This is undesirable in many cases because the antenna must be at a great height to obtain great range. A transmitter located at the antenna cannot be serviced conveniently. Consequently, it is often desirable to locate the antenna and transmitter some distance apart, requiring the use of transmission lines of considerable length. On most occasions this has been the practice. A study of the characteristics of such lines had therefore to be made. Balanced lines using No. 4 B&S and No. 6 B&S wire were investigated. Inductive coupling to the line under test was employed to minimize the effect of reflections and unbalance in the connecting lines from the transmitter. The measurements on the lines were made with the aid of a sensitive thermogalvanometer which made it possible to use very small capacitive coupling to the wire system, thus avoiding disturbance of the natural capacity and resistance of the line. The relatively large amount of power (100 watts) available from the frequency multiplier transmitter was an important factor in making it possible to use very light coupling between the measuring instrument and the wires. The measurements included determining losses and also velocity variations due to moisture on the wires. The loss measurements were carried out at a frequency of 405 megacycles and the velocity measurements at 422 megacycles.

For No. 4 B&S wires spaced one inch apart the calculated value of the attenuation constant per 100 feet is 0.037. The measured average is 0.045. This corresponds to an efficiency of 91.4 per cent for a 100-foot line and 40.7 per cent for a 1000-foot line. The theoretical value of the attenuation constant for a No. 6 B&S wire is 0.042 per 100 feet and the measured value is 0.050. This corresponds to an efficiency of 90.5 per cent for a 100-foot line and 36.8 per cent for a 1000-foot line. It was found that the difference between losses on clean and weathered wires was too slight to be measured. These measurements indicate that it is practical to use balanced, closely spaced, two-wire transmission lines up to several hundred feet in length. For longer lines, the more expensive large diameter concentric conductor type of line must be chosen if good efficiency is to be maintained.

The presence of spacers, if made of good material and if used sparingly, usually add very little to the loss in dry weather. In wet weather, however, the loss and the reflections caused by such spacers are very serious handicaps. They are avoided entirely by using U-shaped half-wave conductor suspension loops, or "metallic insulators."

During rain the increase in loss from the presence of a water film on the wire was found negligible. The wave velocity along the line did, on the other hand, show a very marked decrease. Weathered and polished wires also showed a very marked difference in their ability to hold water. A horizontal line made of new No. 6 B&S wires with one-inch spacing was found to hold enough water to reduce the velocity by three per cent. The distribution of the water was very beady. The thickness of an evenly distributed water film required to cause this amount of variation in velocity was found by calculation to be 0.008 inch. The distribution of the water on a weathered wire was much less beady and was more evenly distributed. The velocity change for maximum water condition was 0.6 per cent, corresponding to a uniform film thickness of 0.002 inch. The maximum reduction in velocity for a weathered No. 4 B&S wire was 0.5 per cent.

From these values, the velocity change on a more open-wire system such as a V antenna can be calculated. For such a system the velocity change for weathered wires amounts to about 0.2 per cent or less. This is fortunate since the change will ordinarily not have sufficient influence upon the characteristics of a hundred-wave antenna to be serious. Such variations and even greater ones can, however, be expected during sleet storms. In some commercial installations it will be necessary

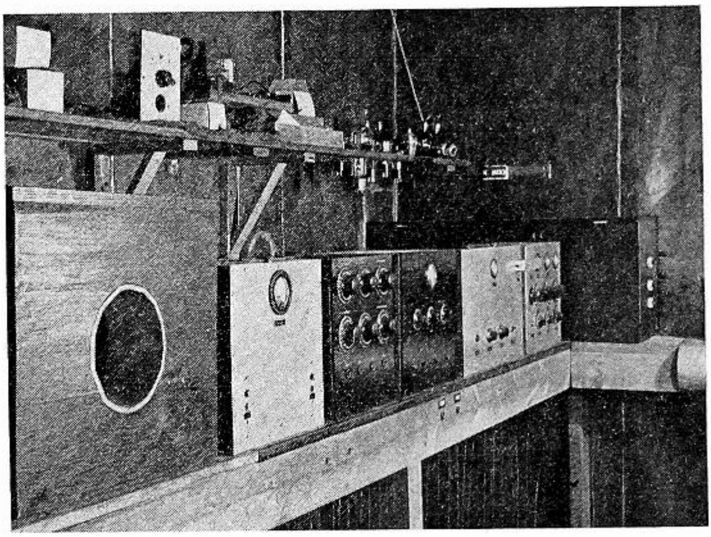


Fig. 21—Rocky Point receiving equipment for 200- to 500-megacycle signals.

to provide for sleet melting. There are a number of ways in which this can be done with long wire antennas.

Propagation Tests

The chief purpose of the transmitter equipment which has been developed in accordance with the principles just outlined has been to study propagation in new frequency regions. Since the results of some of these tests are quantitatively described in detail in a paper by B. Trevor and R. George,[6] only some of the results obtained will be given in this paper.

After preliminary local tests, communication between Rocky Point and Riverhead, a distance of fourteen miles, was established. The 6-watt, 462-megacycle Barkhausen transmitter shown in Figs. 11, 12, and 13

[6] B. Trevor and R. George, "Notes on propagation at a wavelength of seventy-three centimeters," Proc. I.R.E., vol. 23, pp. 461–470; May, (1935).

was used. This equipment was duplicated so that two-way telephony could be carried on. The Rocky Point receiving equipment is shown in Figs. 21 and 22. Both transmitting and receiving antennas, which were of the billboard type shown in Fig. 20, were mounted on supports fifty to seventy-five feet high in order to be in line of sight. As is usual when there is line of sight between transmitting and receiving antennas,

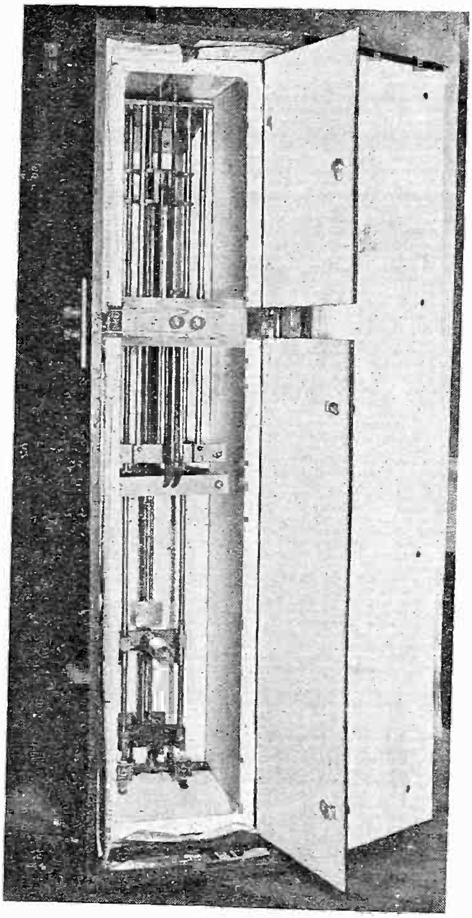


Fig. 22—Triode pendulum detector with trombone tuning circuits. Detail of unit in the background of Fig. 21.

no fading was observed. Rain, snow, and dense clouds of smoke from forest fires seemed to have no effect on the signals. The absence of fading on this circuit was a desirable feature since its purpose was to test the stability of transmitters and receivers. The equipment was used later for propagation studies under more difficult space-circuit conditions.

The Barkhausen transmitter, at Rocky Point, Figs. 11, 12, and 13, was eventually replaced by a frequency multiplier unit, Figs. 14, 15, and 16, capable of delivering up to 15 watts at 432 megacycles. The

Rocky Point-Barkhausen and low power multiplier set-up is shown in Fig. 23. Two horizontal V antennas were also erected. They were both seventy feet above the ground. One was 70 waves long and directed eastward on Riverhead. The other was 100 waves long and directed westward on the Empire State Building in New York City, fifty-six miles away. Signals from these antennas were observed with a portable receiver and a small directive antenna mounted on top of a car. Signals without fading were received at distances up to thirty miles, in many localities considerably below the line of sight. By locating the

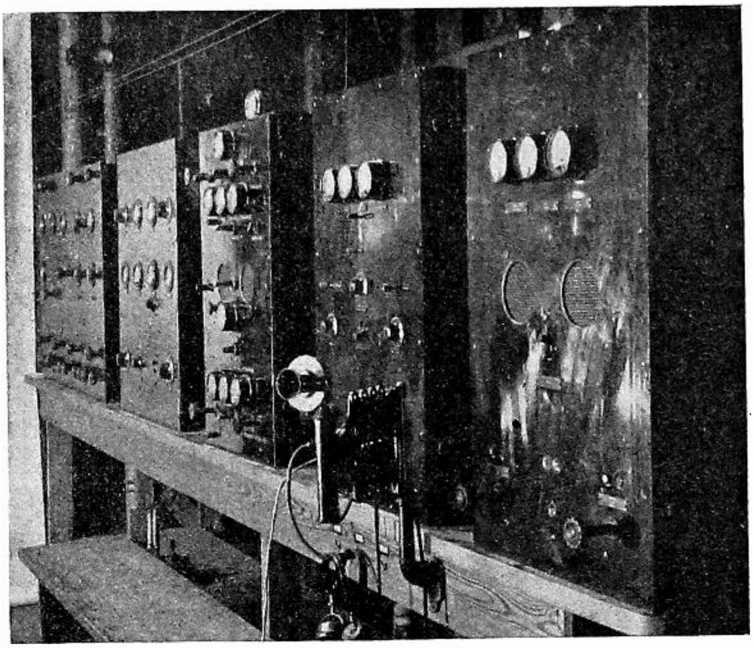


Fig. 23—Rocky Point low power ultra-high-frequency transmitter set-up. From left to right: 15-watt, 432-megacycle frequency multiplier; voltage regulator; 6-watt, 462-megacycle transmitter; compound modulator; and rectifier for modulator.

receiving system on top of high buildings in New York and in airplanes it was found that signals without fading could be received at distances of from fifty to sixty miles when received as much as 500 feet below the line of sight. The distances below the line of sight given here represent actual values taken from contour maps of the terrain between transmitter and receiver. They will, therefore, check only approximately with simple geometric considerations based on the average curvature of the earth.

During these cruises with car and airplane, the unusual fact was observed that the apparent width of the beam was considerably smaller than the calculated width. For example, the beam from the 100-wave V antenna, which should have been about five miles wide over New York, seemed to be only two miles wide at the most.

After the 100-watt frequency multiplier transmitter became available it became possible to increase the range of observation. The frequency used was 411 megacycles. For the approximate distance of sixty miles, signals at ground level, 2000–3000 feet below the line of sight, would fade when completing their last lap over a very variable land and water route. When the receiving end of the route was over an uninterrupted body of water, there was very little fading at the time of observation. At a point 113 miles away, with only a relatively small stretch of water midway, only violently fading signals could be observed, even during prolonged observation periods. These observations were made at an elevation of about 400–500 feet above sea level and 8000 feet below the line of sight. It was also determined, that at least for this particular space circuit, horizontally polarized waves were considerably superior to vertically polarized waves. The fact was also observed that the plane of polarization at the receiver was at all times identical with that at the transmitter.

With the transmitting antenna 70 feet above the surrounding country and 170 feet above sea level, badly fading signals were observed in an airplane, 165 miles away, flying at a height of 7500 feet. In a later flight, after the height of the transmitting antenna had been increased to 120 feet above the surrounding country and after the receiver and the ignition system of the plane had been very substantially improved, surprisingly little distance was added to that obtained in the previous flight. Badly fading signals could still be observed at a distance of 172 miles when flying at an altitude of 7500 feet and two miles below the line of sight. The routes of these signals were chiefly over land.

It appears from these tests that the distance of reliable as well as unreliable signals increases with the increase of power at the transmitter. It is unsafe to predict whether this proportionality will remain when still higher powers are applied. The distance beyond line of sight, for reliable signals, will, of course, also depend upon the nature of the route.

If improved facilities result in much greater distances, no doubt the electrical conditions of the portions of space more remote from the earth's surface will play an important part in determining the propagation characteristics. The indications are that, at the present, most of the phenomena observed are obtained in the region of space near the earth's surface. No doubt such phenomena as nonuniform humidity and the surface condition of water bodies have a great influence upon the characteristics of the propagation along the earth's surface. The shape and nature of the landscape, nedless to say, has an enormous influence.

In Table I, a chart of propagation results, obtained at different

TABLE I

EXAMPLES OF ULTRA SHORT-WAVE TRANSMISSION

Frequency Megacycles	Wavelength Centimeters	Distance		Elevation Meters		Value of *m*	Terminals of Test Circuit	Investigators
		km	miles	*a*	*b*			
60	500	205	128	700	530	3.87	Nice-Corsica	Jouaust, Ferriè Milan
40	750	456	284	1462	0	1.1	Hawaii-Kauai	Beverage, Hansell
40	750	145	90	518	0	1.446	Kauai-Oahu	Peterson, Fifield, Matthews
44 61	682 492	446	278	1916	387	1.32	Empire State Building—Mt. Washington	Jones
527	57	270	168	750	340	1.58	Rocca di Papa—Cape Figari	Marconi, Mathieu
527	57	85	53	750	0	−3.0	Rocca di Papa—Yacht Elettra	"
430	70	90	56	304	51	2	Rocky Point—Empire State Building	Lindenblad, Dow, George, Trevor
411	73	96.5	60	52	18	1.22	Rocky Point—Montauk Point	"
406	74	180	112	70	116	1.16	Rocky Point—Arney's Mount	"
406	74	105	65	70	65	1.43	Rocky Point—Atlantic Highlands	"
406	74	274	170	70	2287	1.88	Rocky Point—Airplane	"

frequencies by various investigators, is shown. In comparing these results the method referred to by Jouaust[7] has been used. A factor *m* is used by which the radius of curvature of the earth must be multiplied to give the curvature of the propagation path. Thus if *m* assumes its smallest value, that of unity, the curvature of the propagation path will be equal to the curvature of the earth. According to Humphrey the value of *m* for light is 5.7. Others have given it values as high as 10. The results given in the table may be judged by the smallness of the factor *m*. It can be seen that for frequencies above 30 megacycles, the Hawaii-Kauai telephone circuit has a very highly curved propagation path. At frequencies above 300 megacycles, very high curving was obtained on the test circuit between Rocky Point and Arney's Mount.

ACKNOWLEDGMENT

The research and development outlined in this paper naturally represent the combined efforts of many individuals. Messrs. O. E. Dow and Bernard Salzberg have contributed much to the transmitter development. Messrs. R. W. George and Bertram Trevor have developed the receiving equipment and made the propagation observations. Mr. E. E. Spitzer and associates at the RCA Radiotron Company and Mr. P. S. Carter of RCA Communications, Inc., have rendered very valuable assistance. The author wishes to express his thanks to Messrs. H. H. Beverage, C. W. Hansell, and H. O. Peterson for their many valuable contributions and for encouragement given.

[7] R. Jouaust, "Some details relative to propagation of very short waves," PROC. I.R.E., vol. 19, pp. 479–489; March, (1931).

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THE GRID-COUPLED DYNATRON*

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Summary—*This paper deals with an improved oscillatory system wherein the dynamic excursion is associated with the secondary emission region of the vacuum tube characteristic, plate current vs. plate voltage. A more or less conventional dynatron oscillator, using a tetrode, is improved by feeding back some of the output power to the inner grid. The maximum increase in oscillatory power was found to be about fifty per cent above the same system with a static control grid potential.*

Introduction

THE limitations of the dynatron and the pliodynatron oscillatory systems are readily discernible when one associates a given electronic device with some R, L, and C circuit in an attempt to increase the amplitude of oscillation by reducing the internal resistance R of the tuned circuit. It can be shown experimentally and graphically that a certain region exists over which the amplitude is materially increased by a reduction in R, however, as larger and larger amplitudes are realized and the peak voltage (E_p+e_z), Fig. 1, approaches the second point on the $i_p=f\ (E_p+e_z)$ characteristic, where one secondary electron is realized on the average for every primary impinging electron, the system exhibits a diminishing return for any effort to improve the circuit by increasing its external impedance. When this region is approached or crossed by the peak amplitude it will be found that the circuit drives the tube for a portion of the cycle in order to produce energy equilibrium. The circuit to be described is one wherein a deformation of the dynamic characteristic is obtained to alter the power relations between the tube and circuit, thus affording a change in the amplitude of oscillation for some fixed circuit and tube conditions.

The Dynatron

Fig. 1 shows diagrammatically a four-element tube arranged with an oscillatory circuit. The electrical arrangement of the electronic device shows that it should operate on the dynatron principle due to Hull. A three-element electronic device is equivalent for the connections shown since the cathode and the inner grid of the electronic device, Fig. 1, can be considered as some equivalent cathode for three-element tube considerations. In this paper, Fig. 1 and curves associated with that

* Decimal classification: R355.9. Original manuscript received by the Institute, July 3, 1934.

figure will be considered to apply to both the three- and four-element electronic device if the inner grid is disposed as shown or in combination with some fixed bias potential.

Let us choose an operating point, Fig. 1, on the plate-current—plate-voltage characteristic somewhere near the first point where, on the average, one secondary electron is collected for every primary impinging electron. If the slope of the curve at this point is such that the value of (dE_p/di_p) is equal to, or less than, the absolute value L/R_LC of the tuned circuit, the system will start to oscillate and will come to

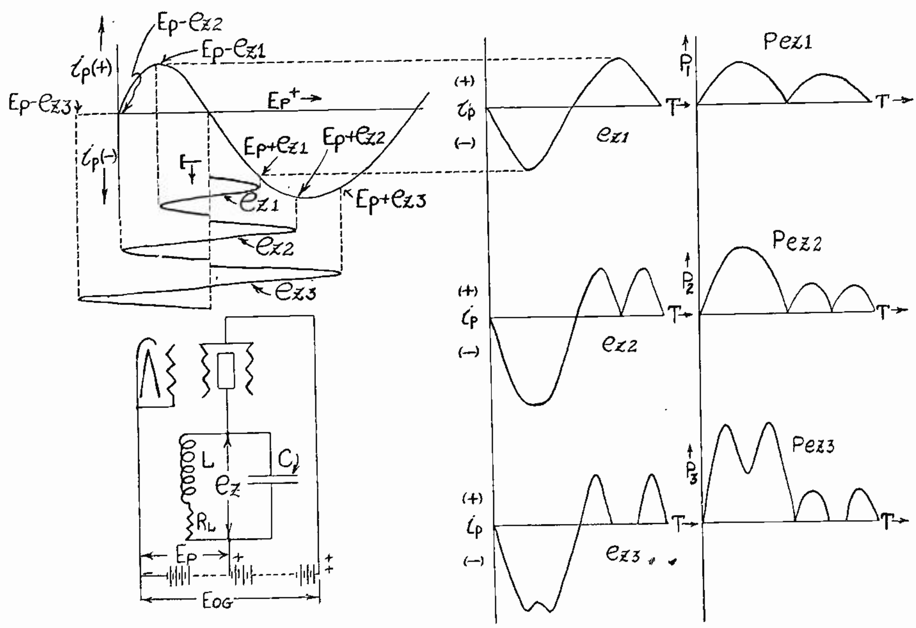

Fig. 1

stable equilibrium at some voltage amplitude e_z, for example, which appears across the tuned circuit, L, R_L, C. The existence of the time varying voltage e_z causes the instantaneous plate voltage to vary between the peak limits (E_p+e_z) and (E_p-e_z) at an angular velocity very nearly equal to $1/\sqrt{LC}$.

Since the excursion of the voltage e_z determines the interchange of current between the tube and the tuned circuit, a qualitative plot of the current to the tuned circuit will show wherein an increase in the amplitude e_z is not met by equal increases in the available energy per cycle from the tube; and thus higher amplitudes are only obtainable by shortening the life of the tube by the application of excess voltages or by using oscillatory circuits that are not ordinarily realizable.

The shapes of the anode current curves and the corresponding oscillatory power curves are shown in Fig. 1 for three amplitudes $e_{z_1} < e_{z_2} < e_{z_3}$. It is permissible to assume a sinusoidal amplitude for e_z in computing the shape of the current and consequently the power curves because with a fair oscillatory circuit the harmonic voltage content is small and the harmonic power contributions are quite negligible. It is evident by inspection that as the amplitude e_z increases from e_{z_1} to e_{z_2}, to e_{z_3} etc., the time per cycle is decreasing within which plate current flows from the plate or anode battery to the tube and circuit, (i.e., plus values of i_p). It is also evident that some peak values of e_z corresponding to an instantaneous anode voltage $(E_p - e_z)$ occur at the "free-wheeling" portion of the cycle and do not enter as products of a current i_p to contribute energy to the oscillatory circuit. Referring to the other half of the cycle where peak values of e_z make the instantaneous plate voltages $(E_p + e_z)$ it is evident, as e_z increases, that the peak values of e_z occur at low values of plate current; and if the excursion is such that the positive region of plate current is entered the oscillatory circuit helps drive the tube and an additional sink is produced for the entire system.

Methods of Increasing the Amplitude

It can be concluded from the simple system of equations for the oscillatory system shown in Fig. 1, and by physical reasoning, that the amplitude of oscillation can be increased in several ways; by a reduction in an effective negative resistance $(-d\ E_p/d\ i_p)$ at the expense of the cathode and supply batteries, by a reduction in the effective resistance of the tuned circuit in order that the circuit might drive the tube for a portion of the cycle, or by a deformation of the dynamic characteristic in such a manner that instead of following a single curve for one set of element potentials, it will sweep across a family of characteristics so as to contribute more or less available energy per cycle to the tuned circuit than is expended by the tuned circuit in some control electrode circuit to bring about this change in the dynamic characteristic.

There is a limit to which the resistance R_L or any contribution to loss in the condenser C, Fig. 1, can be reasonably reduced unless mechanical resonating systems are resorted to. It is also expedient not to overload the electronic device in order that it might have a reasonably long life. Thus, considering some fixed element potentials, some deformation of the dynamic characteristic must be realized by virtue of an oscillatory state in order that an increase in the available energy per cycle from the tube to the oscillatory circuit is to be realized.

Hypothesis of the Improved Circuit

In order to accomplish the previously mentioned dynamic characteristic deformation a dynamic change in some inner or outer grid potential must be developed, in a desired relation with the oscillatory current and favorably related in phase, so as to produce the desired effect. This dynamic change in some control element potential can be produced by any form of mutual coupling of the control element with the oscillatory circuit.

Since the energy associated with the tuned circuit gives rise to the voltage e_s across the circuit, any additional load upon this circuit is equivalent to an effective increase in R_L. On the other hand, even though the effective increase in R_L is inevitable by virtue of some coupled loss from some control electrode, the energy expended in the control element circuit might be offset by a more substantial increase in the available energy per cycle and thus result in a net gain in amplitude. Inasmuch as the available number of primary electrons is limited and the energy dissipated in the internal dynamic resistance of some control electrode might be large for large voltages it would seem that as the introduced voltage due to the oscillatory current was increased, the increase or decrease in oscillatory current would pass through a maximum or minimum point or exhibit a best value for each set of circuit and tube conditions.

Inner and Outer Grid Considerations

Confining the discussion to increases in amplitude with a four-element electronic device one might survey the possibilities of producing a change in the inner or outer grid potential. The outer grid, at least for the tubes available to date, operates as a collector at a rather high direct potential and draws considerable current. It is quite necessary to make an appreciable change in the outer grid potential to produce a substantial change in the tube characteristic. On the other hand, the inner grid can be operated at either a negative, zero, or positive potential with respect to the cathode and it requires relatively small changes in potential to produce an appreciable change in the number of primary electrons leaving the vicinity of the cathode. Since the inner grid exerts such a marked control over the space charge, and the negative resistance $(-dE_p/di_p)$ is inversely proportional to the number of primary electrons, a properly phased control of the inner grid potential will result in a dynamic change in the excursion in as much as the inner grid[1] determines the number of primary electrons which impinge upon the plate.

[1] The effective projected area of the outer grid is not considered.

The Improved Circuit

Fig. 2 depicts one form of the new electrical arrangement that substantiates the foregoing hypothesis. In this circuit the inner grid potential is not only a function of the direct bias potential but also the oscillatory current in the tuned circuit. As drawn, the circuit shows an inductive coupling between the oscillatory circuit and a coil associated with the inner grid. For increases in oscillatory current the mutual can be arranged so that as the point (E_p+e_z), Fig. 2, is approached dynamically, the inner grid potential increases positively so as to release more primary electrons from the cathode area, thus producing an in-

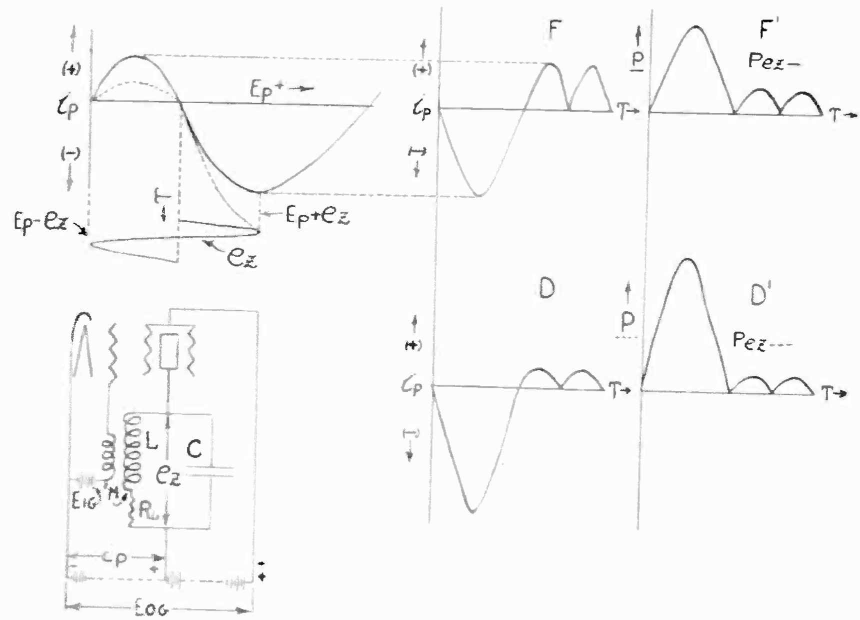

Fig. 2

crease in secondary emission. As the excursion of (e_z) approaches the point corresponding to (E_p-e_z) there is a reduction in the number of primary electrons and hence a decrease in the plate current. In the region denoted by instantaneous values of (E_p-e_z) the contribution of the instantaneous products $(e_z i_p)$ only occur for a small portion of the cycle and at low values of e_z and i_p. Hence, the loss of available energy per cycle from this portion of the cycle is negligible compared to the gain produced on the other half cycle. The net result for the specific operation point designated in Fig. 2 is an increase in the number of primary electrons impinging upon the plate and thus an over-all increase in negative current in the negative portion of the plate-current half cycle and a reduction in the positive region of plate current. The contribution of the positive half cycle of plate current is negligible to that of the negative half for large values of e_z and the net result is a deformation of the dynamic characteristic to produce a desired result. A dynamic characteristic for the circuit of Fig. 2 is shown with this figure by the dotted line, where the full line represents the dynamic characteristic without inner grid excitation. It is evident that an average of the instantaneous products of e_z and i_p integrated over the cycle for the same amplitude e_z will be greater in the case of the dotted dynamic characteristic than that for the full line. The corresponding tube current and power curves are represented by curves D and D' and F and F' appearing as part of Fig. 2.

The excitation of a control electrode by some mutual relation with the oscillatory circuit has been considered on the basis of an increase in the oscillatory amplitude. The converse is possible for an opposing connection to the control electrode, and the effects of this introduced voltage upon the inner grid will be mentioned as it relates to the experimental evidence and the conclusions.

Experimental Observations

The circuit of Fig. 2 was arranged as shown and a study of the oscillatory current in the L, R_L, C circuit was made as a function of mutual (M) and the element potentials. It was found that the circuit exhibited a best value of (M) for each set of tube conditions and the values observed were different for each value and polarity of the inner grid bias potential. The power increase was also found to be quite marked for some settings, and increases in oscillatory power from the tube to the circuit varied from fifteen to fifty per cent. A qualitative résumé of the experimental observations is given in the conclusions, it being deemed irrelevant to set forth quantitative measures for any particular electronic device where a considerable variation in the negative characteristic can be found in tubes of the same type.

Mathematical Relations

The mathematical treatment of the circuit, Fig. 2, is equivalent to those existing in the idealized arrangement where a negative resistance is placed across an L, R_L, and C circuit. It is, however, necessary to interpret the negative resistance for the dynamic characteristic of the tube as composed by a sweeping of the family of characteristics; and in addition the interpretation of R_L must be such as to include any coupled grid or control electrode losses where conditions are such that they might exist. Such an idealized treatment results in a two-mesh network, the determinant of which must vanish in the case of sustained oscillations, i.e.,

$$D = \begin{vmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{vmatrix} = 0.$$

Substitution of the various impedances corresponding to Z_{11}, Z_{12}, and Z_{22}, and a collection of terms results in the conditions for oscillation

$$\left| -R \right| = \frac{L}{R_L C} = \frac{\omega^2 L^2}{R_L} \text{ (collecting the reals)}$$

and the angular velocity

$$\omega = \frac{1}{\sqrt{LC}} \begin{array}{l}\text{approximately} \\ \text{(collecting the imaginary terms)}\end{array}$$

which is consistent with the idealized conditions and will furthermore be found in texts dealing with oscillatory circuits in combination with negative real parameters.

Conclusions

It is well to state that in addition to the effect of either increasing or decreasing the power output of the oscillatory system at any specific frequency the system offers other advantages. With the system shown in Fig. 2 no difficulty was encountered in producing constant voltage variable frequency oscillatory systems for both audio- and radio-frequency ranges. The latter was accomplished by combining the natural voltage-frequency characteristic of such an oscillatory system with compound mutual coupling from the oscillatory circuit to some load circuit. Other combinations such as simple or compound coupling from the oscillatory circuit to some control electrode simultaneously arranged with simple or compound mutual coupling from the oscillatory circuit to the load circuit produced substantially constant voltage-frequency characteristics when the condenser C was made variable.

The performance characteristics of the system, Fig. 2, for operating points in the near region as shown by this figure are summarized as follows for a '24 type tube:

(1) The maximum increase in oscillatory power varied somewhat critically with the value of coupling and a best value was found to exist for each condition of inner and outer grid bias, anode voltage, and circuit impedance.

(2) The maximum increase in oscillatory power was greater for the operating point shown in Fig. 2 when the inner grid bias was in the region of zero than when said bias was either positive or negative.

(3) The voltage introduced on the inner grid which produced the optimum oscillatory current increased with the plate potential.

(4) For increasing plus values of inner grid bias the smaller the coupled voltage necessary to produce maximum oscillatory current.

(5) Small values of oscillatory current produced by poor oscillatory circuit conditions were not increased to any marked degree by the method suggested.

(6) For all values of inner-grid bias other than zero small values of coupling voltage first produced a slight decrease and further augmentation of this voltage produced an increase in the oscillatory power.

(7) Any increase in oscillatory power obtained by this method can of course be reproduced with a noncoupled system by changing the excitation voltages of the tube.

It must be remembered that other operating points and tube potentials produce a variety of intersections between the circuit loss curves and the calculated available power curves plotted as a function of the oscillatory voltage, and thus the experimental findings summarized previously must not be considered to apply generally.

Acknowledgment

The writer is indebted to Messrs. D. G. Fink and J. P. Cavanagh for their aid in obtaining some of the experimental data for this circuit.

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OPTIMUM DESIGN OF TOROIDAL INDUCTANCES*

By

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Summary—*This paper gives the development of formulas for optimum design of single turn toroidal inductances to be used in ultra-high-frequency tank circuits. Two fundamental types of inductances are analyzed. Equations and curves for condition of maximum Q and parallel resistance R' are set forth. The determination of radio-frequency resistance, operating conditions, and possible uses are discussed.*

Introduction

With the recent development of ultra-high-frequency communication has come the problem of transmitter frequency stability. One solution to this problem has been described by Kolster.[1] He has modified the conventional LC tank circuit so that a high Q may be obtained at these superfrequencies. The inductance is composed of a single turn current sheet path of toroidal shape with a slot around the periphery. Capacity for tuning this inductance is obtained by flanges attached to the edges of the slot. The resonant frequency is varied by changing the spacing of the flange plates which changes the circuit capacity. The resistance of this combination is very low and a high Q results even with small values of inductance.

For convenience in mathematical analysis toroidal inductances of single turn may be divided into two fundamental classes. When cut by a plane through the axis the first will show a flux path of circular cross section. The second will have a flux path of rectangular cross section. In this discussion the former will be referred to as a *torus* and the latter as a *toroid*. A third possibility exists by combining the two. This will be called an *oval*. These shapes with their dimensions are shown in Fig. 1.

The Torus

The relation for the inductance[2] of this type is

$$L = 0.0319(r_2 - \sqrt{r_2^2 - r_1^2}) \text{ microhenrys} \tag{1}$$

where r_1 and r_2 are in inches.

* Decimal classification R382. Original manuscript received by the Institute, February 15, 1935; revised manuscript received by the Institute, May 1, 1935.

[1] F. A. Kolster "Generation and utilization of ultra-short-waves in radio communication," Proc. I.R.E., vol. 22, pp. 1335–1354; December, (1934).

[2] "Radio instruments and measurements," Bureau of Standards Circular No. 74, p. 251, (1924).

The resistance can be determined by the following method. The resistance of one increment of the surface will be

$$\Delta R = M \frac{\Delta l}{2\pi a} \tag{2}$$

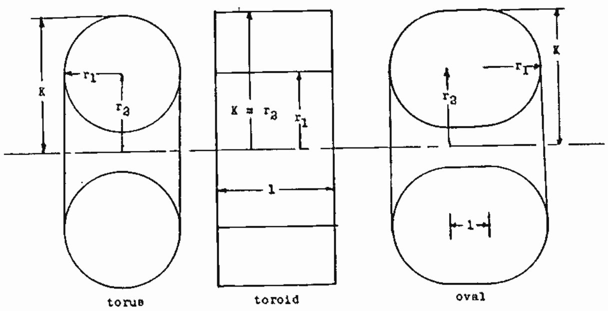

Fig. 1—Types of toroidal inductances.

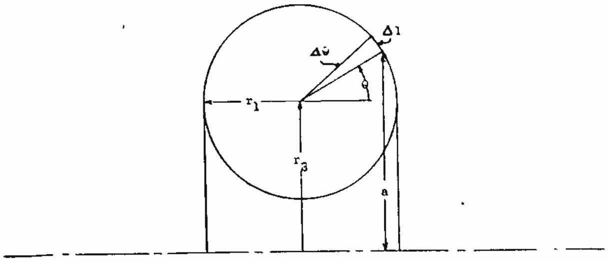

Fig. 2—Method of calculating resistance of torus.

where M is the resistivity constant. It is a function of the material and the frequency and will be discussed later. Also $\Delta l = r_1 \Delta\theta$ and $a = r_2 + r_1 \sin\theta$. The resistance of the torus will be

$$R = 2M \int \frac{\Delta l}{2\pi a} = \frac{r_1 M}{\pi} \int_{-\pi/2}^{+\pi/2} \frac{d\theta}{r_2 + r_1 \sin\theta}. \tag{3}$$

When integrated this will take the form

$$R = \frac{r_1 M}{\pi} \cdot \frac{2}{\sqrt{r_2^2 - r_1^2}} \left[\tan^{-1}\left(\frac{r_2 + r_1}{\sqrt{r_2^2 - r_1^2}}\right) - \tan^{-1}\left(\frac{r_1 - r_2}{\sqrt{r_2^2 - r_1^2}}\right)\right] + C \tag{4}$$

where C is an integration constant equal to zero. By applying

$$\tan(\alpha - \beta) = \frac{\tan\alpha - \tan\beta}{1 + \tan\alpha \tan\beta} \tag{5}$$

to the part within the brackets of (4) it can be shown this is equal to $\pi/2$ so that

$$R = \left[\frac{r_1}{\sqrt{r_2^2 - r_1^2}}\right] M \quad \text{ohms}. \tag{6}$$

When the outside diameter $2K$ and the inductance L of this type have been chosen the two diameters and resistance are automatically determined. Curves for various values L, K, R/M and ratios r_1/r_2 are shown in Fig. 3.

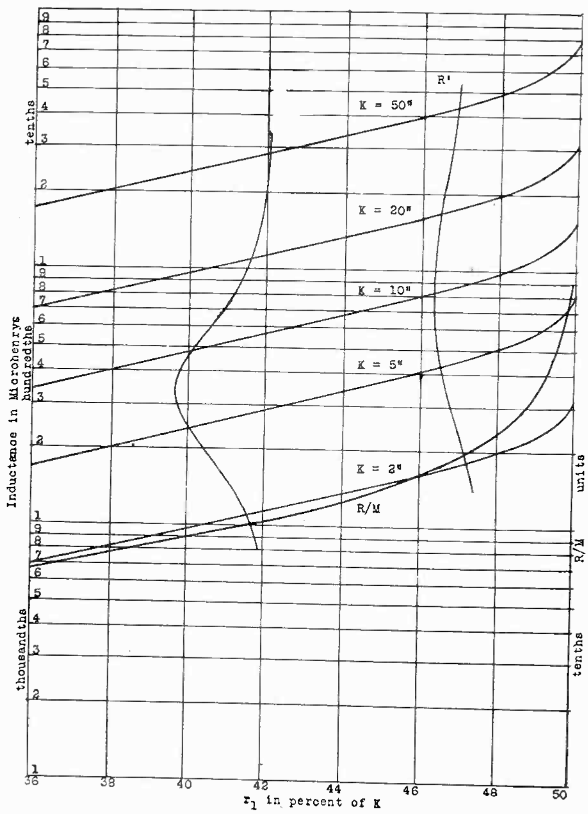

Fig. 3—Graph showing optimum design constants for the torus.

For any fixed frequency the product LC is a constant. From $LC\omega^2 = 1$, $C = 1/L\omega^2$. Also $Q = \sqrt{L/R^2C}$. Let $0.0319 = C_1$ in (1). Substituting these into the equation for Q

$$Q = \frac{1}{R}\sqrt{\frac{L}{C}} = \frac{L\omega}{R} = \frac{C_1\omega}{M}\left[\frac{r_2\sqrt{r_2^2 - r_1^2} - r_2^2 + r_1^2}{r_1}\right] \tag{7}$$

differentiating this with respect to either r_1 or r_2 and setting the derivative equal to zero

$$r_2^2[r_2 + 2r_1] - r_1^3 - \sqrt{r_2^2 - r_1^2}\,[r_1 + r_2]^2 = 0. \tag{8}$$

This is the relation between r_1 and r_2 for any given ω and outside diameter $2K$ $(K = r_1 + r_2)$ to produce the maximum Q. It gives the points where R/M increases at the same rate as L. This is plotted and shown as curve Q in Fig. 3.

By similar procedure curves for maximum effective shunt resistance R' offered by the parallel resonant circuit may be determined.

$$R' = \frac{L}{RC} = \frac{L^2\omega^2}{R}$$

$$= \frac{C_1^2\omega^2}{M}\left[\frac{(2r_2^2 - r_1^2)\sqrt{r_2^2 - r_1^2} - 2r_2(r_2^2 - r_1^2)}{r_1}\right]. \tag{9}$$

The optimum relation between r_1 and r_2 will be found to be

$$2\sqrt{r_2^2 - r_1^2}(r_1^2 - 2r_1r_2 - r_2^2) - (2r_1^3 + 3r_1^2r_2 - 4r_1r_2^2 - 2r_2^3) = 0. \tag{10}$$

It gives the point where R/M increases as the square of L. This is plotted and shown as curve R' in Fig. 3.

The Toroid

The relation for inductance[3] of this type is

$$L = 0.00508l \log_e \frac{r_2}{r_1} \text{ microhenrys} \tag{11}$$

where r_1, r_2, and l are in inches.

The resistance can be found by integration. Fig. 4 shows the fundamental shapes the toroid is composed of; namely, two cylinders and two rings.

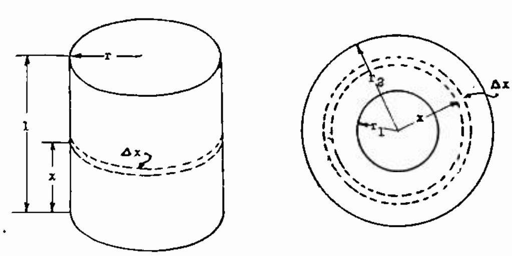

Fig. 4—Component parts of toroid.

The resistance of an element of the ring will be $\Delta R = \Delta x / 2\pi x$. Setting up the integral and integrating between limits r_1 and r_2

[3] J. H. Morecroft "Principles of Radio Communication," Second Edition, pp. 198 and 199.

$$R_R = \left[\frac{1}{2\pi}\log_e\frac{r_2}{r_1}\right] M \text{ ohms.} \tag{12}$$

The resistance of an element of the cylinder will be $\Delta R = \Delta x / 2\pi r$. Setting up the integral and integrating between limits O and l

$$R_C = \left[\frac{l}{2\pi r}\right] M \text{ ohms.} \tag{13}$$

The total resistance of the toroid will be

$$R = \left[\frac{l}{2\pi}\left(\frac{1}{r_1}+\frac{1}{r_2}\right)+\frac{1}{\pi}\log_e\frac{r_2}{r_1}\right] M \text{ ohms.} \tag{14}$$

When an outside diameter $2K$ and inductance L have been chosen it will be possible to find an infinite number of r_1 and l values to satisfy the requirements. From an inspection of the possibilities it will be seen that the limits are shown in Fig. 5.

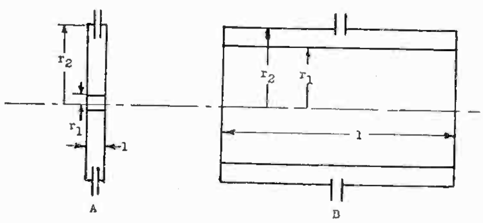

Fig. 5—Possible extreme shapes of toroid.

Both of these figures will have a large resistance for any given inductance and value $2K$. Fig. 5 A will have high resistance because of large ratio of diameters r_2/r_1 and Fig. 5 B because of great length l. To find the optimum ratio of diameters and length l for given outside diameter and inductance to produce minimum R the equations for L and R, (11) and (14), are differentiated with respect to either l or r_1 and set equal to zero. The derivatives with respect to l or r_1 are set equal to each other and solved for l. This gives:

$$= \frac{2r_1r_2\log_e\dfrac{r_2}{r_1}}{r_1+r_2-r_2\log_e\dfrac{r_2}{r_1}}. \tag{15}$$

This value of l is then substituted into (11). With r_2 taken at various values of K the values of L and R/M are determined for given ratios of r_1 to K and shown in Fig. 6.

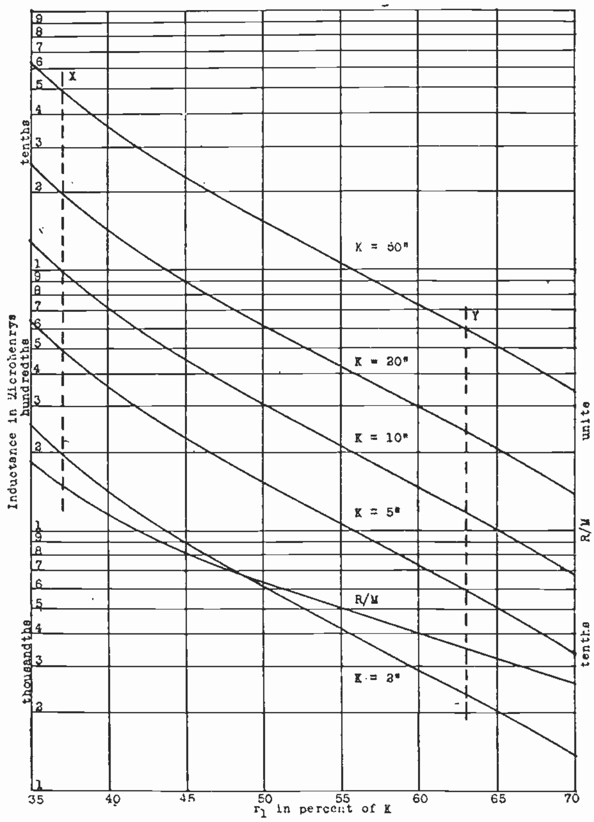

Fig. 6—Graph showing optimum design constants for toroid.

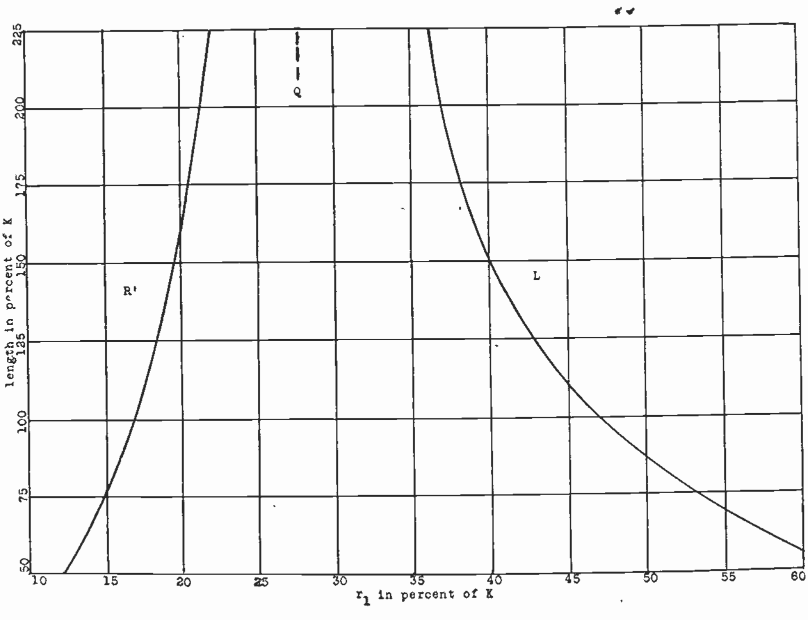

Fig. 7—Graph showing optimum design constants for toroid.

The values of l in per cent of K are determined from (15) and plotted against r_1 in per cent of K and shown by curve L in Fig. 7. The lines denoted by X and Y in Fig. 6 show the values of $l=2K$ and $1/2\,K$, respectively. These are arbitrarily chosen as useful design limits.

By the same procedure as used in (7) Q is determined as

$$Q = \frac{\pi C_1 \omega}{M}\left[\frac{2r_1 r_2 \log_e \dfrac{r_2}{r_1}}{2(r_1 + r_2) - r_2 \log_e \dfrac{r_2}{r_1}}\right] \tag{16}$$

where $C_1=0.00508$ from (11) and $r_2=K$.

When this is differentiated with respect to r_1 there results an equation from which no real value of r_1 can be found. This is because $\Delta R/\Delta L$ never becomes equal to 1 on curves of L vs R/M over the optimum range. By inspecting Fig. 6 it will be seen that as ratio r_1/K decreases the curve R/M becomes more nearly parallel to curves of L. The denominator of (15) will equal zero when $r_1=27.85$ per cent (approximately) of r_2 and for this value an infinite l will result. At this limit $\Delta R/\Delta L=1$ and maximum Q is attained. From this it can be seen that slightly higher values of Q can be obtained for given L by using long, small diameter toroids.

This can be shown in different manner by letting $l=Jr_2$ where J is a constant. Substituting this into (11) and (14) and combining in the manner of (7)

$$Q = \frac{2\pi\omega C_1 J}{M}\left[\frac{r_2 r_1 \log_e \dfrac{r_2}{r_1}}{J(r_1 + r_2) + 2r_1 \log_e \dfrac{r_2}{r_1}}\right] \tag{17}$$

differentiating this with respect to r_1 and simplifying the J drops out and

$$r_2 \log_e \frac{r_2}{r_1} = r_1 + r_2 \tag{18}$$

which is the same conclusion arrived at before and shows that there is no finite value of l that will produce maximum Q.

The equation for optimum R' may be obtained by using the same procedure used to get (17) and (9).

$$R' = \frac{2\pi\omega^2 J^2 C_1^2}{M}\left[\frac{r_1 r_2^2\left(\log_e \dfrac{r_2}{r_1}\right)^2}{J(r_1 + r_2) + 2r_1 \log_e \dfrac{r_2}{r_1}}\right] \text{ ohms.} \tag{19}$$

Differentiating this with respect to r_1 and solving for J

$$J = \frac{2r_1 \log_e \dfrac{r_2}{r_1}}{r_2 \log_e \dfrac{r_2}{r_1} - (r_1 + r_2)} = \frac{l}{r_2}. \tag{20}$$

This is plotted in terms of l as curve R' in Fig. 7 and is also asymptotic to value $r_1 = 27.85$ per cent r_2. Curves similar to those of Fig. 6 could be plotted for maximum R' values.

Throughout this discussion of the toroid the value r_2 has been maintained a constant equal to K.

The Oval

This type is a combination of the torus and the toroid. The inductance of any turn is equal to the summation of all possible pairs of elements of the turn. These mutual inductances depend on the positions of the elements. While not strictly accurate the formula for inductance of the oval can be obtained to a good approximation by combining the formulas for the inductances of the torus and the toroid. Looking at this from a qualitative viewpoint a change in current I in the oval will produce a change in flux linkages in the center rectangular portion approximately equal to that change I would cause in a toroid of equal dimensions. Similarly for the semicircular end sections. As l of the oval approaches infinity or zero causing this shape to approach the toroid or torus as limits the closer the above statements come to being exact. On this basis (1) and (11) may be combined into (21) by simple addition and a proper change of notation for the toroid part. Like the torus $r_1 + r_2 = K$.

$$L = C_1 l \log_e \left(\frac{r_2 + r_1}{r_2 - r_1}\right) + C_2(r_2 - \sqrt{r_2^2 - r_1^2}) \text{ microhenrys} \tag{21}$$

where $C_1 = 0.00508$, $C_2 = 0.0319$.

In the same manner the resistances of the torus and the toroid may be added to find the resistance of the oval.

$$R = \left[\frac{l}{2\pi}\left(\frac{1}{r_1 + r_2} + \frac{1}{r_2 - r_1}\right) + \frac{r_1}{\sqrt{r_2^2 - r_1^2}}\right] M \text{ ohms}. \tag{22}$$

By the same reasoning and processes used in obtaining (15) the optimum length is found to be

$$l = \frac{(K-r_1)\left[C_1\pi\sqrt{K}\sqrt{K-2r_1}\log_e\frac{K}{K-2r_1} - C_2(\sqrt{K}\sqrt{K-2r_1} - (K-2r_1))\right]}{C_1\left[2(K-r_1) - K\log_e\frac{K}{K-2r_1}\right]}. \tag{23}$$

The denominator here will be zero when $r_1 = 36.1$ per cent r_2 and the inside diameter will be 27.85 per cent of K showing the oval has the same characteristics as the toroid when r_1 and r_2 are chosen to give large values of l.

To set up equation for condition of maximum Q let $l = JK$ and $r_1 + r_2 = K$. Substitute these into (21) and (22) and combine in manner of (7).

$$Q = \frac{\pi\omega\sqrt{K}(K-2r_1)}{M}\left[\frac{C_1 J\log_e\frac{K}{K-2r_1} + C_2(K - r_1 - \sqrt{K}\sqrt{K-2r_1}}{r_1\pi\sqrt{K-2r_1} + J(K-r_1)\sqrt{K}}\right]. \tag{24}$$

Differentiating this with respect to r_1, setting the derivitive equal to zero and solving for J the result comes out in the form of a quadratic in J.

$$\begin{aligned} &J^2\left[C_1K^2\left(2K - 2r_1 - K\log_e\frac{K}{K-2r_1}\right)\right] \\ &+ JK\left[\pi C_1\sqrt{K}\sqrt{K-2r_1}\left(2r_1 - K\log_e\frac{K}{K-2r_1}\right)\right. \\ &\left.+ C_2(K-r_1)(\sqrt{K}\sqrt{K-2r_1} - 2(K-r_1))\right] \\ &+ C_2\pi\sqrt{K}\sqrt{K-2r_1}\left[K^{3/2}\sqrt{K-2r_1} + r_1^2 - K^2 + r_1K\right] = 0. \end{aligned} \tag{25}$$

For any values of r_1 up to $1/2K$, J will have two real positive roots. One will be of large value corresponding to the conditions of great length as found in the toroid and the other will be of very small length approaching the values found in the torus as l approaches zero.

Equations to obtain maximum R' can be set up in manner of (9). However they are of little practical use as the result is a third degree equation in J with very complex coefficients. Two roots will be real and positive approximating the conditions found in the toroid and torus while the third one will be negative.

Curves similar to Figs. 3, 6, and 7 could be drawn. However the net result would only be values between the two sets determined for the torus and the toroid because the oval approaches either one of these shapes as limits.

Addendum

Various other odd cross sections could be taken up such as ellipses, triangles, egg-shapes, etc., but it is doubtful whether any increase in either Q or R' would be secured with given over-all dimensions by the use of such shapes.

All the above calculations have been made on the assumption of lumped constants. The torus because of its geometric shape will probably have the least amount of distributed inductance and capacity. Very long toroids will approach a line in characteristics.[4] Such an inductance will have an outside to inside diameter ratio of 3.6 for least attenuation.

In regard to the accuracy of the formulas for inductance (1) and (11) the following may be quoted from reference (2) page 251. "The value so computed is strictly correct only for an infinitely thin winding. For a winding of actual wires a correction may be calculated as shown in the Bulletin of the Bureau of Standards, No. 8, page 125; 1912. The correction is, however, very small." From the above it can be seen that the higher the frequency the more pronounced the skin effect becomes and the nearer this type of inductance approaches the specified infinitely thin winding. In any case the error is probably less than 0.1 per cent.

The method of calculating resistance is open to some question on the basis that the current density at the center portions of the inductance is high from two causes. First, the surface area is less than that at other points. Second, the inductance if of short length may have appreciable stray capacity within from one end to the other. These effects might force the current deeper into the conductor and cause an apparent increase in the resistivity constant. This would cause M to be an inverse function of the diameter. If such a situation could be taken into account the optimum values of any given radius such as r_1 in the toroid and oval and the values of Q and R' in the torus would be increased slightly.

[4] Sterba and Feldman "Transmissions lines for short-wave radio systems," Proc. I.R.E., vol. 20, p. 1171; July, (1932).

The resistance constant M is a function of frequency and the material. The conducting surface layer will have a depth[5] of $6.62/\sqrt{f}$ cm at unit permeability. Copper has a resistivity[6] of 1.72×10^{-6} ohms per centimeter cube. Combining and changing to inches,

$$M = \frac{\rho l}{a} = \frac{1.72\cdot10^{-6}\cdot2.54}{2.54\cdot\dfrac{6.62}{\sqrt{f}}} = 0.260\cdot10^{-6}\sqrt{f} \tag{26}$$

where f is in cycles per second.

TABLE I

Column	A	B	C	D	E	F
Type of Inductance	Toroid	Toroid	Toroid	Torus	Torus	Torus
Frequency in mc			60			
Constant M			2.015×10^{-3}			
Constant LC			7.05×10^{-18}			
Outside Diameter			15 inches			
Inductance in Microhenrys	0.0672	0.0672	0.0730	0.0672	0.0355	0.0635
Capacity in Micromicrofarads	105	105	96.5	105	198	111
r_1 in per cent of K	20	37.8	17.5	46.8	39.6	46.3
l in per cent of K	110	182	110	—	—	—
R in 10^{-3} ohms	3.15	2.72	3.56	3.77	1.75	3.43
Q	8050	9300	7700	6710	7650	6950
R' in 10^3 ohms	204	235	213	165	102	169

Table I shows relatively what various designs will accomplish with fixed diameter $2K$. For comparative purposes a frequency of 60 megacycles is chosen and the largest toroid described by Kolster[1] is taken as a standard. Its characteristics are in column A. Column B is toroid of same LC ratio as A with optimum dimensions for maximum Q and R'. C is toroid to produce maximum R' with same over-all dimensions of A. D is torus using same LC ratio as A. E is torus to produce maximum Q. F is torus to produce maximum R'. Reference to Kolster's article will show a rather wide discrepancy between the Q value he shows and the one in Table I, column A. This is probably due to using different values of M.

Inspection of all the equations derived above will show that the values of Q, R', L, etc., can be increased indefinitely as the diameter and length of the inductance are increased. Just what the limits are is open to conjecture but experience has shown that the diameter $2K$ should be less than a quarter of a wavelength and preferably nearer 10 per cent λ. The length probably should not be greater than $5K$ at most and not over 25 per cent λ. If the inductance is made larger than the suggested dimensions the percentage of distributed constants will in-

[5] F. E. Terman, "Resonant lines in radio circuits," *Electrical Eng.*, p. 1047; July, (1934).

[6] "Standard Handbook," 9th Edition, p. 547.

crease and the equations will not have much meaning as the apparatus will then tend to act as a line.

Kolster has shown that the optimum operating point for the tank will be at a condition where

$$L_0 C_0 \omega^2 = 1 - \frac{1}{Q^2} \tag{27}$$

where L_0 and C_0 are the tank inductance and capacity. The frequency of oscillation will be determined by $L_0 C \omega^2 = 1$. C is the total circuit capacity $C_0 + C_s$ where C_s is circuit stray (the driving tube). Substituting these into (27)

$$L_0 C_0 \omega^2 = L_0(C_0 + C_s)\omega^2 - \frac{1}{Q^2} \tag{28}$$

since,

$$L = 1/\omega^2(C_0 + C_s)$$

$$\frac{C_0}{C_0 + C_s} = 1 - \frac{1}{Q^2}. \tag{29}$$

Solving this for C_0,

$$C_0 = C_s(Q^2 - 1). \tag{30}$$

This will give a very high value of C_0 for any large value of Q. However, it must be remembered that C_s is in series with a large positive reactance in the form of leads across C_0 so that the effect will be the same as tapping C_s down on the inductance instead of directly across C_0.

While these tank circuits enable a high Q to be obtained for frequency stability on wavelengths below ten meters they are also useful in building up a large parallel resistance R' for an efficient power amplifier or to obtain gain in a tuned radio-frequency type receiver. Their selectivity characteristics should be useful for image and noise suppression in a receiver or to reduce harmonic radiation from a power amplifier working on a distorted wave form.

Tuning by change of spacing of the condenser plates produces too critical adjustments to be used for a receiver. One suggested way of accomplishing this is to mount a small trimmer on the condenser rim as shown in Fig. 8. The rough adjustments are made by varying condenser plate spacing in the usual way. The frequency coverage can be determined by spacing of the trimmer plate distance "a". A very slight amount of stray L and C is introduced by this method. The trimmer should preferably be mounted directly opposite the driving tube.

Another simple way of changing the frequency one per cent or less is to insert a thin sheet of mica, isolantite or other insulator between the flanges. This sheet of insulating material should be less than 25 per cent

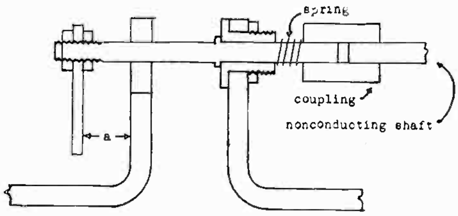

Fig. 8—Design of vernier tuning arrangement.

of the flange spacing to prevent a flashover due to lowering of breakdown potential caused by a too severe change of the electrostatic field.[7] A strip of metal can be used if the reduction of the air gap is not too great.

[7] T. Walmsley, "Notes on the design of radio insulators," Proc. I.R.E., vol. 16, p. 363; March, (1928).

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SOME POSSIBILITIES FOR LOW LOSS COILS*

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Summary—*A toroidal coil wound with a flat conductor so curved that the flat side follows exactly the surface of the toroid has remarkably low losses. The Q ($=\omega L/R$) of such a coil is independent of the number of turns, is proportional to size, and to the square root of frequency. For maximum Q there is a best proportion which is given in the paper for the case of toroids of rectangular and circular cross sections, together with formulas for calculating the losses and the Q for such toroids. These formulas show that values of Q between 1000 and 10,000 are theoretically realizable with reasonable dimensions. The practical difficulties of realizing the theoretical possibilities are discussed, and it is shown that the mechanical construction must have a perfection comparable to the skin depth of current penetration.*

THE ordinary tuning inductance used in radio circuits has an effective resistance at its working frequency which is normally many times the resistance to direct currents. This is because the flux lines, which cut across the conductor, give different portions of the cross section different impedances, with the result that most of the current is concentrated in the small part of the cross section having the lowest impedance. Thus in the single layer solenoid of Fig. 1 the current concentrates in the shaded regions and only a small part of the conductor is effective in carrying current.

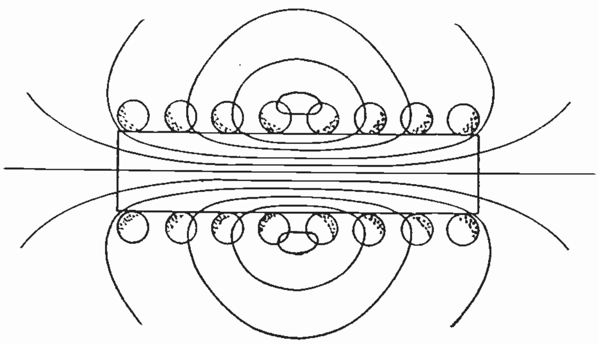

Fig. 1.—Flux and current distribution in a single layer solenoid, showing how the current concentrates in those parts of the conductor circled by the fewest flux lines. The current density is indicated by the depth of shading.

One means of minimizing this tendency is to employ a strip conductor so oriented that the flat side is substantially parallel to the

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flux lines at each point. While this idea may be applied to solenoid coils, its most practicable form is in toroidal coils in which the flat surface of a strip conductor is shaped to fit as nearly as possible the surface of revolution represented by the toroid. Such coils are shown in Fig. 2 and it will be observed that the flux lines are everywhere parallel to the flat surface. The result is that the current is distributed uniformly over one side of the strip instead of being concentrated in a small region as in Fig. 1, and the effective resistance is therefore greatly reduced.

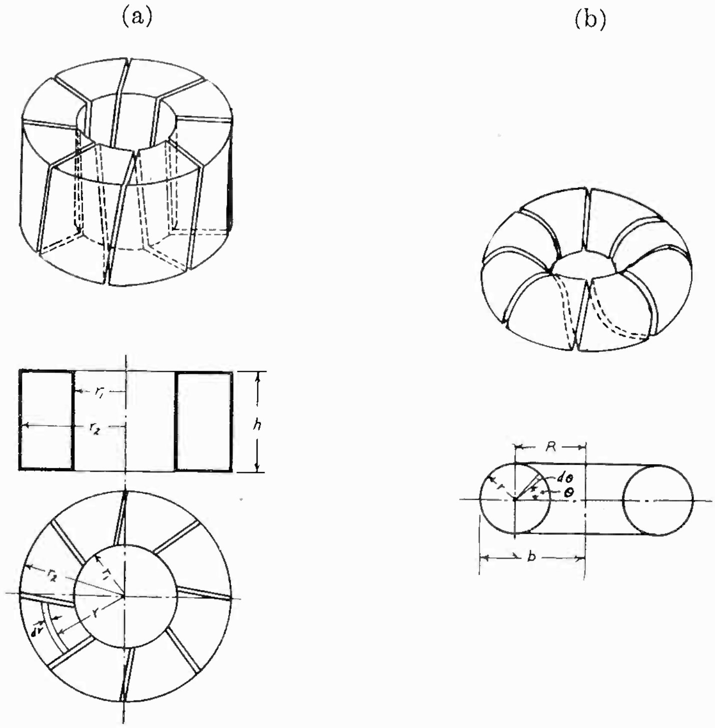

Fig. 2—Toroidal coils wound with strip conductor that is curved to follow the surface of revolution formed by the toroid. Note that the magnetic flux lines produced by current passing through the coil are everywhere parallel to the flat side of the immediately adjacent conductor.

The simple geometrical configuration of a toroid makes it possible to calculate the resistance and Q of toroids such as shown in Fig. 2. Since the depth to which the current penetrates the strip conductor is small compared with the radius of curvature of the surface, one can calculate the resistance using the skin-effect formula for a flat surface.[1] When such a conductor is not too thin the resistance which it offers is the same as though the current were uniformly distributed over a "skin depth," which in the case of copper is $0.00662/\sqrt{f}$ centimeters, where f is in megacycles. A toroid with rectangular cross section (Fig. 2(a))

[1] Kennelly, Laws, and Pierce, "Experimental researches on skin effect in conductors," *Trans. A.I.E.E.*, vol. 34, p. 1953, (1915).

made of copper, and with the space between turns a negligible fraction of the total circumference, hence has a resistance of

$$\text{Resistance} = \frac{\sqrt{f}N\ 1.724 \times 10^{-6}}{0.00662}\left[\frac{hN}{2\pi r_1} + \frac{hN}{2\pi r_2} + 2\int_{r_1}^{r_2}\frac{N}{2\pi r}\,dr\right]$$

$$= 41.5\sqrt{f}N^2\,10^{-6}\left[\frac{h}{r_1} + \frac{h}{r_2} + 2\log_\epsilon\frac{r_2}{r_1}\right] \qquad (1)$$

where N is the number of turns, f the frequency in megacycles, and the dimensions are in centimeters. Since the inductance of the coil is $0.004606\ N^2h\ \log_{10} r_2/r_1$

$$Q = \frac{\omega L}{R} = \frac{303 r_2\sqrt{f}}{\left(\left(\frac{2r_2}{h} + \frac{1+\frac{r_2}{r_1}}{\log_\epsilon\frac{r_2}{r_1}}\right)\right)}. \qquad (2a)$$

If h and r_2 are constant, Q is maximum where $r_2/r_1 = 3.59$. With these optimum proportions

$$Q \text{ for optimum } r_2/r_1 = \frac{84.3\sqrt{f}\,r_2}{\left(1 + 0.556\frac{r_2}{h}\right)}. \qquad (2b)$$

When the toroid has a circular cross section (Fig. 2(b)) the resistance is given by the expression

$$\text{Resistance} = \frac{\sqrt{f}N\ 1.724 \times 10^{-6}}{0.00662 \times 2\pi}\int_0^{2\pi}\frac{d\theta}{\frac{b-r}{r} - \cos\theta}$$

$$= 0.00026\sqrt{f}N^2\ \frac{r}{R}\ \frac{1}{\sqrt{1-\left(\frac{r}{R}\right)^2}}. \qquad (3)$$

Since the inductance in microhenrys is $0.01257\ N^2(R - \sqrt{R^2 - r^2})$, one has

$$Q = \frac{\omega L}{R} = 303\sqrt{f}R\left[\frac{1-\sqrt{1-(r/R)^2}}{\left(\frac{r}{R}\right)}\right]\sqrt{1-\left(\frac{r}{R}\right)^2}. \qquad (4a)$$

If the outer radius $b=R+r$ is kept constant as r/R is varied, then Q will be greatest when $r/R=0.71$. For this optimum proportion

$$Q \text{ for optimum proportion } = 52.1\, b\sqrt{f}. \tag{4b}$$

The outstanding properties of the toroidal coils of Fig. 2 revealed by (2) and (4) are:

First, the Q is independent of the number of turns.

Second, the Q is proportional to the square root of frequency so that, unlike ordinary coils, the higher the frequency the greater the Q.

Third, the Q is directly proportional to the linear dimension of the coil provided the proportions are kept constant.

Fourth, with fixed outer dimensions there is an optimum inner diameter which will give the highest Q.

The values of Q theoretically obtainable with reasonable proportions are amazingly high. Thus a toroid such as shown in Fig. 2(a) one foot high and one foot in outer diameter will have a Q of 3200 at 10 megacycles.[2] At lower frequencies even higher values are possible because the size can be increased inversely with the frequency. Thus at 1000 kilocycles a coil ten times the size of the above example would have a Q of about 10,000, and while the dimensions appear absurdly large at first glance they are not necessarily so impracticable when it is realized that there is no external magnetic field, and that a toroid of such large dimensions would need have only a single turn.

Investigation was made of the possibilities of a multilayer toroidal coil of the type herein discussed, but as a result of the analysis outlined in the Appendix it was found that the inside turns would carry large eddy currents unless the thickness of the inside conductor was of the same order of magnitude as the "skin depth." It is therefore necessary that the winding be confined to a single layer, in spite of the fact that a multilayer construction permits the use of wider turns.

The analysis and discussion given above assumes that perfect mechanical symmetry is obtained, and in particular that the surfaces of the conductor which carry the current are sections from a perfect figure of revolution. The increase in resistance occasioned by imperfections cannot be predicted mathematically, but the analysis given in the Appendix for the proximity effect in a multilayer coil indicates

[2] It is interesting in this connection to note that F. A. Kolster using what is essentially a one turn toroid of the type shown in Fig. 2(a) but with the inner radius somewhat less than the optimum value, has reported values of Q of several thousand at 20 to 60 mc. F. A. Kolster, "Generation and utilization of ultra-short waves in radio communication," Proc. I.R.E., vol. 22, pp. 1335–1354; December, (1934).

that the actual losses will be considerably greater than those existing under ideal conditions if the mechanical imperfections of the surfaces exceed appreciably the "skin depth." This means that each turn must be so curved and orientated that when taken in relation to the adjacent turns it is not out of line more than about $0.00662/\sqrt{f_{mc}}$ centimeters. This represents a high degree of precision, particularly at the higher frequencies, and makes the problem of completely realizing the theoretical possibilities a difficult one.

A number of constructional methods suggest themselves. One possibility would be to use a solid insulating core upon which a copper coating could be plated. Turns could then be formed by cutting the coating. A variation of this method is to use a wax or type-metal core which is melted out after the turns are formed and securely fixed in place by fastening to insulating rings. The coil shown in Fig. 2 (a) can be built up using pieces cut from cylinders for the strips along the h direction, and flat members for the ends.

The type of toroidal coil herein described fills in the gap between the low frequencies at which litz wire performs satisfactorily, and the frequencies high enough to utilize the remarkable properties of resonant lines to obtain high Q.[3] Throughout this frequency range it is theoretically possible to realize coil Q's of several thousand. Such coils when operated at substantially constant temperature would give sufficient frequency stability when used in an oscillator to meet, without difficulty, the requirements for code transmitters. It even appears that a tuned-circuit oscillator incorporating an inductance such as mentioned above having a Q of 10,000 could rather easily give the frequency stability required of broadcast stations provided the inductance was constructed well enough to have good mechanical stability.

The simplifications and economies that would result from being able to dispense with the inflexible crystal and its power amplifiers and frequency doublers, are too obvious to need discussion. Hence, while it is admittedly a difficult matter to realize the full theoretical possibilities of the new type of low loss toroid, the constructional problem is not insoluble, and the results which are to be obtained appear to be well worth the trouble.

Appendix

Proximity Effect in Miltilayer Toroid

Consider the two-layer toroid shown in Fig. 3. The current distribution and resistance for the outer layer is exactly the same as though

[3] F. E. Terman, "Resonant lines in radio circuits," *Electrical Eng.*, vol. 53, p. 1046; July, (1934).

no inner layer were present, but the magnetic flux produced by the outer layer induces eddy currents in the inner layer. These eddy currents are superimposed upon the current distribution that would exist in the absence of the outer layer, and can be thought of as being the consequence of a "proximity effect."

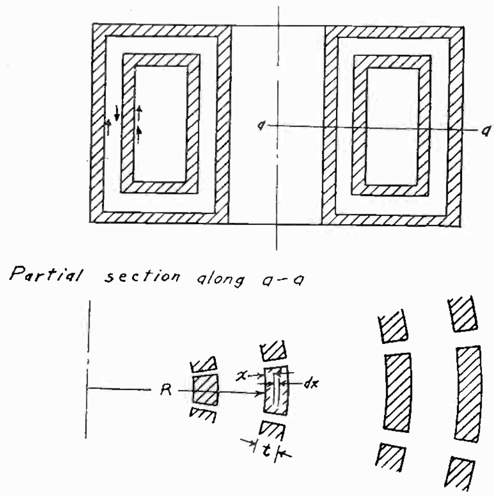

Fig. 3—Multilayer toroid wound with strip conductor.

In the case of the vertical members in Fig. 3 these eddy currents flow up one side of the inner layer and return down the other side as indicated, and greatly increase the loss in the conductor. It is possible to calculate the current distribution in the inner layer for these vertical members. The differential equation that applies is still the same as applies to the outer layer,[1] but the boundary conditions are now different. The solution is of the form

$$i_x = A \cosh \alpha x + B \sinh \alpha x$$

where,

i_x = vector current density at point x (Fig. 3(a))
$\alpha^2 = 4\pi\ \omega\gamma \times 10^{-9}$
γ = conductivity of the metal
x = distance into inner conductor from edge (Fig. 3(a))
A and B = constants of integration determined by the boundary conditions, which are

$$(1) \quad \text{at } x = 0,\ H = \frac{NI}{2\pi R}$$

$$(2) \quad \int_0^t i_x \frac{2\pi R}{N}\, dx = I$$

where,

N = turns in the outer layer

I = current passed through the toroid

H = magnetic intensity produced by current in the outer layer

t = thickness of inner layer conductor.

The first of these conditions states that the magnetic intensity at the edge of the inner layer adjacent to the outer layer is produced solely by the ampere turns of the outer layer. The second condition states that the total net current flowing in the inner conductor is the current passed through the coil. The eddy currents merely add to the current density at some points and subtract from it at others.

Substituting the first condition into the equation, and making use of the relation $di/dx = \alpha^2 H$ obtained when setting up the original differential equation, gives

$$B = \frac{\alpha N I}{2\pi R}.$$

Similarly, substituting the second condition into the equation leads to the result

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Some of the results have been studied in connection with a practical communication problem which was concerned with skipping of signals in short-distance transmission along one of the airways. The results are used to determine the limiting frequency for any distance up to a few hundred miles.

Information of the type presented here should prove useful in the study of the properties of the upper atmosphere, as well as in the interpretation of communication problems.

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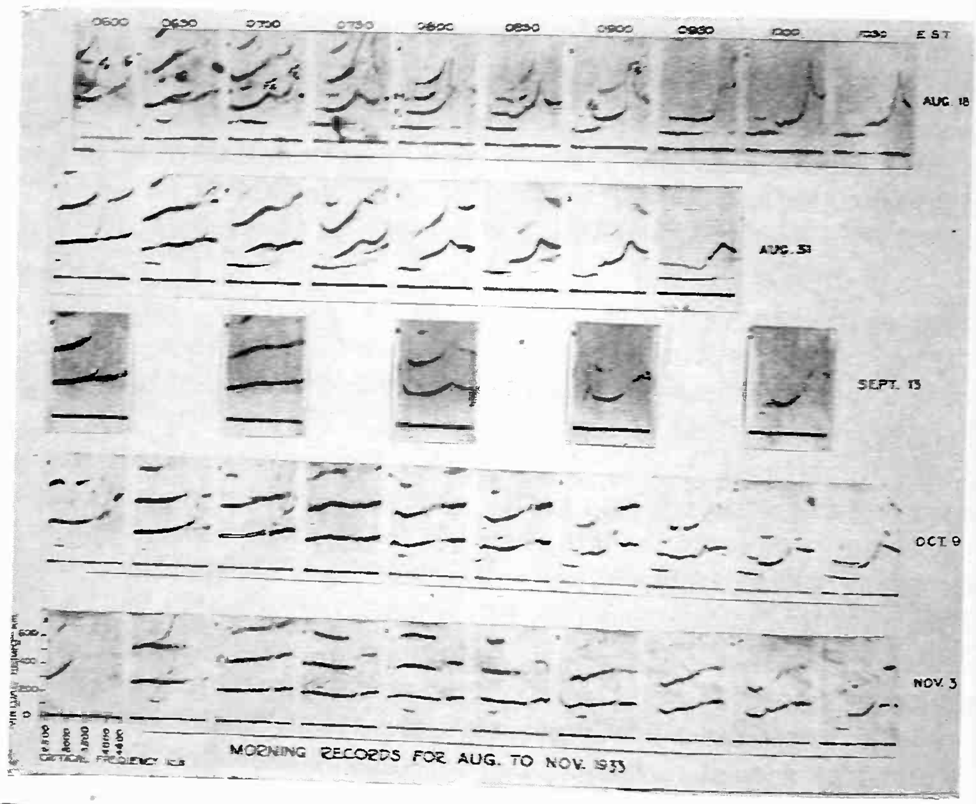

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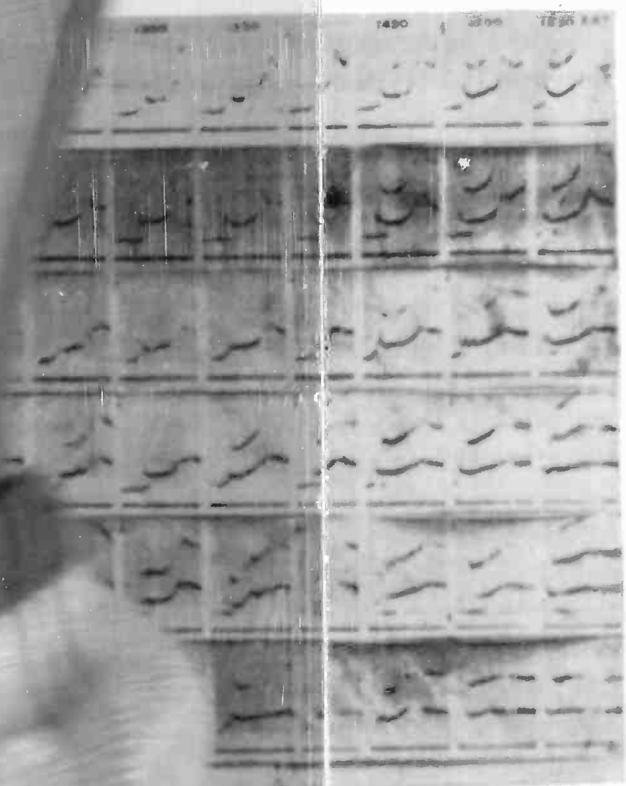

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no inner layer were present, but the magnetic flux produced by the outer layer induces eddy currents in the inner layer. These eddy currents are superimposed upon the current distribution that would exist in the absence of the outer layer, and can be thought of as being the consequence of a "proximity effect."

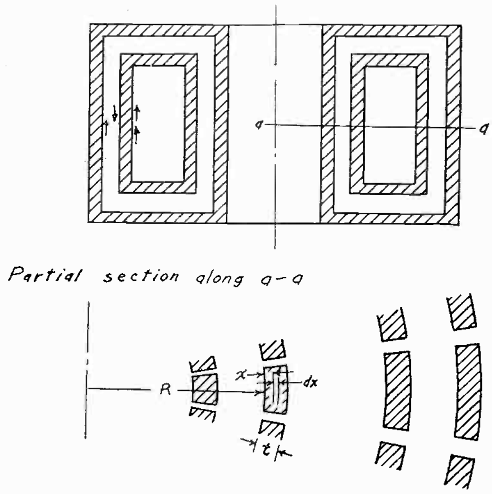

Fig. 3—Multilayer toroid wound with strip conductor.

In the case of the vertical members in Fig. 3 these eddy currents flow up one side of the inner layer and return down the other side as indicated, and greatly increase the loss in the conductor. It is possible to calculate the current distribution in the inner layer for these vertical members. The differential equation that applies is still the same as applies to the outer layer,[1] but the boundary conditions are now different. The solution is of the form

$$i_x = A \cosh \alpha x + B \sinh \alpha x$$

where,

i_x = vector current density at point x (Fig. 3(a))

$\alpha^2 = 4\pi\,\omega\gamma \times 10^{-9}$

γ = conductivity of the metal

x = distance into inner conductor from edge (Fig. 3(a))

A and B = constants of integration determined by the boundary conditions, which are

$$(1) \qquad \text{at } x = 0, \; H = \frac{NI}{2\pi R}$$

$$(2) \quad \int_0^t i_x \frac{2\pi R}{N}\, dx = I$$

where,

N = turns in the outer layer
I = current passed through the toroid
H = magnetic intensity produced by current in the outer layer
t = thickness of inner layer conductor.

The first of these conditions states that the magnetic intensity at the edge of the inner layer adjacent to the outer layer is produced solely by the ampere turns of the outer layer. The second condition states that the total net current flowing in the inner conductor is the current passed through the coil. The eddy currents merely add to the current density at some points and subtract from it at others.

Substituting the first condition into the equation, and making use of the relation $di/dx = \alpha^2 H$ obtained when setting up the original differential equation, gives

$$B = \frac{\alpha N I}{2\pi R}.$$

Similarly, substituting the second condition into the equation leads to the result

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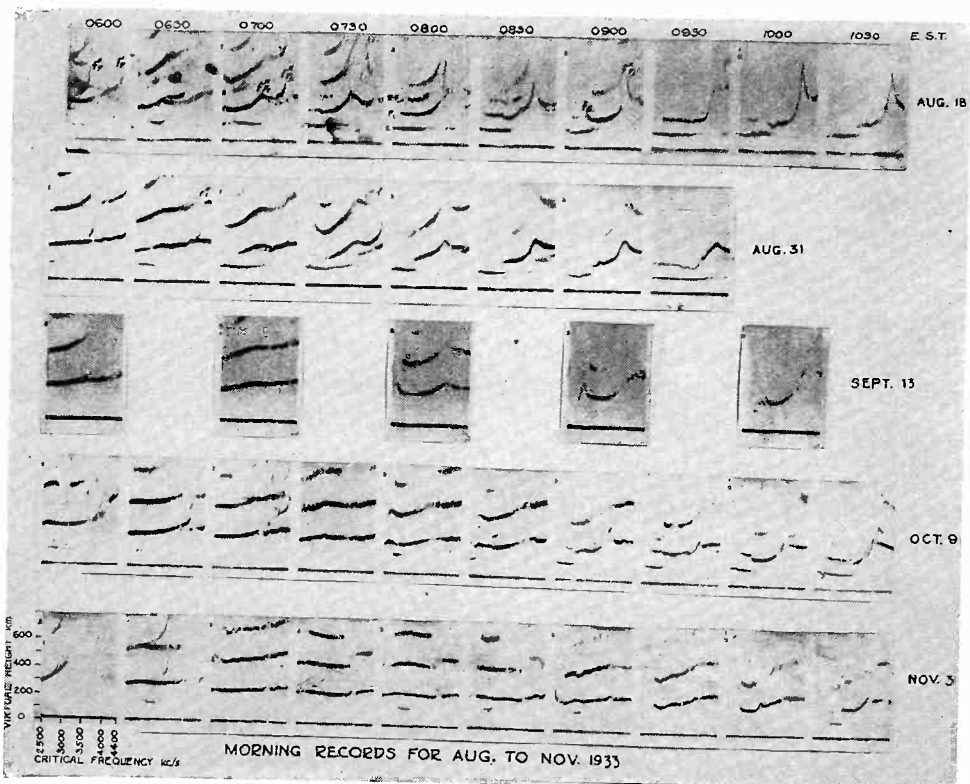

Fig. 1—Records taken during the morning at half-hour intervals showing the changes in character of the patterns from August to November, 1933.

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record at 0600 E.S.T. for August 18 shows the ordinary and extraordinary rays returned from the F region with the critical frequencies[3] f_F'' and f_F'. The critical frequency for each ray occurs at the point where the trace approaches the vertical. By 0700 the ordinary ray critical frequency is above 4000 kilocycles and formation of the F_1 layer is beginning to appear. The pattern at 0800 shows F_1 refraction with one multiple, as well as both E and E–F reflections.[4] E–F reflections are those which penetrate the E layer, then make two round trips between the top of the E and the bottom of the F layer before finally coming

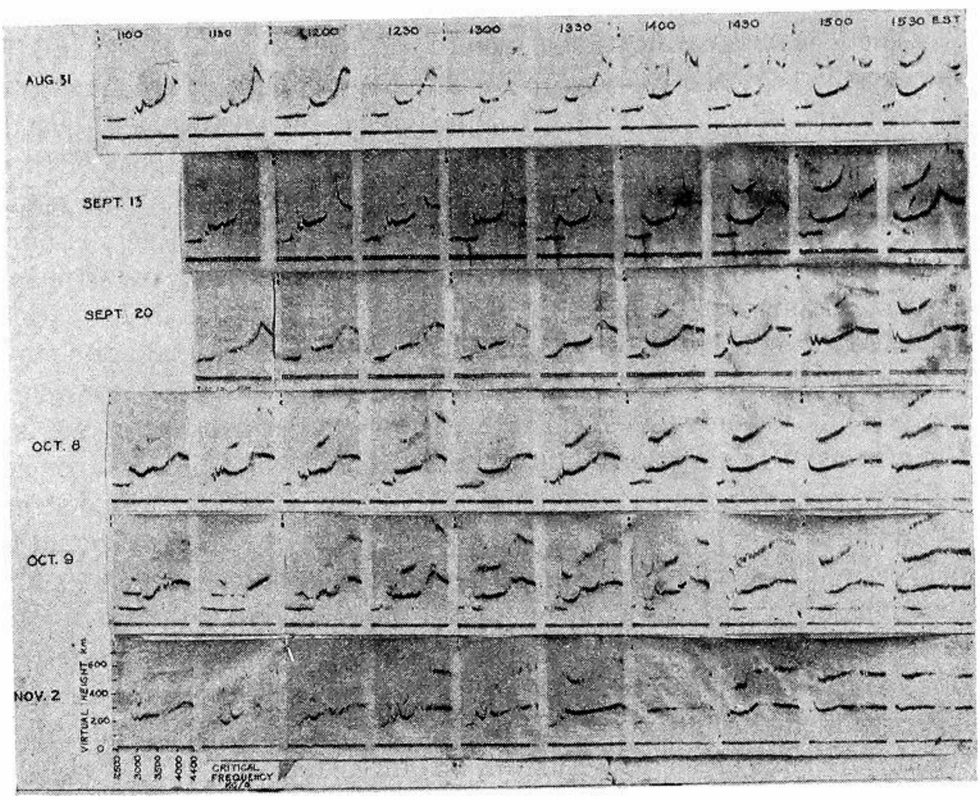

Fig. 2—Records showing changes in the character of the patterns from August to November, 1933, between 1100 and 1530 E.S.T.

down through the E layer to earth. This type of reflection is quite common and can be differentiated from the others because its trace

[3] The following nomenclature for critical frequencies is used:
f_E for E-layer critical frequency.
f_{F_1}'' for F_1-layer critical frequency. Ordinary ray.
f_{F_1}' for F_1-layer critical frequency. Extraordinary ray.
f_F'' for F-layer critical frequency. Ordinary ray.
f_F' for F-layer critical frequency. Extraordinary ray.
Attention has just been called to nomenclature adopted at the London meeting of the U.R.S.I. in September, 1934. The U.R.S.I. nomenclature will be used in future publications.

[4] Called *M* reflections by Ratcliffe and White in England.

has the same shape as that of the F-layer trace below it while its separation from the F trace is about the same as the separation between the E and F traces. This indicates that the reflecting E layer is fairly thin at times, perhaps less than 10 kilometers in thickness when E-F reflections appear.

The pattern at 0900 shows both the E and F_1 critical frequencies. The F_1 critical frequency becomes more pronounced later in the morning. This critical frequency is for the ordinary ray. The extraordinary ray can be seen at times but its critical frequency is usually above the limits of the present recording system.

The F-layer stratification is seen to become less pronounced and the decrease in both the E and F_1 critical frequencies can be seen as winter approaches. The records of Fig. 2 show the changes from August to November, 1933, between 1100 and 1530. At 1100 on August 31 there is a triple critical effect between the usual E and F_1 layers. "Fine structure"[5] of this type which has been noted by other observers is quite common and indicates other strata between the usual layers. Multiple critical frequencies do occur at times of rapid increase in ionization, especially at sunrise. This is to be expected because the rate of change of frequency is not great enough to keep up with the rapid increase in ionization. It is not likely, however, that this is the cause of the fine structure shown above. At 1130 there are only two critical frequencies between E and F_1 and at 1200 and after there is only one. Fine structure is in evidence on many of the other records of Fig. 2. It should be pointed out that many of the patterns recorded are very complex and that it is often difficult and sometimes impossible to interpret the results. With three or more layers, some of them giving double refraction, with reflections directly from, as well as back and forth between layers, added to multiple reflections and refractions, it can be understood that some of the patterns will be quite complex.

The results of the year's measurements of E-layer critical frequencies are shown in the graphs of Figs. 3, 4, and 5. The time scale is divided into half-hour intervals and the frequency scale is divided into intervals of 50 kilocycles. The points falling within a given interval are plotted within the rectangle corresponding to that interval. When there is only one critical frequency it is shown as a dot. When the critical frequencies are multiple (i.e. where fine structure exists) as shown in Fig. 2, the lowest one is plotted as a dot and the second one as a cross. The values higher than the second are not indicated.

[5] J. P. Schafer and W. M. Goodall, *Nature*, vol. 131, p. 804; June 3, (1933); E. V. Appleton, *Nature*, vol. 131, p. 872; June 17, (1933); J. A. Ratcliffe and E. L. C. White, *Nature*, vol. 131, p. 873, June 17, (1933).

The curves represent hourly averages of the lowest critical frequencies. The points representing measurements on the half hour are not included in the average curves because they are relatively few in num-

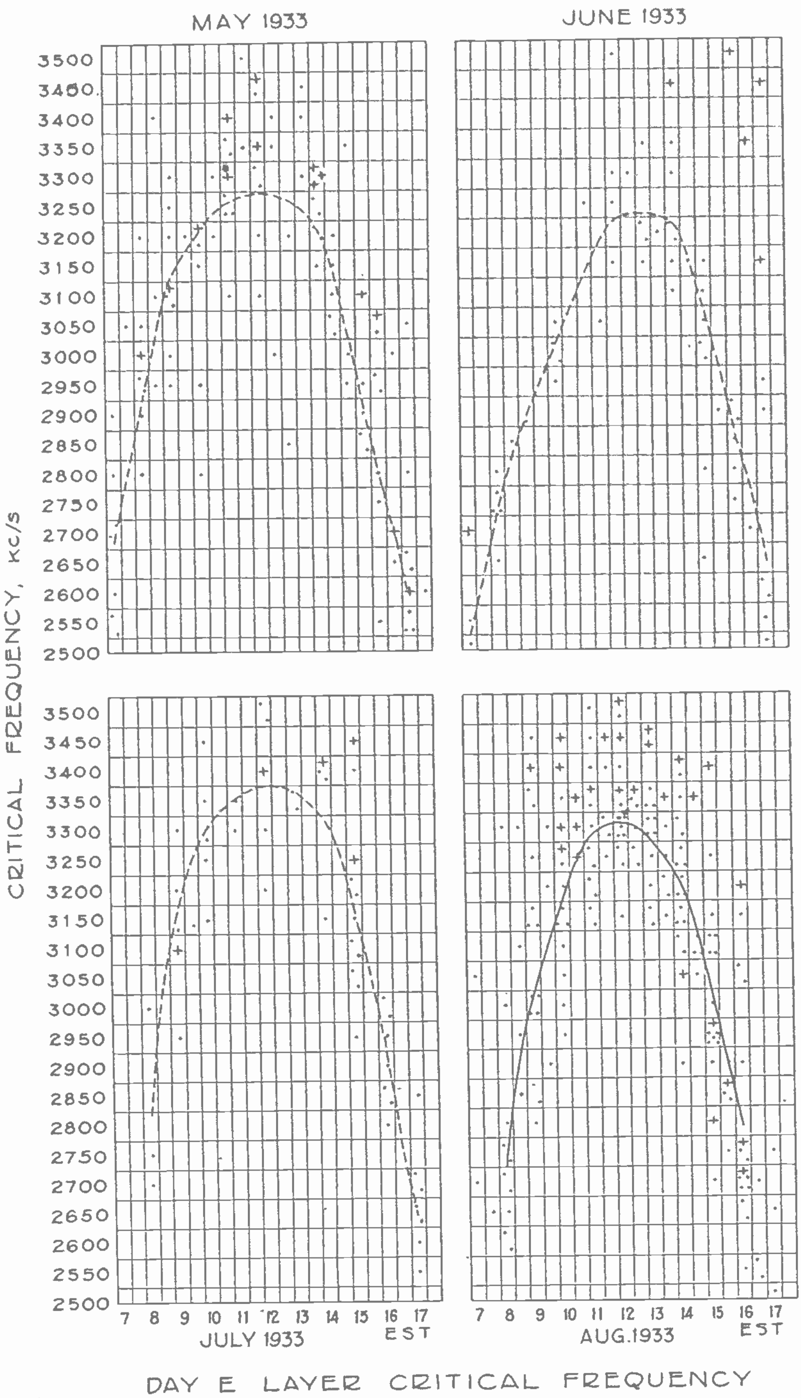

Fig. 3—Day E-layer critical frequencies for May, June, July, and August, 1933. Curves represent hourly averages of lowest critical frequencies.

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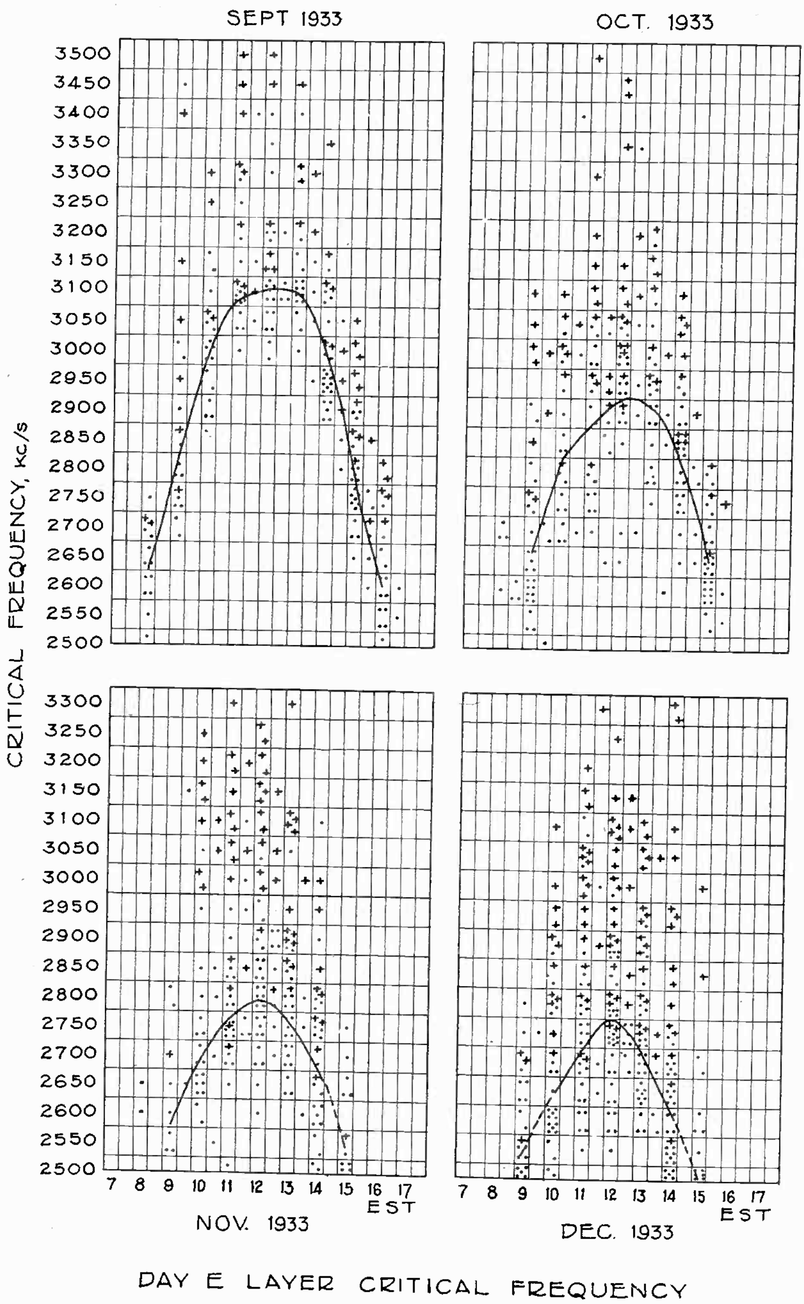


Fig. 4—Day E-layer critical frequencies for September, October, November, and December, 1933. Curves represent hourly averages of lowest critical frequencies.

are dotted because the measurements were relatively few in number during these months and the average value shown may not represent an accurate average for the month. It is likely that the maximum of the average curves for May and June should be higher. Most of the

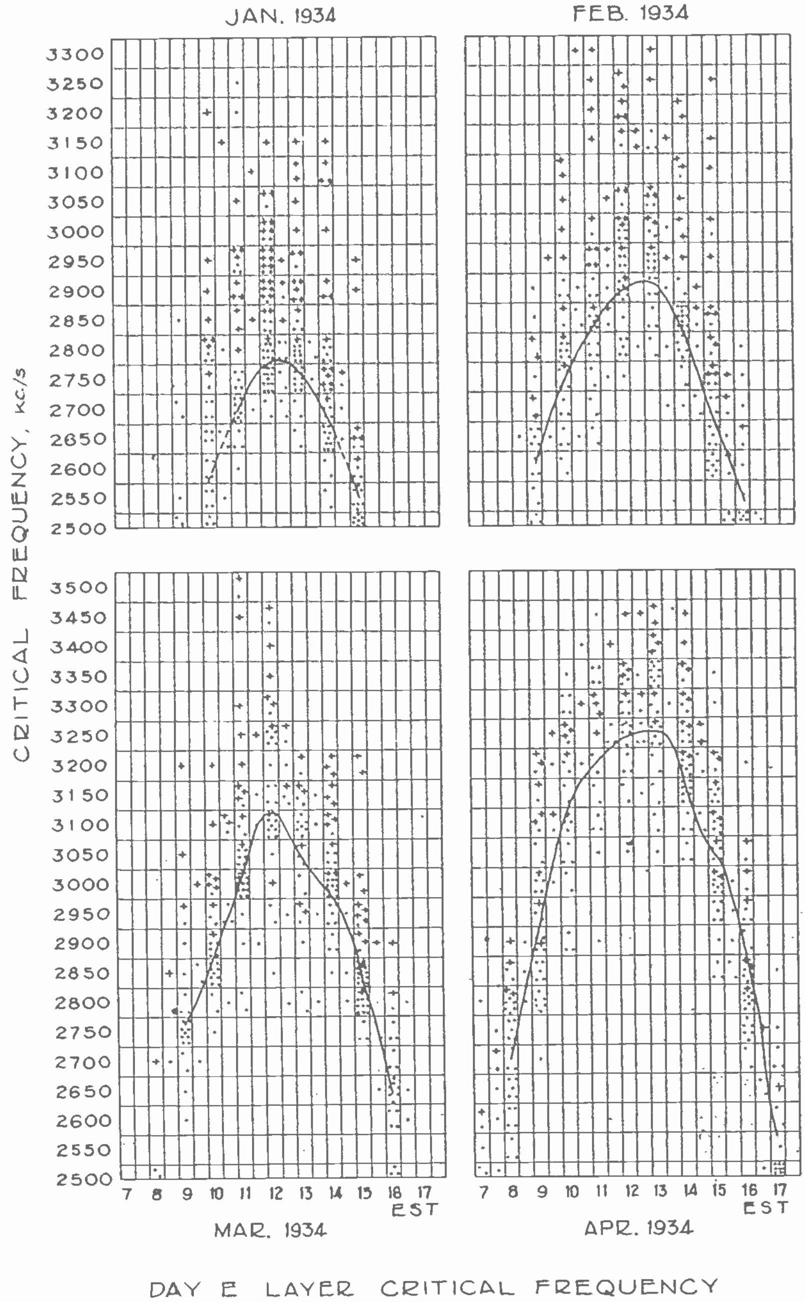
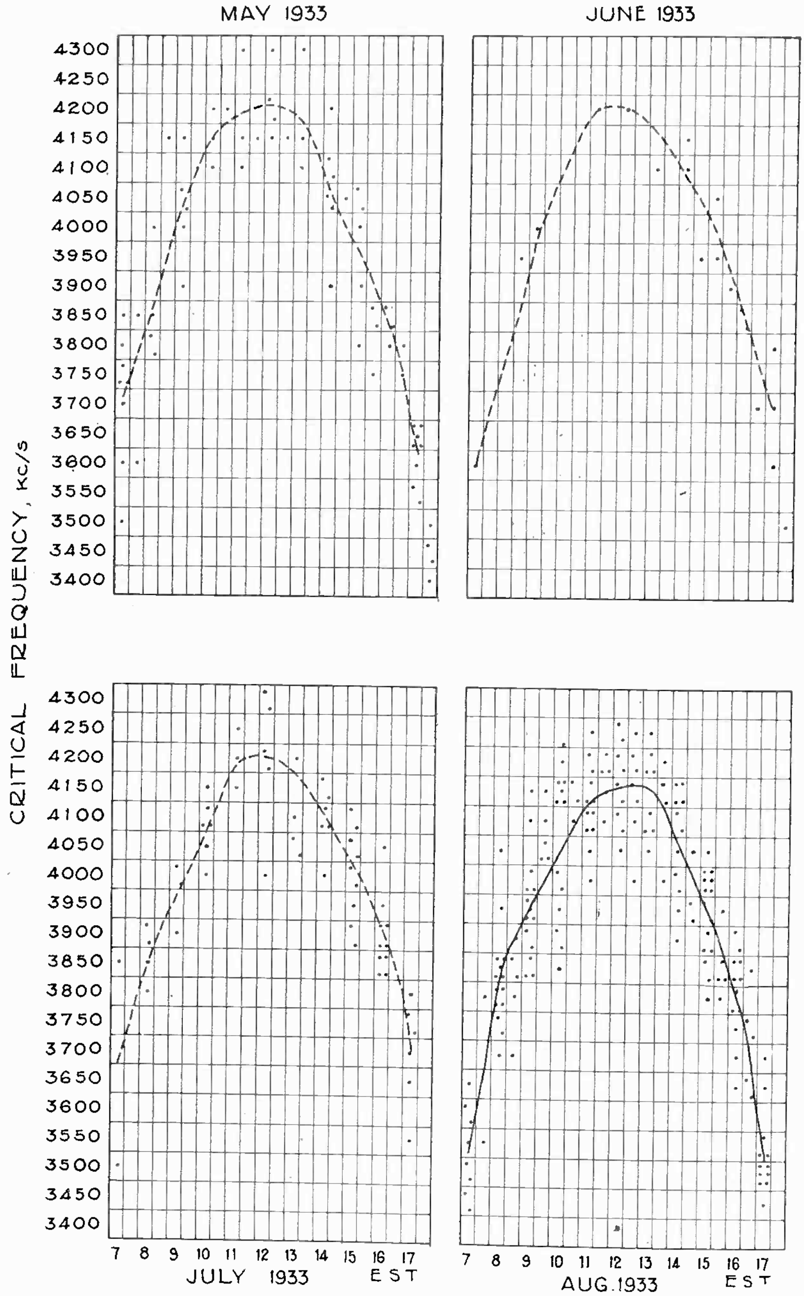


Fig. 5—Day E-layer critical frequencies for January, February, March, and April, 1934. Curves represent hourly averages of lowest critical frequencies.

curves show some irregularity, which can be understood, considering the wide scatter of the points. Some of the scatter may be caused by the fact that one of the critical frequencies of a multiple group may disappear between records and it is usually impossible to tell which one is missing. As a result the lowest one of the group which appears in the

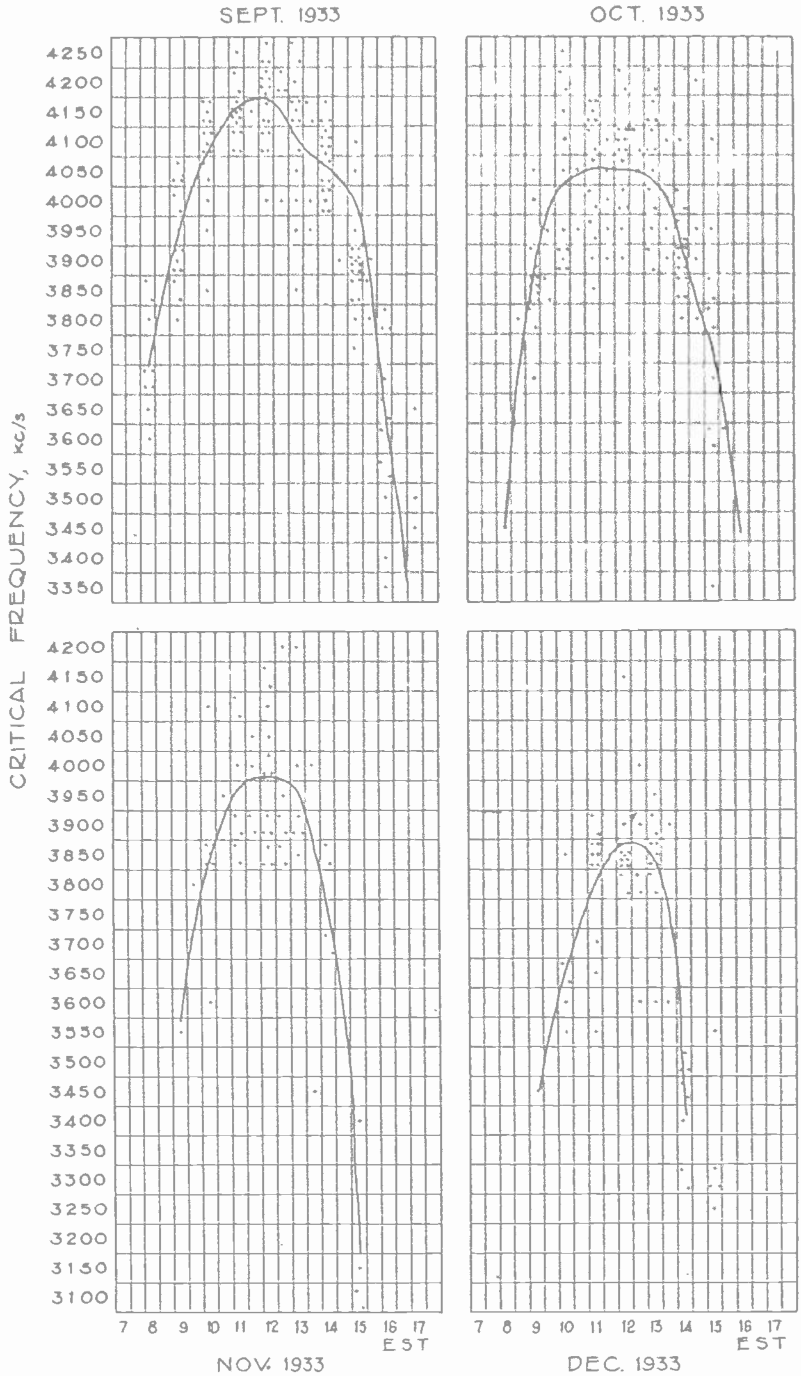


DAY F_1 LAYER CRITICAL FREQUENCY

Fig. 6—F_1-layer critical frequencies for May, June, July, and August, 1933. Curves represent hourly averages.

average curve may actually represent the critical frequency for a higher stratum. The maximum of the average critical frequency curve for each month is shown in Table I. In each case the maximum comes very near to noon.

The results of critical frequency measurements of the ordinary ray



DAY F_1 LAYER CRITICAL FREQUENCY

Fig. 7—F_1-layer critical frequencies for September, October, November, and December, 1933. Curves represent hourly averages.

in the F_1 layer for the year are shown in Figs. 6, 7, and 8. The curves represent averages of measurements made on the hour only. Here, as for the E-layer results, the points are plotted in rectangles representing

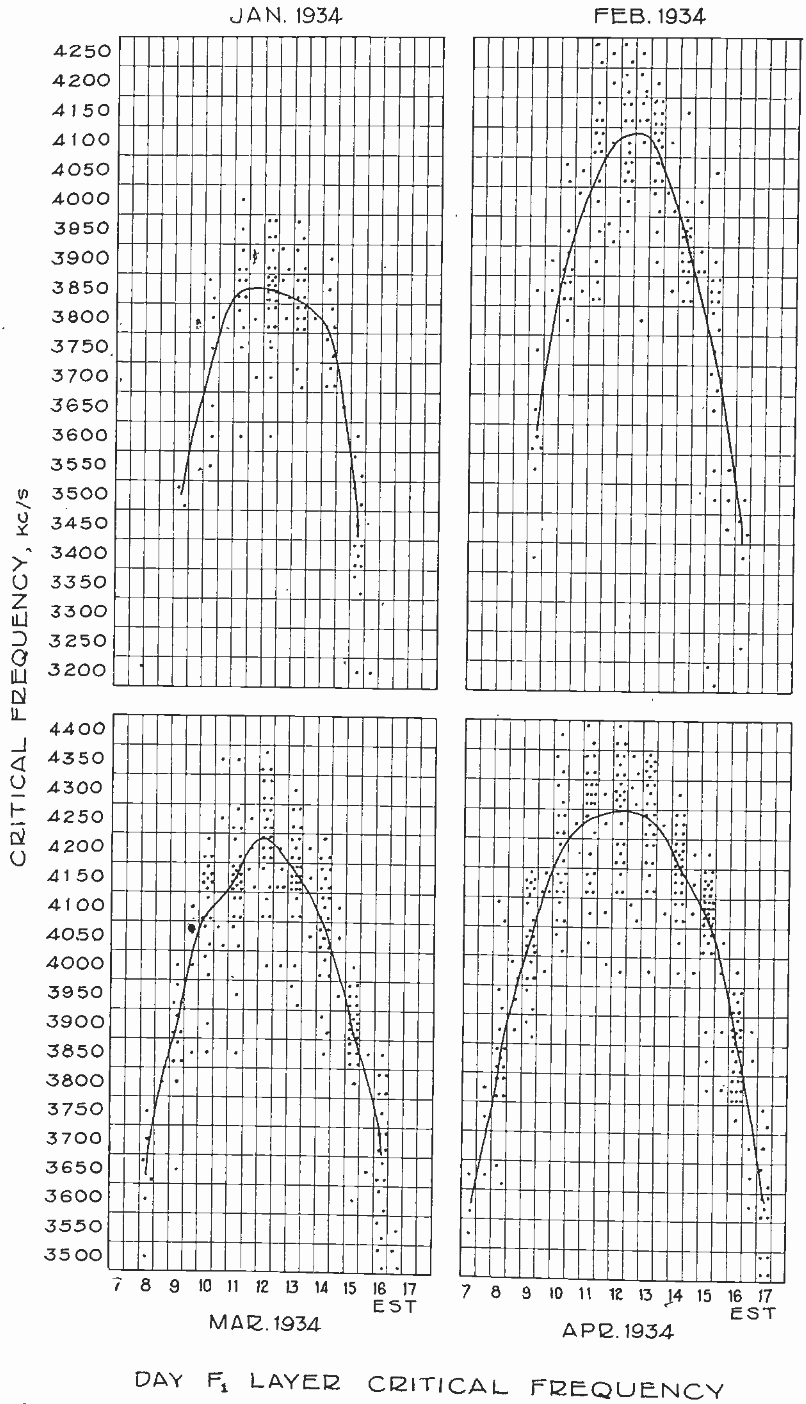


Fig. 8—F_1-layer critical frequencies for January, February, March, and April, 1934. Curves represent hourly averages.

intervals of one-half hour and 50 kilocycles. Stratification does not appear between the F_1 and F_2 layers so that only one critical frequency is in evidence. During most of the day the extraordinary ray critical frequency is above the highest frequency recorded by the present system. As mentioned above, the F-layer stratification appears only during the daytime, and is much more pronounced in summer than in winter. The maximum of the average curve for each month is shown in Table I. As for the E layer, the maximum comes very near to noon.

TABLE I

MAXIMA OF AVERAGE CRITICAL FREQUENCY CURVES

For Day E Layer			For Day F Layer		
Year	Month	Maximum of Average Critical Frequencies	Year	Month	Maximum of Average Critical Frequencies
		kc			kc
1933	May	3270(?)	1933	May	4205(?)
1933	June	3230(?)	1933	June	4205(?)
1933	July	3375(?)	1933	July	4205(?)
1933	August	3310	1933	August	4160
1933	September	3105	1933	September	4175
1933	October	2925	1933	October	4050
1933	November	2795	1933	November	3980
1933	December	2775	1933	December	3870
1934	January	2780	1934	January	3850
1934	February	2910	1934	February	4115
1934	March	3120	1934	March	4215
1934	April	3255	1934	April	4275

III. Night F-Layer Results

As mentioned before, the frequencies within the band covered by the present system are usually returned from the F layer at night. Fig. 9 shows the type of record obtained for six consecutive nights in October, 1933. During the first night the critical frequency for the extraordinary ray fell below 2500 kilocycles at 2000 E.S.T., increased to a maximum of 3920 kilocycles at 0400 and then fell to a minimum of about 3750 kilocycles at 0600. Both rays are in evidence from 0200 to 0600, inclusive. The ordinary ray is at the left and the extraordinary at the right. Although accurate determination of the actual separation between the two critical frequencies will await a more nearly simultaneous determination of the two values, the separation is very nearly 800 kilocycles when the ordinary ray critical frequency is at 2500 kilocycles. Theory shows that with this separation the ions here are undoubtedly electrons. The changes during the second night are quite different. A minimum critical frequency of 3450 kilocycles occurred at 2100 followed by a maximum of 4100 kilocycles at 0000 and another minimum below 2500 kilocycles at 0500. The critical frequency increased also during the other four nights but it did not fall below 2500 kilocycles unless possibly during the third night when F refractions were obscured by strong sporadic E reflections at 2300 and 0000.

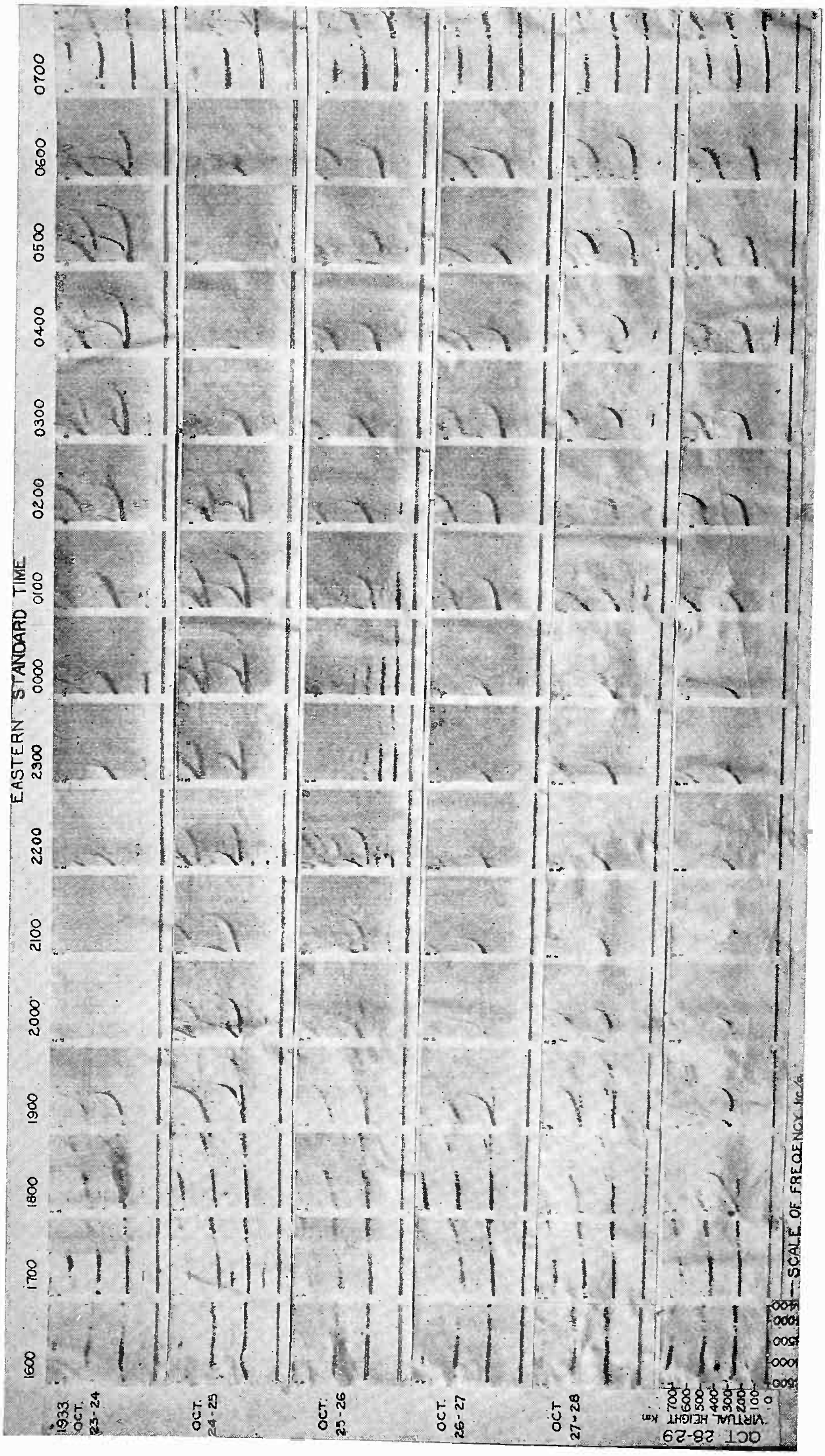


Fig. 9—Showing records obtained for six consecutive nights in October, 1933. Note E-layer reflections at 2300 and 0000 E.S.T. during third night.

The greater portion of the critical frequencies obtained at night in this band were for the extraordinary ray. The curves of Figs. 10, 11, and 12 have been plotted from hourly averages of this critical fre-

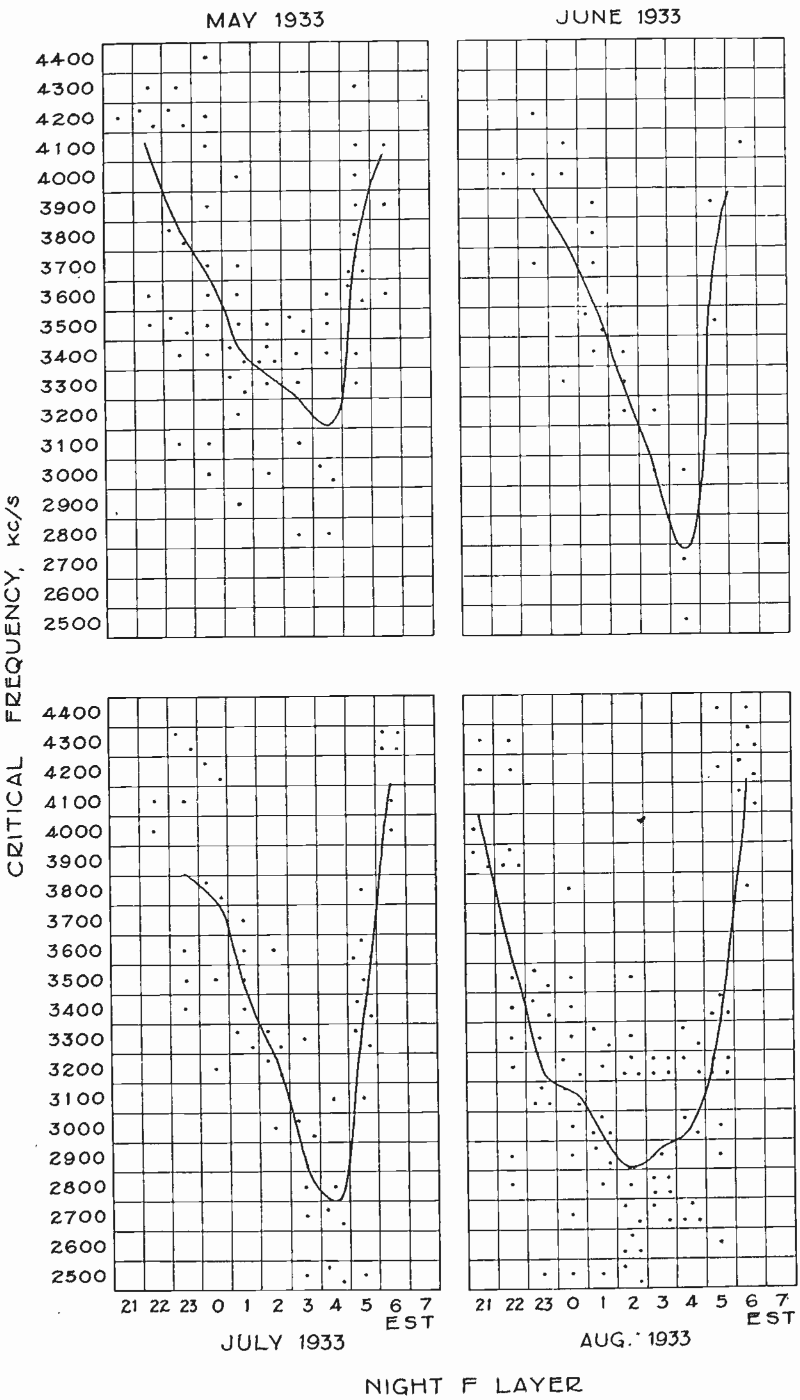


Fig. 10—Night F-layer critical frequencies with hourly average curves for May, June, July, and August, 1933.

quency for each month. The points are plotted in squares representing intervals of one hour and 100 kilocycles. The character of the average curves for the winter months is of particular interest. This is especially

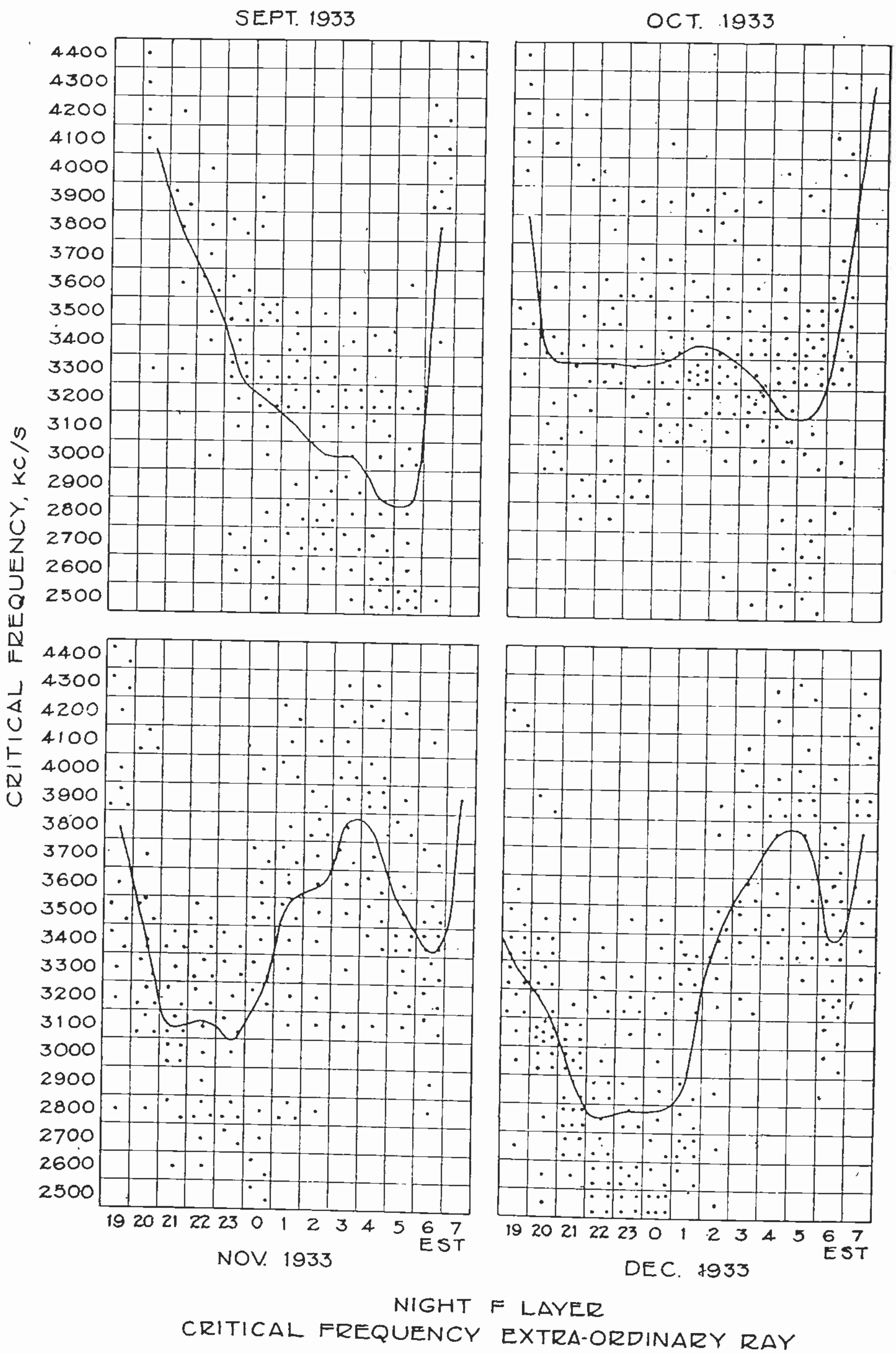
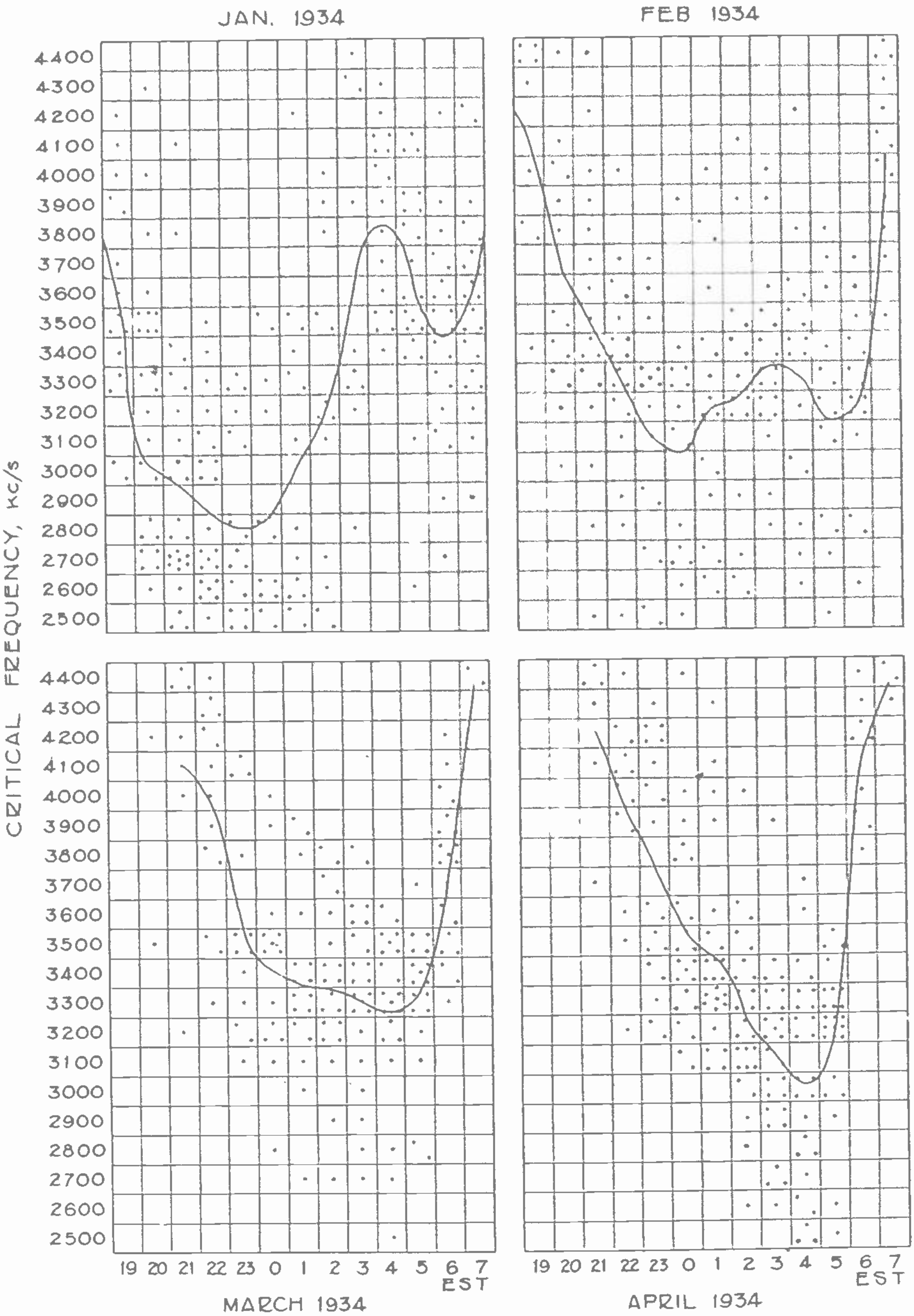


Fig. 11—Night F-layer critical frequencies with hourly average curves for September, October, November, and December, 1933.

true for November, December, and January. It will be noted that on the average during these months there is a definite increase in density of ionization during the night followed by a decrease before sunrise. Table II shows the maximum and minimum values of the average



NIGHT F LAYER
CRITICAL FREQUENCY, EXTRA-ORDINARY RAY.

Fig. 12—Night F-layer critical frequencies with hourly average curves for January, February, March, and April, 1934.

TABLE II

MAXIMA AND MINIMA OF AVERAGE NIGHT CRITICAL FREQUENCY CURVES FOR THE F LAYER

Month	1st Minimum, kc	Time of Occurrence E.S.T.	Maximum, kc	Time of Occurrence E.S.T.	2nd Minimum, kc	Time of Occurrence E.S.T.
May, 1933	3160(?)	0400				
June, 1933	2735(?)	0400				
July, 1933	2750(?)	0400				
August, 1933	2850	0200				
September, 1933	2825	0430				
October, 1933	3330	2300	3400	0100	3150	0430
November, 1933	3045	2300	3830	0320	3370	0600
December, 1933	2800	2200	3820	0430	3430	0600
January, 1934	2800	2320	3820	0400	3440	0600
February, 1934	3045	0000	3335	0320	3155	0510
March, 1934	3265	0410				
April, 1934	3010	0400				

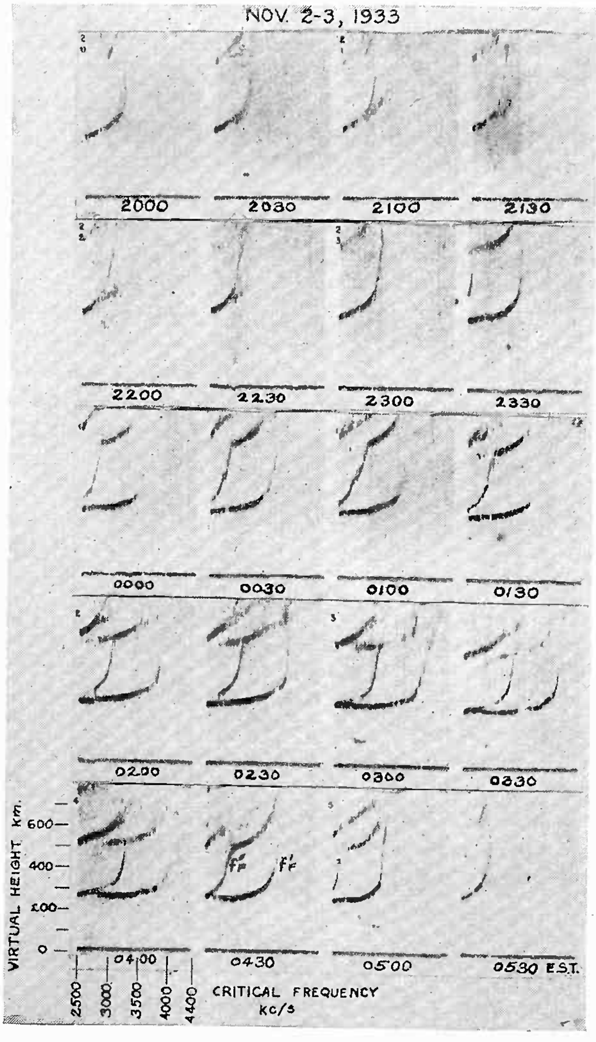


Fig. 13—Records of November 2 and 3, 1933, showing increase of F-layer critical frequency during the night. Between 2230 and 0330 the critical frequency for the extraordinary ray increases from 2950 to 4070 kilocycles representing more than a 100 per cent increase in maximum electron density. f_F'' represents ordinary ray critical frequency. f_F' represents extraordinary ray critical frequency.

curves for each month with the time of occurrence. Calculation shows that during December the maximum electron density more than doubles between 2200 E.S.T. and 0430 E.S.T. No explanation is offered to account for this increase in ionization during the night. It seems to be centered about midwinter. Fig. 13 shows records taken at half-hour

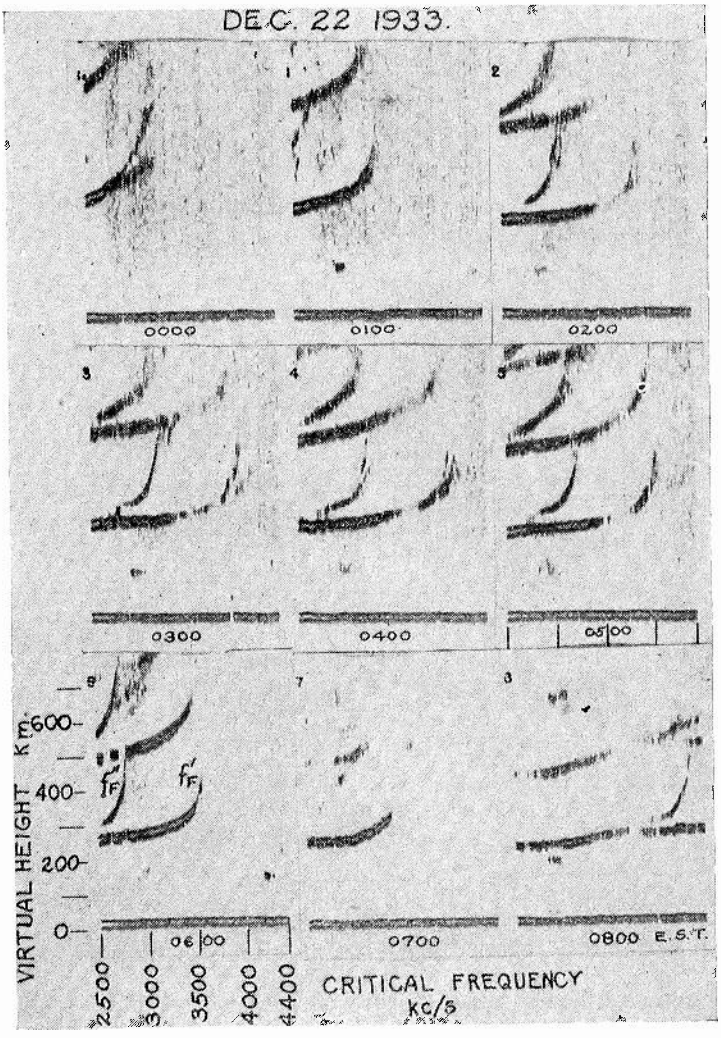


Fig. 14—Records showing increase of F-layer critical frequency during the night, December 22, 1933.

intervals during the night of November 2–3, 1933. The extraordinary critical frequency has a minimum value of 2950 kilocycles at 2230 and increases to a maximum of 4070 kilocycles at 0330 and then drops to another minimum of 3030 kilocycles at 0530. This represents more than a 100 per cent increase in maximum electron density in five hours. Fig. 14 shows similar changes during the early morning of December 22.

Fig. 15 shows some of the different types of records obtained at night. I of Fig. 15 shows sporadic E reflections at 2000 and 2100 E.S.T. E-F reflections together with F ordinary and extraordinary

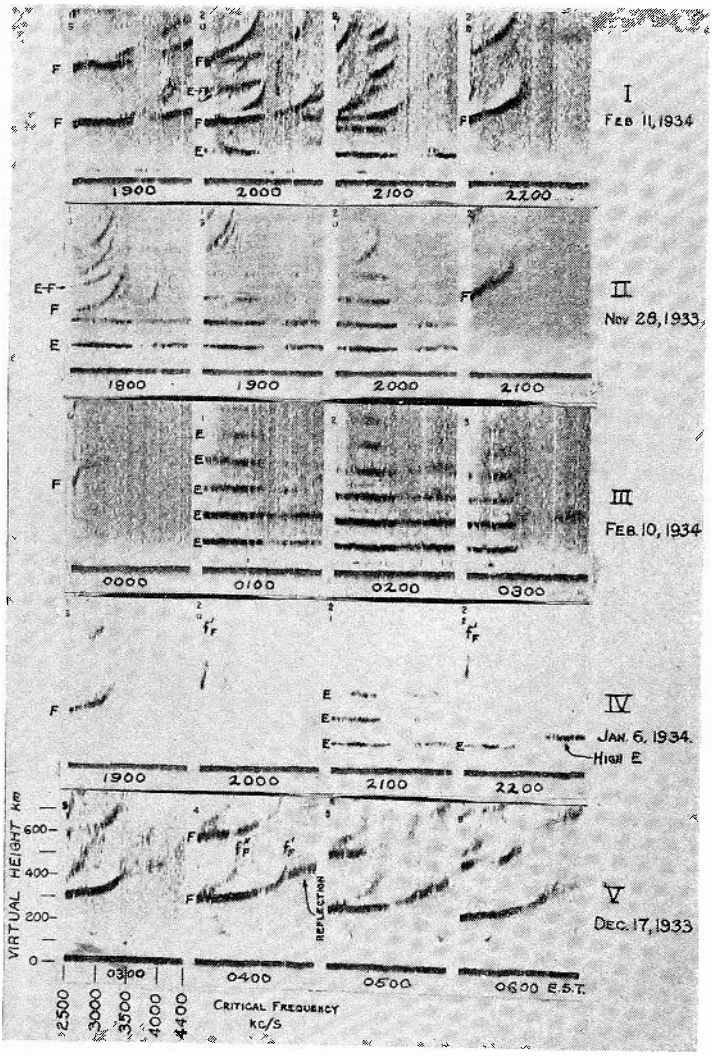


Fig. 15—Showing different types of records obtained at night. I shows sporadic E reflections at 2000 and 2100 E.S.T. E-F reflections together with F ordinary and extraordinary rays are also visible. II shows strong E reflections with E-F and F at 1800 and 1900. E is strong with F weak at 2000. Only F extraordinary ray is visible at 2100 with critical frequency about 3200 kilocycles. III shows F extraordinary ray at 0000. Strong E reflections with 5 multiples appear at 0100 and continue through 0200 and 0300. IV shows F extraordinary ray at 1900 and 2000. E and 2 multiples show at 2100. At 2200 F extraordinary ray is visible with E reflections. At this time E layer appears to be stratified which is unusual for these conditions. From 2500 to 3400 kilocycles the virtual height is 120 kilometers while from 3400 to 4400 the height is 170 kilometers. V shows reflections at frequencies above the critical frequency for refraction for the F layer. Thus at 0400 the extraordinary ray critical frequency is 3950 kilocycles while the reflection continues beyond 4400 kilocycles.

rays are also visible. In II strong E reflections with E-F and F are visible at 1800 and 1900. E is strong with F weak at 2000 E.S.T. Only the F extraordinary ray is visible at 2100 with critical frequency about 3200 kilocycles. In III of Fig. 15 only the F extraordinary ray is visible at 0000. Strong E reflections with five multiples appear at 0100 and continue through 0200 and 0300. In IV F extraordinary shows at 1900 and 2000. E and two multiples show at 2100. At 2200 F extraordinary is visible with E reflection. At this time the E layer appears to be stratified, which is unusual for these conditions. From 2500 to 3400 kilocycles the virtual height is 120 kilometers, while from 3400 to 4400 the height is 170 kilometers. In V reflections at frequencies above the critical frequency for refraction are shown for the F layer. Thus at 0400 the extraordinary ray critical frequency is 3950 kilocycles while the reflection continues beyond 4400 kilocycles.

IV. Application of Results to a Practical Communication Problem

Recently one of the air transport companies experienced difficulty in communicating at night between ground stations and aircraft along one of the routes in northeastern United States using a frequency of 3257.5 kilocycles. Conditions were reported to be especially bad during September, 1934. Since the frequency used is within the band covered by the measurements described above, the data were examined to determine what behavior would be expected at this frequency. The data for the period September 11–30, 1933, were compared with those for the same period of 1934, and are shown in Table III. The figures without letters in this table indicate extraordinary critical frequencies for the F layer in kilocycles. The discussion is in terms of the extraordinary ray since the problem is concerned with the limiting frequency which will return to earth. As mentioned before the ionization which gives sporadic reflections from the E layer will, at times, support transmission at frequencies considerably above the critical value for the refracted ray. When sporadic reflections are returned from the E layer the highest frequency returned is indicated in the table by the letter *E*. The reflections occur only a few times during September and are relatively unimportant during the period included in the table. Reflections from the F layer are indicated by the letter *R* after the highest frequency reflected. Where both reflection and refraction occur in the F layer two numbers are given for the hour, the critical frequency for the refracted ray being given alone while the highest reflected frequency is followed by the letter *R*. These reflections are also relatively unimportant compared to the refractions. When they do appear the maximum frequency returned is usually not much higher than that returned by refraction. Records were obtained during 118 of the 140

TABLE III

Maximum Frequencies Returned from the Ionosphere at Normal Incidence

Night Sept.	Hour, E.S.T.						
	2300	0000	0100	0200	0300	0400	0500
			September 11–30 inclusive, 1933				
11–12	3220	3150	3200*R*	3100*R*	2600	*X*	*X*
12–13	3250	3250	3100*R*	3100*R*		2900*R*	3000*R*
13–14		*X*	2700	2770	3150	3150	2600
14–15	2700	2550		2700	2680	2550	2530
15–16	2600	2550		2710 3100*R*	2700 3100*R*	2620	*X*
16–17	3250	3230	2800 3200*R*	2770	2770	2700	2720
17–18		3500	3340	2930	*X*	*X*	*E*3900
18–19	2730		2710	2740		2610	
19–20	3450	3230	3230	3210	2980	2700	2800
20–21	3330	3170	3250	3130	3290	*E*2900 2770	*E*4400
21–22	3270	3100	2910	2770	2900		
22–23							
23–24	3350	3500	3460	*E*3400 3200	*E*4400 3170	*E*3400 3350	3220
24–25	3630	3500	3430	3310	3150	3180	3010
25–26		3530	3220	2570	*E*4000	*E*4400	*E*3300 3180
26–27		2870	2570	2730		2970	3200
27–28	3770	3510 2750*R*	3400 3800*R*	3310	3230	3150	
28–29	3200	3280	3180	3250	3140	3090	3000
29–30	3080	2950	2790	2910	2910	3050	2970
30–Oct. 1	3470	3100	3150	3200	3200	3270	
Average	3220	3135	3080	2955	2950	2875	2840
			September 11–30 inclusive, 1934				
11–12		3900	3350	2950	2770	2700	2760
12–13	4200	4100	3900	3670	3170		3000
13–14	4250	3950	3750	3500 3900*R*	3550 3800*R*	3400 3800*R*	3350 3800*R*
14–15	4500	4000	3550	3400	3330	3180	3200 3800*R*
15–16	4060	3820	3600	3510	3280	*E*3500 3260	3060
16–17	4180	4050	3670	3370	3000	2550	*X*
17–18	3650	3700*R*	3680*R*	3450*R*	3500*R*	3600*R*	3900*R*
18–19	3630	3530	3650	3720	3570	3600	3350
19–20	4000	3900	3600	3330	2850	2750	3100
20–21	3800	3500	3200	3200	3000	3020	2940
21–22	4500	4270	3650	3240	2900	2760	*E*2650 2770
22–23	3300	3050	2770	2760	2720	2730	*E*weak 2800
23–24	3750	3600	3650	3330	2950 3400*R*	*X*	3150
24–25	4300	4070	3870	3350	3300	2700	2500 3100*R*
25–26	*E*3500	*E*3200	2800 3400*R*	2870 3500*R*	3300	*E*3440 2950	3250
26–27	3650	3500	3650	3830	3530	3400	3320
27–28	3150	3130	3160	3130	3180	3110	*E*3700
28–29	3500	3380	3280	3250	3300	3400	3450
29–30	3600	3730	3650	3600	3470	3580	3530
30–Oct. 1	3130	3000	3120	*E*3000 3160	*E*2950 3160	3150	3200
Average	3840	3690	3470	3320	3175	3040	3060

hours of this period in 1933 and during 138 of the 140 hours for the corresponding period of 1934. *X* indicates that no signals are returned at frequencies above 2500 kilocycles.

The curves of Fig. 16 are plotted from the hourly averages of the F layer extraordinary ray critical frequency and may be used together with the lower curve of Fig. 17 to determine, on the average, the highest frequency which would be expected to be useful for transmission for any short distance between transmitting and receiving stations.

For example, the average critical frequency at 4:00 a.m. during the period September 11–30, 1934, is 3040 kilocycles. Then from the lower

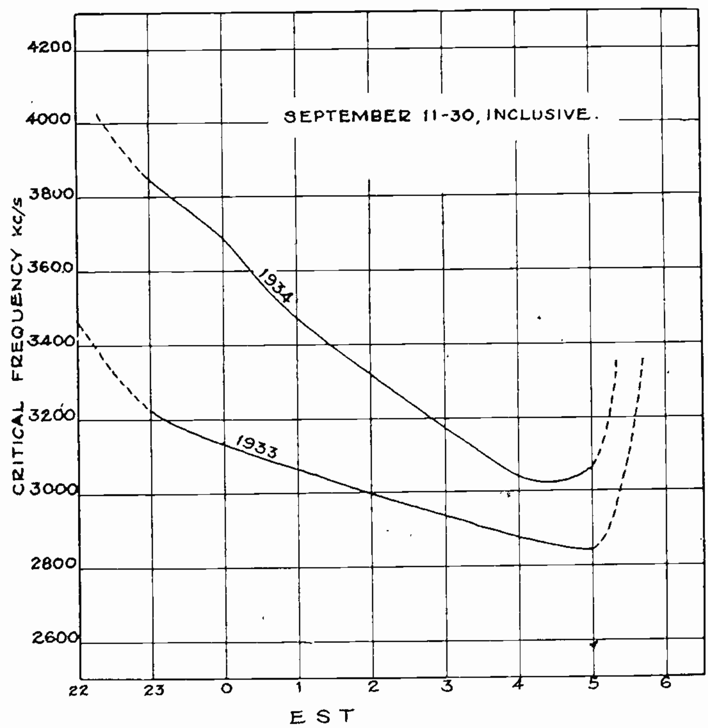


Fig. 16—Curves showing hourly averages of night F-layer extraordinary ray critical frequencies for periods September 11–30, 1933, and for September 11–30, 1934. Note that minimum for 1934 is about 180 kilocycles higher than for 1933.

curve of Fig. 17 the factor for 100 miles, say, is found to be 1.06 and for 200 miles is 1.21. Then on the average during this period no sky-wave frequencies above 1.06×3040 or 3225 kilocycles would be received at distances nearer than 100 miles from the transmitter and no frequencies above 1.21×3040 or 3680 kilocycles would be received at 200 miles or nearer. The variability from night to night is quite large so that individual values of critical frequency are often much different from the average value. Table III shows that the critical frequencies from which the average value was taken for 4:00 a.m. range from 2700 to 3600 kilocycles. The curves of Fig. 17 may be used together with table III to determine the limiting frequency for any distance at any hour of this period. The factors given by the curves of Fig. 17 are simply secants of the angle of incidence at the layer.

It will be noted that the curve for September, 1933, shown in Fig. 16 is somewhat different from that shown in Fig. 11. This is due to the fact that the curve of Fig. 11 represents the whole month while the other curve represents only the last twenty days of the month.

It is interesting to note that the minimum of the average curve for 1934 is about 180 kilocycles higher than that for 1933. Since the trend of the sun-spot curve is upward it might be inferred that the critical frequencies will increase as the sun-spot cycle progresses upward but since these observations extend over such a short period of time it is not advisable to make any predictions at this time. The provisional sun-spot numbers total 57 for this period of September, 1933, and 106 for the same period of 1934.

Table IV is prepared from Table III and shows the percentages of the time that the critical frequency for the extraordinary ray falls below 2500, 2750, 3000, and 3250 kilocycles during the periods September 11–30, 1933, and 1934. The percentages of the time that 2750, 3000, and 3250 kilocycles were above the limiting frequency for a distance of 100 miles and, similarly, when 3000 and 3250 kilocycles were above the limiting frequency for 200 miles, are also given in Table IV.

TABLE IV

PERCENTAGES OF THE TIME DURING THE PERIODS SEPTEMBER 11–30, 1933 AND 1934, FOR WHICH LIMITING FREQUENCIES FELL BELOW SPECIFIED VALUES

(Data covers hours 2300–0500; 11:00 P.M.–5:00 A.M. E.S.T.)

f = frequency specified	Critical frequency falls below f		f exceeds limiting frequency			
			for 100 miles		for 200 miles	
	1933	1934	1933	1934	1933	1934
(kc)	(per cent)	(per cent)	(per cent)	(per cent)	(per cent)	(per cent)
2500	5.1	1.45	—	—	—	—
2750	25.4	5.8	10.2	2.9	—	—
3000	44.0	17.4	30.5	12.3	5.1	1.45
3250	74.5	39.2	46.5	22.4	15.3	2.9

It is likely that the satisfactory ground-wave range at this frequency is only thirty or thirty-five miles so that transmission is mainly by sky wave. The results of this study show that at times night transmissions over short distances at a frequency of 3257.5 kilocycles pass through the ionosphere and are lost from the earth. The results also indicate that a lower frequency such as 2750 kilocycles passes through the ionosphere at a given angle a much smaller percentage of the time. It would be necessary to go below 2500 kilocycles to obtain practically complete freedom from skipping. The other F-layer critical frequency graphs indicate that transmissions at a frequency of 3257.5 kilocycles will often pass through the ionosphere during any season. These graphs (Figs. 10, 11, and 12) together with the lower curve of Fig. 17, may be used to determine, on the average, the limiting frequency for any time of the year for any distance.

The communication circuit of the air transport company was not in operation during September, 1933, but the results show that even more trouble would have been experienced then than in 1934.

In allocating frequencies for a given type of service a consideration of data of the type shown here should prove useful. World-wide in-

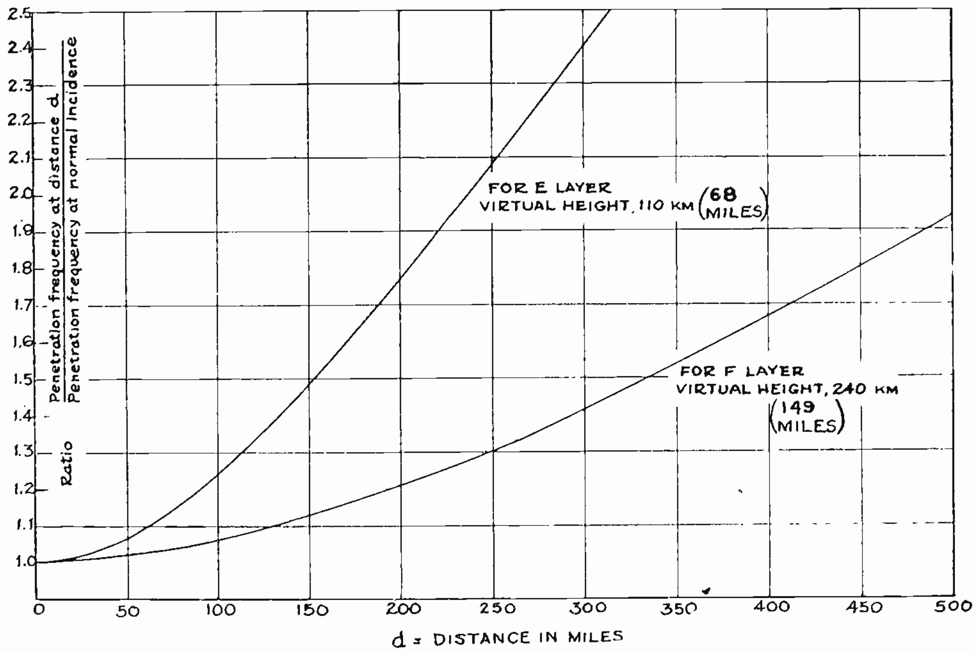


Fig. 17—Curves giving approximate multiplying factors for determining limiting frequency of penetration for any distance between transmitting and receiving stations. These factors multiply the critical frequencies for normal incidence.

formation will be necessary for an intelligent allocation of frequencies to be used in different geographical locations and for different types of service.

V. Conclusions

An attempt is made to picture the results of the year's critical frequency measurements in the solid diagram of Fig. 18. The hourly averages of the day F_1 critical frequencies above 3600 kilocycles are shown in the solid curves at the left. The curves below these represent averages of the day E critical frequencies above 2650 kilocycles. The solid curves at the right represent averages of night F critical frequencies below 3800 kilocycles. It is evident that the band used (2500–4400 kilocycles) gives an important and interesting part of the total cross section. The day E- and F_1-layer critical frequencies are seen to follow

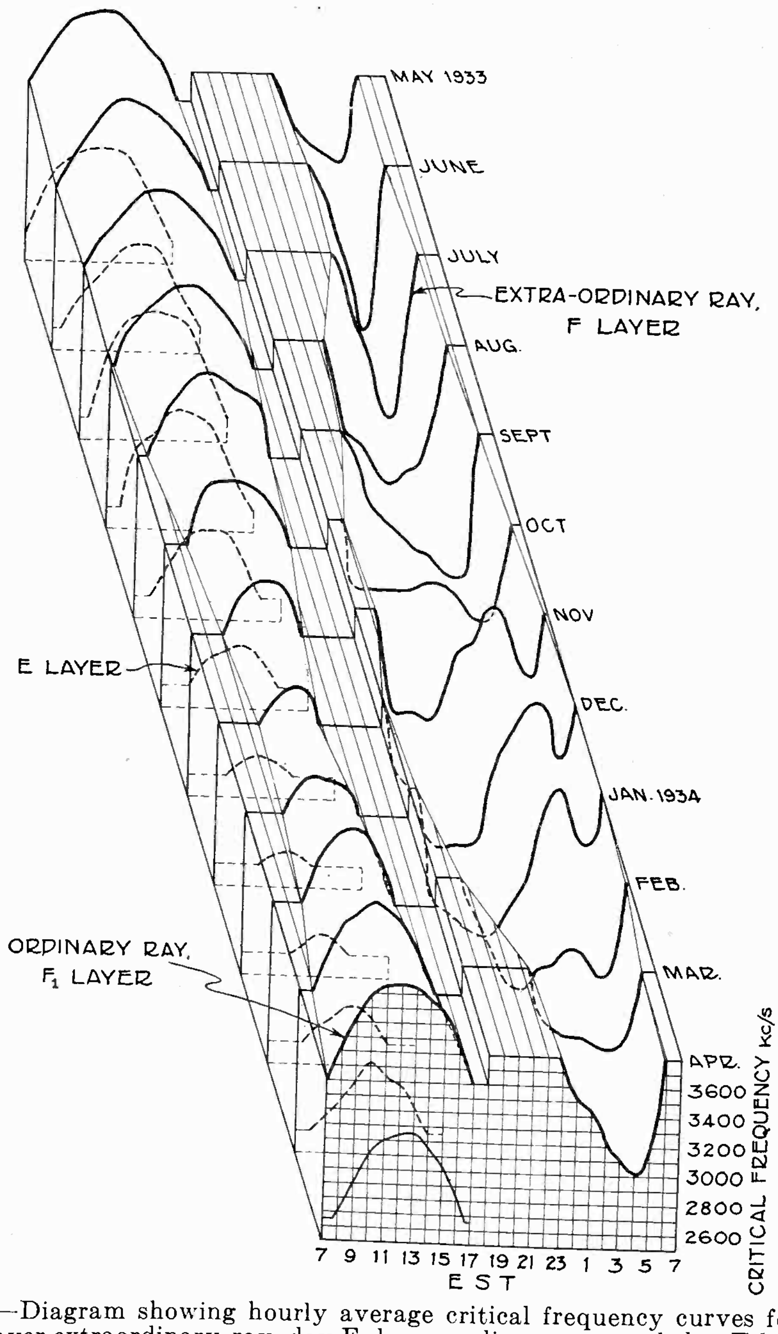


Fig. 18—Diagram showing hourly average critical frequency curves for night F-layer extraordinary ray, day F_1-layer ordinary ray, and day E-layer, for each month, May, 1933–April, 1934.

in phase with the sun both diurnally and seasonally. During the winter night the F-layer critical frequency drops until an hour or two before midnight, then increases until about 0400, after which it decreases again before sunrise. The maximum density of ionization frequently more than doubles after the first minimum.

A detailed analysis of the data has not yet been made with relation to magnetic storms. There have been no severe magnetic storms coincident with the hourly observations during the last one and a half years. Appleton has reported[6] that during magnetic storms in the northern Norway polar region reflections from the ionosphere were absent at all frequencies. No such pronounced effect has been observed in this latitude during the past one and a half years. It is expected that the severity of magnetic storms will increase with the advance of the sun-spot cycle during the next four or five years so that the relation between magnetic storms and changes in the ionosphere will be more easily recognized.

The probable effect of solar disturbances likewise will be demonstrated only by the accumulation of data over a longer period of time. Although the average night F-layer critical frequency curve for September, 1934, is considerably higher than that for September, 1933, no predictions are made at this time that this indicates a definite trend upward in the future. Data accumulated during the next four or five years should show how much sunspots do effect critical frequencies.

Information of the type shown here should prove useful in the study of the properties of the upper atmosphere as well as in the study of radio transmission.

Although the results obtained give a considerable part of the whole cross section it is desirable to extend the present system so that all of the critical frequencies will be obtained for the twenty-four hours. When more complete information of this type is available for different parts of the world and when the results are compared with actual transmission data a more complete understanding of sky-wave transmission should follow.

[6] E. V. Appleton, R. Naismith, and G. Builder, *Nature*, vol. 132, p. 340· September 2, (1933).

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DISSIPATION IN PHASE-COMPENSATING NETWORKS*

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Summary—*The effects of dissipation in the lattice-type, phase-compensating network are considered. There are three effects: (1) the phase shift is slightly different from the ideal calculated value; (2) there is introduced an attenuation; (3) the image impedance varies, especially near the frequencies of resonance and antiresonance of the lattice arms. The third effect is most important as it causes highly inconvenient reflections. Two methods are described for avoiding this effect, one of the methods having the special advantage that it avoids effect (1) at the same time.*

Introduction

A SINUSOIDAL wave, $E\cos(\omega t+\theta)$ suffers attenuation and phase shift during transmission, so that the received wave is $Ee^{-\beta}\cos[\omega t+\theta-\alpha_0]$. $e^{-\beta}$ is the diminution factor, β the attenuation in nepers, and α_0 the phase shift. Whatever β and α_0 are, the received wave is a replica of the sent wave, since both are sinusoidal.

If the sent wave is not sinusoidal we can express it as a Fourier series or integral with component or partial waves of the form given above. These partial waves have attenuation β and phase shift α_0, which are in general functions of ω. In order that the received wave be a replica of the sent wave it is necessary that β be independent of ω and α be of the form ωt_0 plus a multiple of π; for telephony it is sufficient that β be independent of ω and $\alpha_0=\omega t_0+$constant. The procedure adopted by transmission engineers[1] is to equalize the attenuation at all frequencies by means of "equalizing networks" and then to insert phase-compensating networks, which introduce very small attenuation (ideally no attenuation) but a phase shift α so that $\alpha_0+\alpha$ is of the form ωt_0+constant.

Before discussing phase-shift networks, a few words will describe the effects of attenuation and phase shift. If β is greater at high frequencies than at low, transmitted speech sounds woolly; if β is greater at low frequencies the speech sounds tinny. It was found sufficient for reasonable lengths of transmission line to equalize β; for great lengths it is found that this is not good enough, for if α_0 is not of the form ωt_0+constant, each syllable is preceded by a whistling noise. In the case of picture transmission the nonlinearity of the phase-shift curve

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[1] *See* Zobel, *Bell Sys. Tech. Jour.*, pp. 438–534, July, (1928).

causes a blurring effect; furthermore, for exact reproduction the constant added to ωt_0 must be a multiple[2] of π.

There are several kinds of networks, notably the lattice and bridged T, which can be designed to give zero attenuation (in the ideal case) at all frequencies and a desired phase shift. This paper discusses the departure of these networks from the ideal case caused by unavoidable resistance in the inductances that are used.

Lattice-Type, Phase-Compensating Network

Fig. 1 shows the well-known lattice network. If Z_1 and Z_2 are inverse networks, the lattice network is of the constant-resistance type; i.e., the image impedances are a constant resistance at all frequencies.

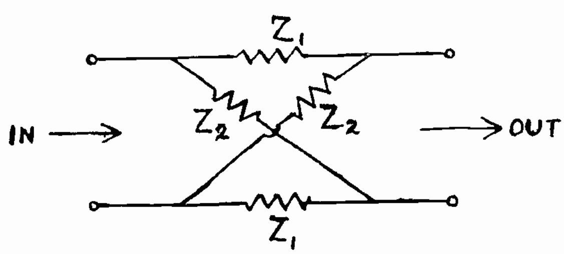

Fig. 1

Furthermore if the network consists wholly of reactive elements, there is no attenuation over the whole of the frequency range, so that the network is suitable for phase compensation. This facilitates design, for the system can be equalized for attenuation, paying no regard to phase, by the use of equalizing networks (described by Zobel[1]), and then the phase can be compensated without upsetting the conditions of equalized attenuation.

The lattice network[3] of Fig. 1 is symmetrical so that $Z_{i1} = Z_{i2} = Z_i$, say, the image impedance. The open-circuit impedance is

$$Z_0 = \tfrac{1}{2}(Z_1 + Z_2)$$

and the short-circuit

$$Z_s = 2Z_1Z_2/(Z_1 + Z_2),$$

so that

$$Z_1 = \sqrt{Z_0Z_s} = \sqrt{Z_1Z_2}$$

[2] *See* Zobel,[1] p. 501 *et seq.*, and Starr, "Electric Circuits and Wave Filters," Ch. XII.

[3] *See* K. S. Johnson, "Transmission Circuits," pp. 130–134; or Starr,[2] pp. 170–174.

and,

$$\tanh T = \sqrt{Z_s/Z_0} = 2\sqrt{Z_1 Z_2}/(Z_1 + Z_2),$$

giving,

$$\tanh (T/2) = \sqrt{Z_1/Z_2}. \tag{1}$$

Z_1 and Z_2 are made inverse to one another, so that $Z_1 Z_2 = R^2$. This is done by making an inductance L in Z_1 correspond to a capacity L/R^2 in Z_2, a resistance R_1 to a resistance R^2/R_1, and a capacity C to an inductance CR^2, a series combination to a parallel combination, and a parallel combination to a series combination. For example the impedances of Figs. 2(a) and 2(b) are inverse with product R^2.

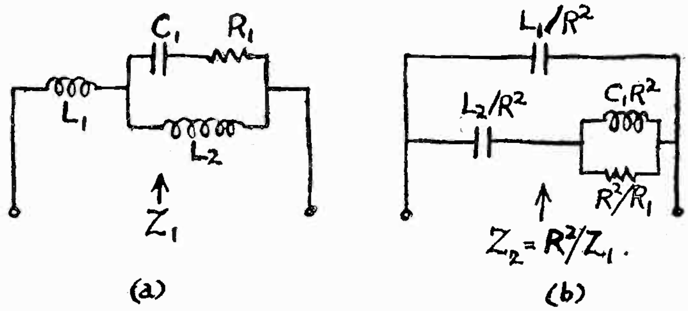

Fig. 2

This being so

$$\left.\begin{aligned} Z_i &= R \\ T &= 2 \tanh^{-1} (Z_1/R) \end{aligned}\right\}. \tag{2}$$

and,

If Z_1 is a purely reactive impedance, jX_1, say, Z_2 is purely reactive also and equal to $-jR^2/X_1$. Then,

$$\begin{aligned} T &= 2 \tanh^{-1} (jX_1/R) \\ &= j2 \tan^{-1} (X_1/R), \end{aligned}$$

so that,

$$\left.\begin{aligned} \beta &= 0 \\ \alpha &= 2 \tan^{-1} (X_1/R) \end{aligned}\right\} \tag{3}$$

and,

Type A Network

In this case $Z_1 = j\omega L$ and $Z_2 = 1/(j\omega L/R^2)$, the network shown in Fig. 3.

Then,

$$\left.\begin{aligned} \alpha &= 2\tan^{-1}(\omega L/R) \\ &= 2\tan^{-1}(\omega/\omega_0), \\ \omega_0 &= R/L = 1/\sqrt{LC}. \end{aligned}\right\} \tag{4}$$

where,

$$\frac{d\alpha}{d\omega} = \omega_0/(\omega_0^2 + \omega^2) \tag{5}$$

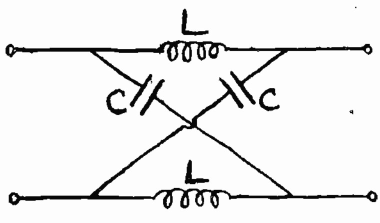

Fig. 3

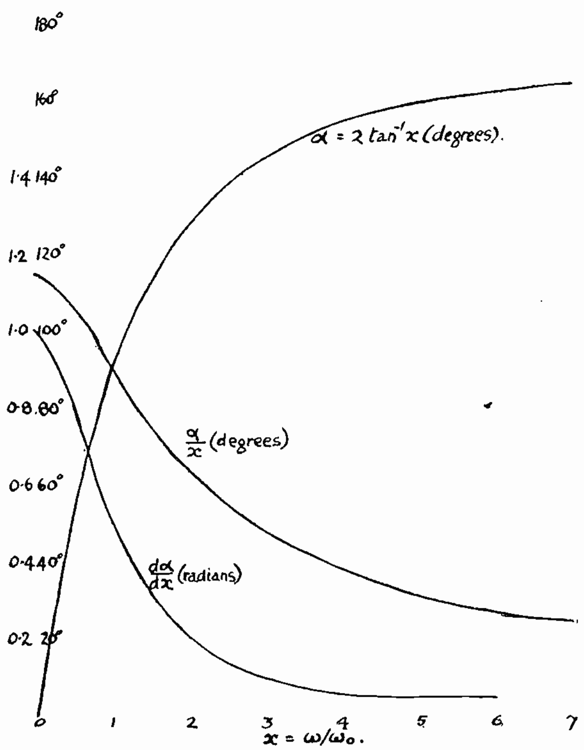

Fig. 4

and,

$$\frac{\alpha}{\omega} = \frac{2}{\omega}\tan^{-1}\left(\frac{\omega}{\omega_0}\right). \tag{6}$$

α, $\dfrac{d\alpha}{d\omega}$, and $\dfrac{\alpha}{\omega}$ are shown in Fig. 4.

Effect of Resistance in the Coils

In any actual case, however, the coils L will have some resistance, r, say. The leakage of the condensers is negligible in comparison. Let,

$$\omega L/r = Q, \tag{7}$$

the factor of the coil. Then,

$$\begin{aligned} Z_1 &= \sqrt{(j\omega L + r)/j\omega C} \\ &= R\sqrt{1 + 1/jQ} \\ &\doteq R\left(1 + \frac{1}{8Q^2}\right) + jR/2Q. \end{aligned}$$

Q will be of the order of 50 to 300, depending upon the frequency and the type of coil, so that $R/8Q^2$ may be neglected, but $jR/2Q$ introduces an irregularity which will produce undesired reflection effects. In addition the propagation constant will be modified. The effect can be illustrated simply by imagining the network of Fig. 3 to be terminated by a resistance R. Then the ratio of the emerging to the ingoing current is

$$\begin{aligned} &\frac{Z_1 - Z_2}{Z_1 + Z_2 + 2R} \\ &= \frac{j\omega L(1 + 1/jQ) - 1/j\omega C}{j\omega L(1 + 1/jQ) + 1/j\omega C + 2R}. \\ &= k\angle\overline{\phi}, \text{ say.} \end{aligned}$$

It will be found that[4]

$$\phi \doteq 2\tan^{-1}(\omega/\omega_0) + 2\omega^2 LC/Q(1 + \omega^2 LC)^2$$

[4] The numerator is approximately

$$j\omega L\left(1 + \frac{\omega_0^2}{\omega^2}\right) \angle\overline{1/Q\left(1 + \frac{\omega_0^2}{\omega^2}\right)}$$

and the denominator

$$j\omega L\left[1 - \frac{\omega_0^2}{\omega^2} - j\frac{2\omega_0}{\omega} - j\frac{1}{Q}\right]$$

$$\doteq j\omega L\sqrt{\left(1 - \frac{\omega_0^2}{\omega^2}\right) + \frac{4\omega_0^2}{\omega^2} + \frac{4\omega_0}{\omega Q}} \quad \angle\overline{\frac{\dfrac{2\omega_0}{\omega} + \dfrac{1}{Q}}{1 - \dfrac{\omega_0^2}{\omega^2}}}.$$

and,

$$k \doteq 1 - 2R/Q\omega L(1 + 1/\omega^2 LC)^2.$$

Thus the phase shift is increased by $2\omega^2 LC/Q(1+\omega^2 LC)^2$ while an attenuation is introduced of the amount

$$2R/[Q\omega L(1 + 1/\omega^2 LC)^2] \text{ nepers.}$$

In practice several such networks will be in tandem and so each network will not be correctly terminated. The actual phase shift obtained will be modified by reflections, and will be difficult to calculate.

Fortunately, this undesirable state of affairs can be avoided by allowing for the unavoidable dissipation in the coils, *after design*, by the following method.

Correction for Dissipation

The networks Z_1 and Z_2 are made truly inverse, after the correct values of L and C have been calculated, by placing resistances R^2/r across the condensers[5] C. If desired, leakage in the condensers G can be allowed for by making their resistances assume the value $1/(r/R^2-G)$, but this will be usually unnecessarily accurate because r will vary with frequency somewhat. This method will eliminate reflections, since the network now will be really a constant-resistance network, but in addition it will be shown that the phase shift of the network assumes the calculated value for a network *without* dissipation. This will help design considerably.

Proof will next be given that the employment of correcting resistances leads to ideal phase shift but introduces some attenuation.

Let Z_1 be an impedance composed of coils and condensers in which a typical coil has an impedance $z=j\omega L(1+1/jQ)$, and a typical condenser has an impedance $z=1/j\omega C(1+1/jQ)$. Z_2 is made inverse by placing resistance across the corresponding condensers to allow for dissipation in coils of Z_1 and resistances in series with coils corresponding to leakages in condensers in Z_1. The same is done for Z_1 to allow for losses in Z_2. Then,

$$Z_i = \sqrt{Z_1 Z_2} = R,$$

and,

$$T = 2 \tanh^{-1} (Z_1/R).$$

But,

$$Z_1 = f[j\omega L_1(1 + 1/jQ_1),\ 1/j\omega C_2(1 + 1/jQ_2),\ \cdots].$$

[5] British Patent, No. 342,307.

Let,

$$z_1 = j\omega L_1, \quad z_2 = \frac{1}{j\omega C_2}, \quad \text{etc.},$$

the impedance of the elements in the ideal case, and,

$$j\omega L_1\left(1 + \frac{1}{jQ_1}\right) = z_1 + \Delta z_1,$$

$$\frac{1}{j\omega C_2\left(1 + \dfrac{1}{jQ_2}\right)} = z_2 + \Delta z_2, \text{ etc.}$$

Then,

$$\begin{aligned} Z_1 &= f[z_1 + \Delta z_1, z_2 + \Delta z_2, \cdots] \\ &= f[z_1, z_2, \cdots] + \Delta z_1 \frac{\partial f}{\partial z_1} + \Delta z_2 \frac{\partial f}{\partial z_2}, \cdots, \end{aligned}$$

to the first order, where $f[z_1, z_2, \ldots]$ is the value of Z_1 in the ideal case and so with the derivatives.

$f(z_1, z_2, \ldots)$ is clearly a pure imaginary $= j\phi(\omega)R$, say, and the derivatives are real positive quantities. Also,

$$\Delta z_1 = j\omega L_1\left(1 + \frac{1}{jQ_1}\right) - j\omega L_1 = \frac{\omega L_1}{Q_1},$$

and,

$$\Delta z_2 = \frac{1}{j\omega C_2\left(1 + \dfrac{1}{jQ_2}\right)} - \frac{1}{j\omega C_2} = \frac{1}{\omega C_2 Q_2}, \tag{8}$$

to the first order, so that $\Delta z_1, \Delta z_2, \ldots$ are all positive real quantities to the first order. Then,

$$T = 2 \tanh^{-1}\left(\frac{Z_1}{R}\right) = j2 \tan^{-1}\left[\phi(\omega) + \frac{1}{jR} \Sigma \frac{\partial f}{\partial z} \Delta z\right].$$

Expanding by Taylor's theorem this gives

$$T = j2 \tan^{-1}[\phi(\omega)] + 2\left[\Sigma \frac{\partial f}{\partial z} \Delta z\right] \frac{1/R}{1 + \phi^2(\omega)},$$

giving a phase shift $2 \tan^{-1}[\phi(\omega)]$, which is the value of the ideal case, and an attenuation of

$$\frac{2/R}{1+\phi^2(\omega)} \cdot \Sigma \frac{\partial f}{\partial z} \Delta z \text{ nepers.} \tag{9}$$

This last summation is taken over all the elements in one arm of the lattice network, and Δz is given for a coil or condenser by (8).

Special Case. Corrected A Type Network

This is shown in Fig. 5.

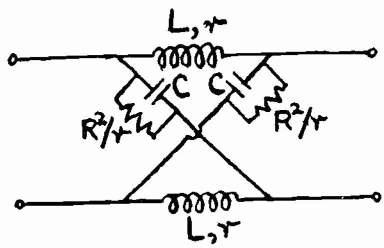

Fig. 5

The phase shift for this is the same as for the ideal network of Fig. 3; viz., $\alpha = 2 \tan^{-1}(\omega L/R)$.

In this case $z = j\omega L$, $f(z) = z$, and $\phi(\omega) = \omega L/R$. Equation (9) then gives the attenuation as

$$\frac{2/R}{1+\omega^2L^2/R^2} . 1 . \frac{\omega L}{Q} = \frac{2\omega/\omega_0}{Q(1+\omega^2/\omega_0^2)} \text{ nepers.}$$

Ideal Type B Network

This is shown in Fig. 6 for the ideal case.

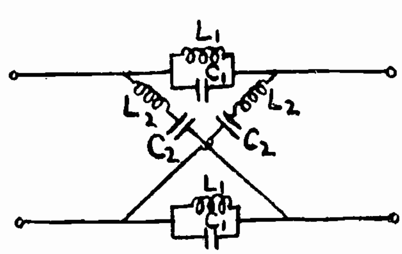

Fig. 6

In order that Z_1 and Z_2 be inverse

$$\frac{L_1}{C_2} = \frac{L_2}{C_1} = R^2.$$

$$Z_1 = j\omega L_1/(1 - \omega^2 L_1 C_1),$$

giving,

$$\begin{aligned} \alpha &= 2 \tan^{-1} \omega L_1/R(1 - \omega^2 L_1 C_1) \\ &= 2 \tan^{-1} [\gamma x/(1 - x^2)] \end{aligned} \tag{10}$$

where,

$$\left.\begin{aligned} x &= \omega/\omega_0, \\ \omega_0 &= 1/\sqrt{L_1 C_1} \\ \gamma &= \omega_0 L_1/R = \sqrt{L_1/L_2} \end{aligned}\right\}. \tag{11}$$

and,

Type B Network with Dissipation

Suppose the coils L_1 have resistance R_1 and L_2 resistance R_2. Let $\omega L_1/R_1 = Q_1$ and $\omega L_2/R_2 = Q_2$, then,

$$Z_1 = \frac{j\omega L_1\left(1 + \dfrac{1}{jQ_1}\right)}{1 - \omega^2 L_1 C_1\left(1 + \dfrac{1}{jQ_1}\right)},$$

$$Z_2 = j\omega L_2\left(1 + \frac{1}{jQ_2}\right) + \frac{1}{j\omega C_2}.$$

$$\therefore\ Z_i = \sqrt{Z_1 Z_2}$$

$$= \sqrt{\frac{L_1}{C_2}}\left[\frac{\left(1 + \dfrac{1}{jQ_1}\right)\left(1 - \omega^2 L_2 C_2 \cdot 1 + \dfrac{1}{jQ_2}\right)}{1 - \omega^2 L_1 C_1\left(1 + \dfrac{1}{jQ_1}\right)}\right]^{1/2}. \tag{12}$$

It is found in practice that the impedance of an A type network, even when uncorrected, is fairly constant over a reasonable range of frequency, although it has the reactive part $jR/2Q$. At very low frequencies, say down at 50 cycles, this does not hold, and the method described above is necessary.

The uncorrected B type, however, has an image impedance which varied rapidly in the neighborhood of $\omega_0 = 1/\sqrt{L_1 C_1} = 1/\sqrt{L_2 C_2}$. The variation may be thirty or fifty per cent of the normal impedance.

D. C. Espley (late of the International Telephone and Telegraph Laboratories) found experimentally that if the Q's of the coils are made equal the variation is smoothed out. It is easy to see that this is so, for if we put $Q_1 = Q_2$, and $L_1 C_1 = L_2 C_2$, equation (12) gives

$$Z_i = \sqrt{(L_1/C_2)(1 + 1/jQ_1)},$$

which is of the same form as the impedance for the A type network and is fairly constant over wide ranges of frequency except at very low frequencies.

Espley corrected the impedance of the B network by making the Q's of the coils equal by inserting a resistance in series with the coil which has the larger Q. By using a resistance box, he varied the inserted resistance so as to keep the Q's equal. This reduced the impedance of the B type network to that of the uncorrected A type network, which was sufficiently constant in his range of frequency.

This result is unexpected, for if we have Q's of values 80 and 50 say, the best we can do is to reduce both Q's to 50, and this gives less variation of image impedance than the Q's of values 80 and 50. So values of 50 and 50 are better than values of 80 and 50. Of course values of 80 and 80 will be better than either, but Q cannot be easily increased.

The reason why equalization of the Q's helps is because of the behavior of the factor

$$\left\{[1 - \omega^2 L_2 C_2(1 + 1/jQ_2)]/[1 - \omega^2 L_1 C_1(1 + 1/jQ_1)]\right\}^{1/2}$$

whose square is

$$\left[(1 - \omega^2 L_2 C_2) - \frac{\omega^2 L_2 C_2}{jQ_2}\right] \div \left[(1 - \omega^2 L_1 C_1) - \frac{\omega^2 L_1 C_1}{jQ_1}\right]$$
$$= \left[1 + \frac{j\omega^2 L_2 C_2}{Q_2(1 - \omega^2 L_2 C_2)}\right] \div \left[1 + \frac{j\omega^2 L_1 C_1}{Q_1(1 - \omega^2 L_1 C_1)}\right]. \qquad (13)$$

When,

$$\frac{\omega^2 L_1 C_1}{Q_1(1 - \omega^2 L_1 C_1)} < 1$$

i.e.,

$$\omega^2 L_1 C_1 < \frac{Q_1}{1 + Q_1},$$

expression (13) becomes approximately

$$1 + \frac{j\omega^2 L_1 C_1}{1 - \omega^2 L_1 C_1}\left(\frac{1}{Q_2} - \frac{1}{Q_1}\right). \qquad (14)$$

Let,

$$\omega^2 L_1 C_1 = \frac{KQ_1}{1 + Q_1},$$

where,

$$K < 1.$$

Then,

$$(14) = 1 + j\frac{KQ_1}{1 + Q_1 - kQ_1}\left(\frac{1}{Q_2} - \frac{1}{Q_1}\right).$$

If $Q_1 = mQ_2$, and Q_1 and Q_2 are both large, this becomes

$$1 + j\frac{K}{1 - K}\left(\frac{m - 1}{Q_1}\right). \tag{15}$$

If m is not equal to 1, the second term of this last expression can become large as K approaches the value 1. (K must not be taken too near to unity, for then the step from (13) to (14) will not be justified as an approximation.)

Hence, unless $Q_1 = Q_2$, (13) will vary appreciably near the region of $\omega^2 L_1C_1 = \omega^2 L_2C_2 = 1$.

A very simple way of seeing this is the following:

When $\omega^2 L_1C_1$ is very small, (13) is clearly nearly equal to unity. Similarly when $\omega^2 L_1C_1$ is very large,

$$(13) = \frac{1 - \omega^2 L_2C_2(1 + 1/jQ_2)}{1 - \omega^2 L_1C_1(1 + 1/jQ_1)}$$

$$= \frac{1 + 1/jQ_2}{1 + 1/jQ_1},$$

and if Q_1 and Q_2 are reasonably large, this $= 1$. At $\omega^2 L_1C_1 = \omega^2 L_2C_2 = 1$,

$$(13) = \frac{1 - (1 + 1/jQ_2)}{1 - (1 + 1/jQ_1)} = \frac{Q_1}{Q_2} = m.$$

If m is not equal to unity we have a considerable variation in (13) and hence in the image impedance, which is $R\sqrt{(1+1/jQ_1)}$ at low frequencies, $R\sqrt{(1+1/jQ_2)}$ at high frequencies, and $[R\sqrt{(1+1/jQ_1)}] \times \sqrt{m}$ at $\omega = \omega_0 = 1/\sqrt{(L_1C_1)} = 1/\sqrt{(L_2C_2)}$.

If this variation in image impedance were gradual it might be allowed for, so that the impedance would be almost constant in the desired range. But the variation is not gradual, for (15) shows that the impedance is nearly equal to $R\sqrt{(1+1/jQ_1)}$ except near $\omega^2 = 1/L_1C_1$, near which frequency, therefore, the whole of the variation takes place. This region is thus one of rapidly varying impedance. This would cause large reflection losses and would upset the phase shift also. The

Q's should thus be made equal at $\omega = 1/\sqrt{(L_1C_1)} = 1/\sqrt{(L_2C_2,)}$ since here only is there a rapid change of impedance.

We may correct for impedance more exactly by the method described in the beginning of the paper. We do this by putting a resistance R^2/R_2 in parallel with C_1, and a resistance R^2/R_1 across C_2. This is shown in Fig. 7.

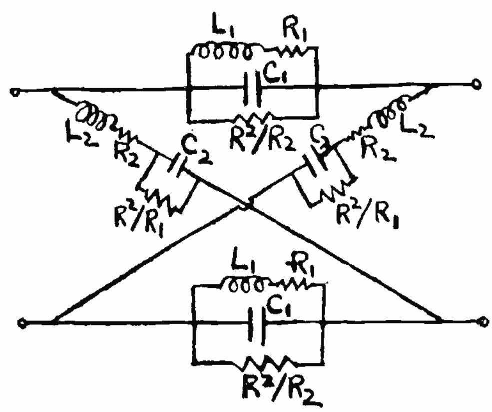

Fig. 7

R_1 and R_2 should be the alternating-current resistances at the critical frequency $\omega = 1/\sqrt{L_1C_1} = 1/\sqrt{L_2C_2}$, if this frequency is in the transmitted range, otherwise at the mid-frequency. This method will correct impedance irregularities and, besides, it will produce a phase shift equal, to a first approximation, to the value for the ideal network. The unavoidable attenuation can be calculated by (9).

Recapitulation of Results

Ideal, type A network. (*Fig. 3.*)

$$\beta = 0 \text{ and } \alpha = 2 \tan^{-1}(\omega/\omega_0)$$

where,

$$\omega_0 = R/L = 1/\sqrt{LC}.$$

Image impedance is

$$R = \sqrt{L/C}.$$

Actual, type A network. This has a resistance r in the coils. Let,

$$\omega L/r = Q.$$

Then,

$$\beta = 2R/\left[Q\omega L\left(1 + \frac{1}{\omega^2 LC}\right)^2\right] \text{ nepers,}$$

$$\alpha = 2 \tan^{-1}(\omega/\omega_0) + 2\omega^2 LC/[Q(1 + \omega^2 LC)^2,$$

and,

$$Z_i \doteq R + jR/2Q.$$

Corrected, type A network. (*Fig. 5.*) Here,

$$\beta = \frac{2\omega/\omega_0}{Q(1 + \omega^2/\omega_0^2)} \text{ nepers,}$$

$$\alpha = 2 \tan^{-1}(\omega/\omega_0),$$

and,

$$Z_i = R.$$

Ideal, type B network. (*Fig. 6.*)

$$\beta = 0, \qquad \alpha = 2 \tan^{-1}\left(\frac{\gamma x}{1 - x^2}\right),$$

and,

$$Z_i = R,$$

where,

$$x = \omega\sqrt{LC} \text{ and } \gamma = \sqrt{L_1/L_2}.$$

Actual, type B network. Coil L_1 has resistance R_1 and coil L_2 has R_2. Let $\omega L_1/R_1 = Q_1$, and $\omega L_2/R_2 = Q_2$. The image impedance is

$$Z_i = \sqrt{\frac{L_1}{C_2}}\left[\frac{\left(1 + \frac{1}{jQ_1}\right)\left(1 - \omega^2 L_2 C_2 \cdot \overline{1 + \frac{1}{jQ_2}}\right)}{\left(1 - \omega^2 L_1 C_1 \cdot \overline{1 + \frac{1}{jQ_1}}\right)}\right]^{1/2}$$

$$= R\left[\frac{\left(1 + \frac{1}{jQ_1}\right)\left(1 - \frac{\omega^2}{\omega_0^2} \cdot \overline{1 + \frac{1}{jQ_2}}\right)}{\left(1 - \frac{\omega^2}{\omega_0^2} \cdot \overline{1 + \frac{1}{jQ_1}}\right)}\right]^{1/2}.$$

When $Q_1 = Q_2$ this reduces to the simple form

$$Z_i = R\left(1 + \frac{1}{jQ_1}\right)^{1/2},$$

and the reflection effects are negligible except at low frequencies. In general $Q_1 \neq Q_2$ and there is a rapid and large variation of Z_i in the neighborhood of $\omega = \omega_0$, at which frequency

$$Z_i = R\left(1 + \frac{1}{jQ_1}\right)^{1/2}\left(\frac{Q_1}{Q_2}\right)^{1/2}.$$

Corrected, type B network. (*Fig. 7.*) Here β is given by equation (9); α is the same as for the ideal case; viz.,

$$2 \tan^{-1}\left(\frac{\gamma x}{1 - x^2}\right);$$

and Z_i is R.

BOOK REVIEWS

Drawings and Drafting Room Practice.—American Standard. 24 pages. American Standards Association, 29 West 39th Street, New York, N. Y. Price 50¢.

The American Standards Association has issued a printed pamphlet containing the drawings and drafting room practice adopted as an American Standard by the American Standards Association, May, 1935. The standards which are covered include the arrangement of views, line work, dimensioning, sheet sizes, and lettering and are the result of several years' work on the part of a large representative committee under the chairmanship of Professor Franklin DeR. Furman of Stevens Institute of Technology under the sponsorship of the Society for the Promotion of Engineering Education and The American Society of Mechanical Engineers. The pamphlet bears the A. S. A. designation Z 14.1—1935.

*L. E. Whittemore

The Fundamentals of Radio, by R. R. Ramsey, Professor of Physics, Indiana University. Second edition, revised and enlarged, 426 pages. Published by the Ramsey Publishing Co., Bloomington, Ind., Price $3.50.

In this work the author has "endeavored to give the basic theory of radio as it is exemplified in modern practice." A knowledge of electrical theory, such as may be gained in a first course in physics, is presupposed. With this understanding, the first four chapters on elementary electricity are to be considered as a concise and convenient review of fundamental principles, and are not intended as a thorough presentation of the subject. The introductory chapters on capacity and inductance are likewise to be classed under the same heading.

The main treatment of the radio fundamentals follows a spiral approach. That is, more general principles are first passed in review, and a return is made later to consider them in more detail and to point out their applications in practice. The book is especially complete and valuable in the practical information it contains and in the full descriptions of apparatus. The teacher will find it a compact guide to modern practice.

The book is well provided with diagrams of circuits and is profusely illustrated with cuts of apparatus. In the case of the more complicated vacuum tube circuits a more detailed explanation of the circuits would be helpful to such readers as have had no previous acquaintance with vacuum tube "hookups."

This second edition is about fifteen per cent larger than the first edition, written in 1929, and the rapid progress of recent developements has necessitated considerable changes and revision. The author is to be congratulated on having succeeded in covering so much ground in a treatise of moderate size. The presentation is clear and logical, but a rather unusual staccato style of writing is employed throughout.

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BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained without charge by addressing the publishers.

The RCA Manufacturing Company, Radiotron Division, Harrison, N. J., has issued Application Note No. 48 on the graphical determination of the decrease in an inductance produced by a coil shield and Application Note No. 49 on the construction of a top-cap shield for metal tubes.

The Western Electric 400-watt radio transmitting equipment is described in a booklet issued by that organization which should be addressed at 195 Broadway, New York City. Another booklet describes studio speech input equipment for broadcast purposes and a third booklet describes a midget transmitter designed for private airplane flyers.

A leaflet issued by Thordarson Electric Manufacturing Company of 500 W. Huron Street, Chicago, Ill., describes a kit for the construction of a combined condenser capacitance and leakage testing instrument.

The Handee grinder produced by the Chicago Wheel and Manufacturing Company of 1101 W. Monroe Street, Chicago, Ill., is described in a leaflet issued by that organization.

Volume 2 of their booklet on "Service Hints" has recently been issued by the Hygrade Sylvania Corporation of Emporium, Pa.

Sensitive Research Instrument Corporation of 4545 Bronx Boulevard, New York City, has issued leaflet Volume 1, No. 4 on "Instruments for Use with Photo Electric Cells."

Remote controls for automobile radio receivers are described in Bulletin No. 37 issued by DeJur-Amsco Corporation, 95 Morton Street, New York City.

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Lindenblad, Nils Erik: Born October 30, 1895, at Norrkoping, Sweden. Studied mechanical engineering, Norrkoping Polytechnic Institute, 1912–1915; electrical engineering, Royal Institute of Technology, Stockholm, 1916–1919. Immigrated to the United States, October, 1919. Drafting department and radio engineering department, General Electric Company, 1920; engineering staff, RCA Communications, Inc., 1920 to date. Member, Institute of Radio Engineers, 1934.

Reber, Grote: Born December 22, 1911, at Chicago, Illinois. Received B.S. degree in electrical engineering, Armous Institute of Technology, 1933. With M. G. Harbula, 1933. Engineer, General Household Utilities Company, 1934; engineer, household radio department, Stewart-Warner Corporation, 1935. Radioman U. S. N. R., 1929–1933. Member, American Radio Relay League; American Rocket Society. Associate member, Institute of Radio Engineers, 1933.

Starr, A. T.: Born 1905, at London, England. Received B. S. degree, London University, 1925; B.A. degree (Wrangler), Cambridge University, 1927; M.A. degree, Cambridge University, 1931; Ph. D. degree, London University, 1934. Honorary mention, Smith's Prize, 1929. International Telephone and Telegraph Laboratories, 1928–1931; lecturer, Faraday House Electrical Engineering College, 1931 to date. Associate member, Institute of Radio Engineers, 1931.

Terman, Frederick Emmons: See Proceedings for March, 1935.

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The Institute of Radio Engineers

Incorporated

330 West 42nd Street, New York, N.Y.

APPLICATION FOR ASSOCIATE MEMBERSHIP

(Application forms for other grades of membership are obtainable from the Institute)

To the Board of Directors

Gentlemen:

I hereby make application for Associate membership in the Institute of Radio Engineers on the basis of my training and professional experience given herewith, and refer to the members named below who are personally familiar with my work.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I will be governed by the constitution of the Institute as long as I continue a member. Furthermore I agree to promote the objects of the Institute so far as shall be in my power, and if my membership shall be discontinued will return my membership badge.

....
(Sign with pen)

....
(Address for mail)

....
(Date) (City and State)

Sponsors:

(Signature of references not required here)

Mr. .... Mr. ....

Address .... Address ....

City and State .... City and State ....

Mr. ....

Address ....

City and State ....

The following extracts from the Constitution govern applications for admission to the Institute in the Associate grade:

ARTICLE II—MEMBERSHIP

Sec. 1: The membership of the Institute shall consist of: * * * (c) Associates, who shall be entitled to all the rights and privileges of the Institute except the right to hold any elective office specified in Article V. * * *

Sec. 4: An Associate shall be not less than twenty-one years of age and shall be a person who is interested in and connected with the study or application of radio science or the radio arts.

ARTICLE III—ADMISSION AND EXPULSIONS

Sec. 2: * * * Applicants shall give references to members of the Institute as follows: * * * for the grade of Associate, to three Fellows, Members, or Associates; * * * Each application for admission * * * shall embody a full record of the general technical education of the applicant and of his professional career.

ARTICLE IV—ENTRANCE FEE AND DUES

Sec. 1: * * * Entrance fee for the Associate grade of membership is $3.00 and annual dues are $6.00.

ENTRANCE FEE SHOULD ACCOMPANY APPLICATION

(Typewriting preferred in filling in this form) No. ....

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

Name ....
(Give full name, last name first)

Present Occupation ....
(Title and name of concern)

Business Address ....

Permanent Home Address ....

Place of Birth ....Date of Birth ....Age ......

Education ....

Degree ....
(college) (Date received)

TRAINING AND PROFESSIONAL EXPERIENCE

DATES	

Record may be continued on other sheets of this size if space is insufficient.

Receipt Acknowledged ....Elected ....Deferred............

Grade ....Advised of Election ....This Record Filed...........

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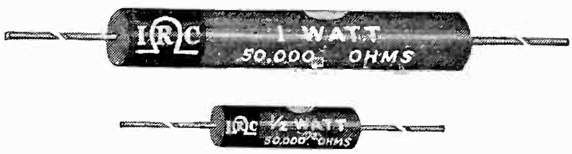

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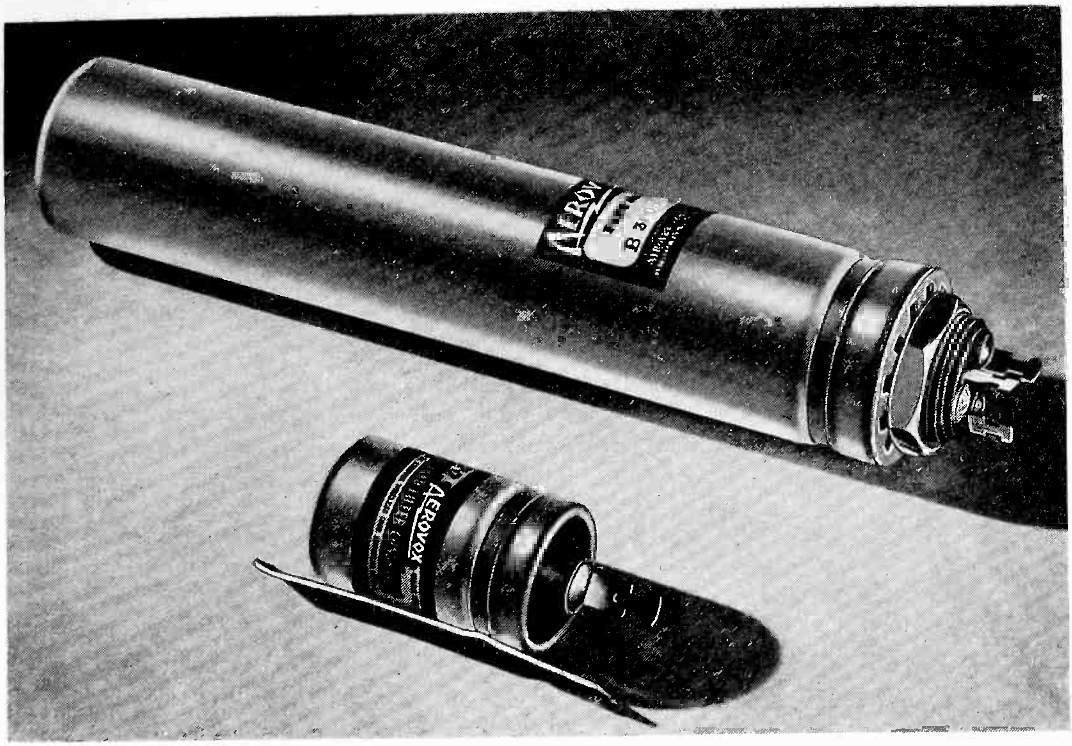

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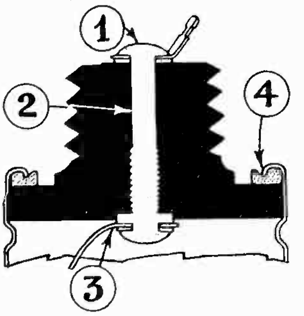

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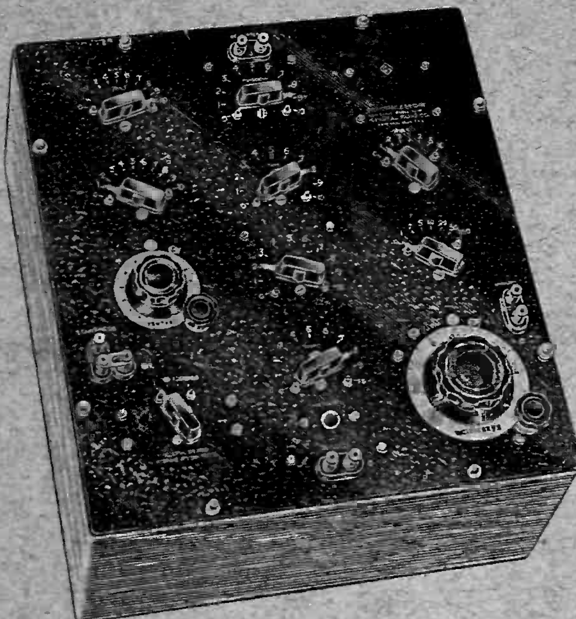

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