BELL LABORATORIES RECORD

BARKHAUSEN OSCILLATOR

F. B. LLEWELLYN

THE BRIDGED T EQUALIZER

A. L. STILLWELL

BALANCING CROSSTALK

M. A. WEAVER

AUGUST 1935 Vol. XIII No. 12

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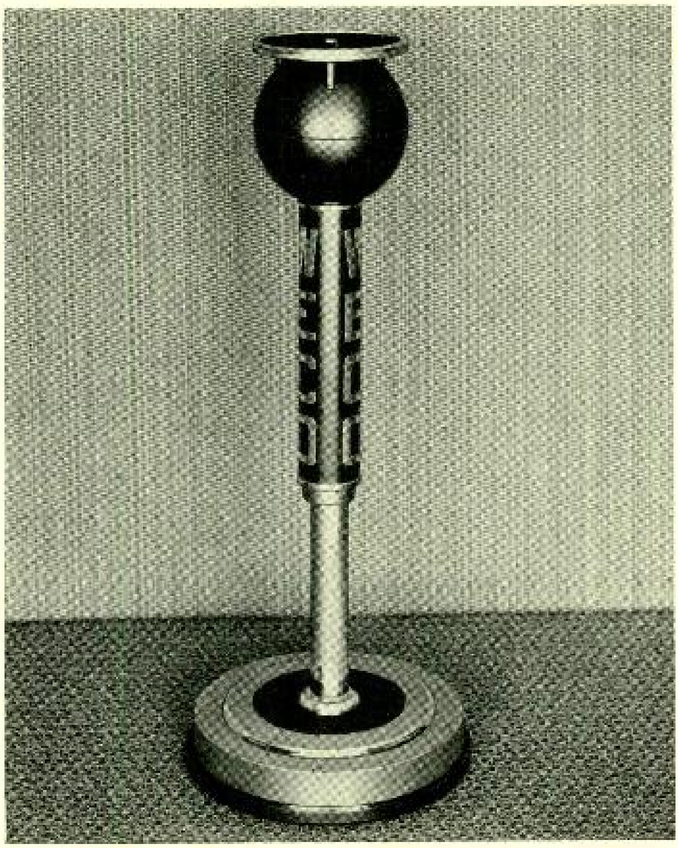

Non-directional Microphone

VOLUME THIRTEEN—NUMBER TWELVE

for

AUGUST

1935

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The Barkhausen Oscillator

By F. B. LLEWELLYN
Radio Research

THE Barkhausen oscillator for ultra-high frequencies has been the subject of many complicated analyses. As with other developments, the history of its theory has gone through three stages; starting with a simple but incomplete concept, advancing through a more and more involved mathematical analysis, and finally yielding the important properties in simple enough form to be stated in everyday language.

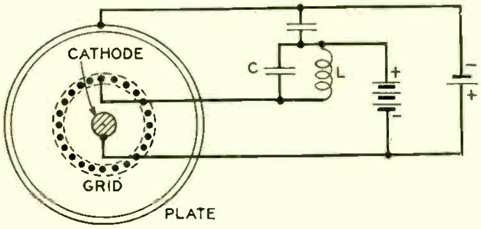

Fig. 1—This general arrangement of tube elements and circuit is used in the production of Barkhausen oscillations

The elementary concept upon which the explanation of the Barkhausen oscillator was originally based was that of an electron or group of electrons which danced back and forth through the opening, in a positively polarized grid. Among other things this concept did not show why the number of electrons going in opposite directions through the grid was not always the same. The mathematics which followed led in many cases into strange paths, but finally was placed on a sound basis. When this had been done it was found that the explanation could be made in a perfectly straightforward way, and that the original concept of the dancing electron contained only enough truth to delay and complicate the process of arriving at the correct answer.

Physically the Barkhausen oscillator consists of a filamentary cathode surrounded by a cylindrical grid and plate, as shown in Figure 1. The grid is operated at a positive potential, the plate is biased to a slightly negative potential, and a tuned circuited LC is usually connected between them.

Electrons starting out from the cathode are attracted by the positive grid, and move with increasing velocity in that direction. Many of the electrons hit the grid, and become of no further concern. Many others, however, pass between the grid wires without hitting them and move toward the plate with decreasing ve-

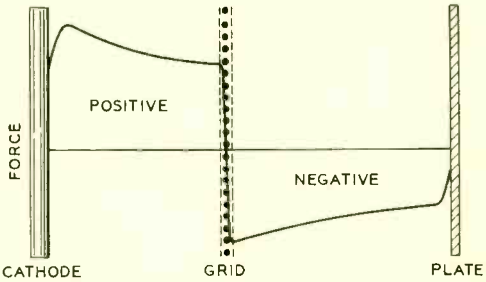

Fig. 2—When a constant potential is applied to the grid of a vacuum tube constructed as shown in Figure 1, the force exerted on the electron at various points in its trip across the tube is as shown by the curve. The work done on the electron is therefore measured by the area under the curve

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locity, coming to rest just before reaching the negative plate and then starting back toward the grid again. As before, a given electron may hit the grid or may miss it again and continue on toward the cathode, when the cycle is repeated.

If the tube were not oscillating, the story would now be complete. A convenient way of seeing what causes the oscillations in the tube is to investigate what happens when a transient oscillation is produced in the LC circuit by some external means. If the forces produced on the moving electrons by this transient deliver energy to the electrons, the transient will die out; but if the electrons, having acquired kinetic energy from the grid battery, can oppose the forces set up by the transient and thus deliver energy to the circuit, the transient will persist or build up as a continuous oscillation instead of dying out.

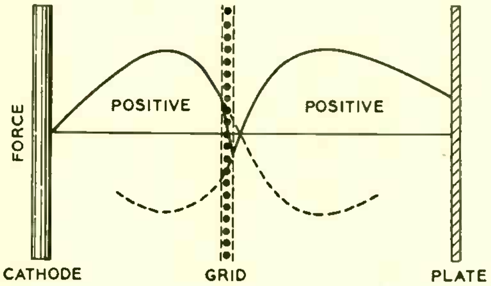

Fig. 3—Because the work done on an electron of the useless type by an alternating force is positive, the electron abstracts energy from the alternating-current transient

In the absence of the transient, an electron starting from the filament moves in the direction of the force from the positive grid, and so draws energy from the grid battery. After passing through the grid wires, the electron moves against the force from the grid, thus returning energy to the grid battery. When it comes to rest in the grid-plate space, the entire amount of energy picked up in the cathode-grid space has been restored to the battery. During the return trip, the same sequence again occurs: the energy which was abstracted in moving toward the grid is restored in moving away from it. The net result is that energy is neither absorbed from nor delivered to the external circuit and batteries by such an electron. This fact is made graphically evident in Figure 2, where the force from the batteries which acts on the electron is plotted as ordinate while the distance the electron has moved from the cathode is taken as abscissa. The area under the curve consequently measures the work done on the electron during its trip across the tube: the positive work done between cathode and grid is just equal to the negative work done between the grid and the plate.

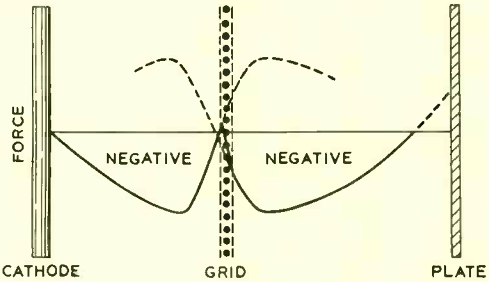

Fig. 4—With an alternating force, an electron of the useful type delivers energy to the transient, at the expense of the battery producing the constant force

When the transient is introduced into the LC circuit, conditions are considerably changed. Superposed on the force diagrammed in Figure 2 is the force set up by the alternating grid potential. Since the latter force alternates, the resultant forces acting on electrons which start out from the cathode at different times in the alternating-current cycle will be quite

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different. For purposes of illustration, it is sufficient to trace the history of two electrons only. One of these starts out just at the time when the alternating force begins acting in the same direction as the constant force of Figure 2. The other starts out a half cycle later, when the alternating force begins to oppose the constant force.

In the first case, the alternating force increases in intensity as the electron moves along, then decreases to zero, then reverses and opposes the motion, and finally completes the cycle by becoming positive again. If the electron passes through the grid mesh just before the alternating force returns to its first zero value, the action of the force upon it is as shown in Figure 3. At the instant the electron passes through the grid, of course, the direction of the force acting on the electron reverses, not because of any abrupt change in the grid potential, but because the grid is now located behind the electron instead of ahead of it. As the electron moves on toward the plate, however, the alternating force decreases to zero and then reverses. Thus during both halves of the cycle the force acts in the same direction as the electron is moving, and delivers energy to it, as can be checked by reference to the area under the curve in Figure 3. In other words, the transient in the external circuit has done work on this particular electron, and the electron, by taking energy away from the circuit, has produced a tendency for the transient to die out.

There is nothing in this behavior that gives encouragement to the maintenance of oscillations. The only consolation comes from noting that the electron is moving faster when it approaches the plate than it would if no alternating forces had acted on it, and consequently it will hit the plate even though the latter be at a slightly negative potential. Thus this useless, and in fact harmful, electron is at least prevented from doing still further harm by being removed through the plate from the scene of action.

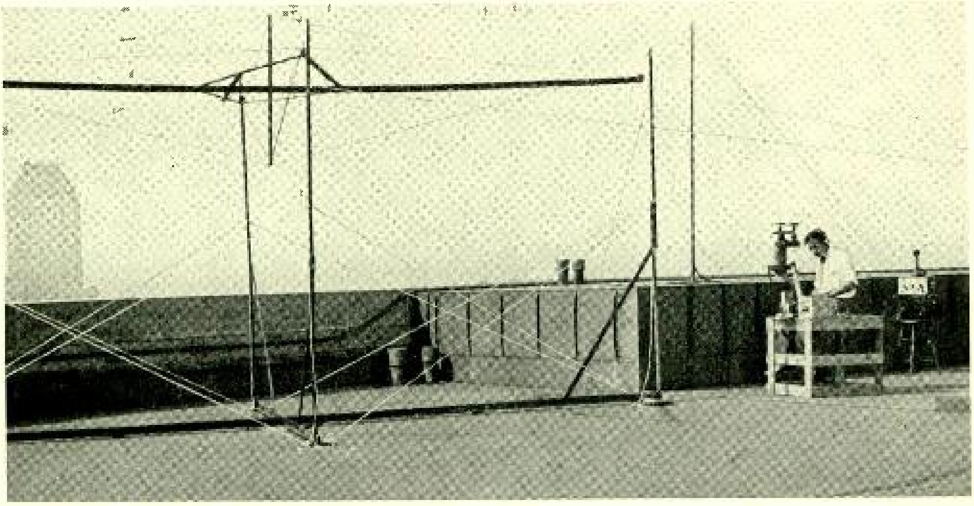

Fig. 5—During certain experiments on ultra-short-wave transmission, a receiving layout was placed on the roof of the New Jersey Telephone Company's headquarters building in Newark, to receive 60-centimeter waves transmitted from Barkhausen oscillators located in New York

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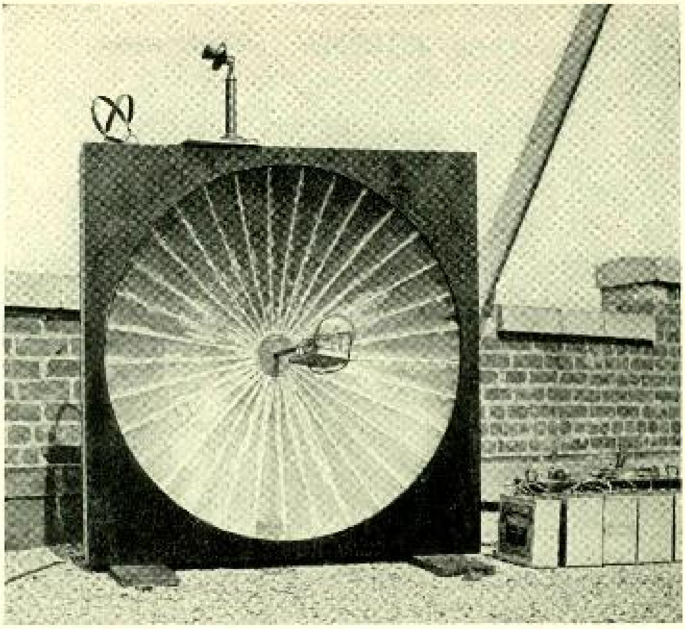

Fig. 6—Experiments were also performed in which a transmitter, incorporating a Barkhausen oscillator and associated with a parabolic reflector, transmitted 28*-centimeter waves from the roof of the Graybar-Varick Building to the New York Telephone Company's building at* 140 *West Street*

Fortunately the other electron, that leaves a half cycle later than the worthless one just dismissed, is more useful. From the very start the alternating force opposes the motion of the new electron, but cannot stop it because the alternating force is never larger than the constant force of Figure 2. The electron is therefore doing work against the alternating force, delivering energy to the transient in the external circuit. As the electron progresses, the phase of the force changes as shown in Figure 4. Unlike Figure 3, the reversal in direction occurs before the electron has reached the grid, because the force opposing the motion has decreased the speed. When passage through the grid mesh again reverses the direction of the force, the agreeable electron continues to deliver energy to the circuit transient, as it approaches the plate. Having lost much of its velocity in transferring its energy to the circuit, the electron comes to rest before hitting the plate, and then, urged by the constant force of Figure 2, starts on its return journey toward the grid. At about the same time, the phase of the alternating force again reverses and again opposes the motion, so that the hapless electron must give up still more of its energy to the growing transient.

Another passage through the grid follows, accompanied by another reversal in the phase of the alternating force, and the tormented electron must again yield energy acquired from the constant force to the hungry transient. The energy loss brings the electron to a halt before it reaches the cathode, the phases again reverse, and the cycle starts over again.

In its round trip, the useful electron of Figure 4 supplies to the transient nearly twice as much energy as the useless one of Figure 3 abstracted in its one-way trip. Moreover, the useful electron reaches the cathode again at just the right time to join with other electrons of the useful type and augment their relative number. The action is consequently progressive, building up more and more useful electrons.

In practice, operating conditions modify somewhat the mechanism described. For example, space charge

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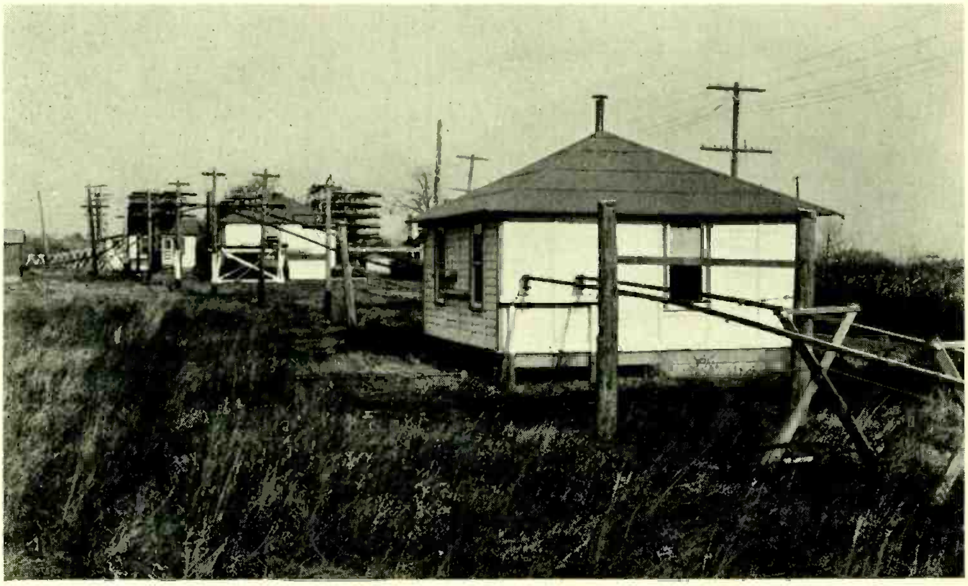
near the cathode produces more harmful electrons than useful ones and is to be avoided. Space charge near the plate causes a shift between the phase of the grid voltage and the force acting on the electrons, which in general tends to raise the frequency of oscillation. On the other hand, space charge in general makes the electrons move slower. Since the slower motion tends to decrease the frequency, the net result of space charge near the plate is only a small decrease in frequency. The tuning of the external circuit can also modify the frequency by about thirty per cent, but for fixed values of plate, grid, and filament battery voltage, there is a particular tuning adjustment which gives maximum output.

There is a simple approximate expression for determining the proper grid voltage and size of vacuum tube to produce oscillations of a given wavelength. For example, a tube with a plate diameter of one centimeter, and with 100 volts applied to the grid, will produce oscillations with a wavelength somewhere between 100 and 50 centimeters, corresponding to a frequency between 300 and 600 megacycles, depending on the circuit adjustments.

It is interesting to note that the same kind of analysis here used to illustrate the workings of the Barkhausen oscillator can be applied to the well-known feedback oscillators operating with negative grid and positive plate, and shows that the two are not very different from each other after all. The Barkhausen oscillator will probably prove very useful in the fields of ultra-short-wave transmission, which are rapidly coming into commercial application.



Experimental installation of large coaxial conductor lines near Phœnixville, Pennsylvania

Balancing Crosstalk in Toll Cables

By M. A. WEAVER
Transmission Development

FOR twenty-five years the Bell System has been using toll cables in which two twisted pairs are in turn twisted together to form "quads." Two "side" circuits are operated in each quad, one over each pair. In addition, a "phantom" circuit is operated in each quad, using one pair as the outgoing path and the other pair as the returning path. When many circuits run as close together as they do under these circumstances, crosstalk is especially likely to take place between them.

The crosstalk which would occur if the wires were straight and parallel is greatly reduced by twisting the wires and pairs together, for each twist acts in the same way as a transposition in an open wire line.* Even with the greatest care and skill in manufacture, however, further crosstalk reduction measures have always been found necessary during installation of the cable.

Crosstalk between circuits in different quads is minimized by splicing successive lengths of cable in such a manner that no two quads are adjacent for more than a small part of their total length. Until recently the sole means for minimizing the crosstalk between circuits in the same quad was by the method of "capacitance unbalance test splicing."

The lack of suitable balance, among the many capacitances between wire and wire and between wire and sheath, would be the principal source of crosstalk in a loaded cable if this test splicing were not done. By splicing successive lengths of cable so as to play one unbalance off against another, the net unbalance is so greatly reduced that other sources of crosstalk become about equally important. In the last five years, it has been found economical to spend less time on test splicing and to preserve the same degree of unbalance reduction by using supplementary balancing condensers.

The nature of the capacitance unbalances within a quad is shown in Figure 1, where wires 1 and 2 form one side circuit and wires 3 and 4 the other. The direct capacitances between the four wires of the quad are indicated at A, B, C and D; and those

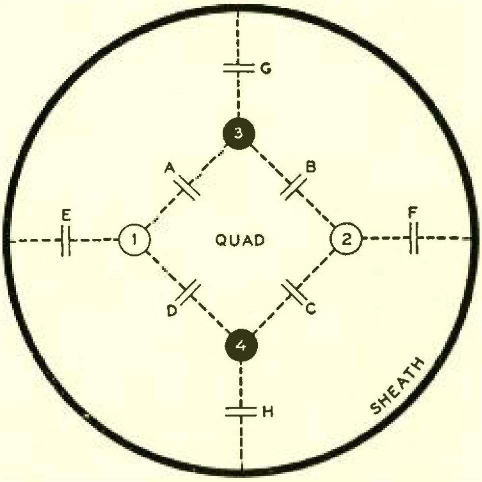


Fig. 1—Crosstalk between the three circuits of a quad (two side and one phantom) would be excessive without reduction of the capacitance unbalance, in which eight capacitances are involved

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*Record, *November*, 1934, *p.* 66.

between each wire and the sheath (including all other wires in the cable) are shown at E, F, G and H. The direct capacitances between the two wires of each pair are not indicated since they do not affect the capacitance unbalance. In a perfectly constructed quad A, B, C and D would all be the same, E and F would be the same, and G and H would be the same. In an actual quad this symmetry is not exactly obtained. At the time the cable is installed, the effect of the dissymmetries in causing crosstalk is measured with a 4A Capacitance Unbalance Set. The several lengths of cable in each loading section are then spliced in such a way as to approach symmetry of the direct capacitances for the loading section.

Until 1919, such tests and splices were made at seven places in each loading section. For this purpose, each loading section was divided into eight approximately equal lengths as indicated in Figure 2 and splices occurring between test points were made. Capacitance unbalance tests were then made at splices 1, 3, 5 and 7, measuring the unbalances in the lengths to the left and to the right of each point, and those lengths were so spliced as to minimize the net unbalance. While making the tests at 1 and 3, a watch was kept to determine whether any large net unbalances remained in either length 0-2 or length 2-4. If such unbalances remained in either length, similarly large unbalances were deliberately built up in the other length so that opposing unbalances of approximately equal magnitude would result in the lengths to be joined at 2. Similar tests were made and coordinated on lengths 4-6 and 6-8. This procedure was then repeated at 2 and 6, and finally at 4.

About fifteen years ago, tests at splices 1, 3, 5 and 7 were discontinued. The resulting three-test plan was made permissible by the lower capacitance unbalances per reel length resulting from improved cable design and manufacturing processes, and effected a considerable reduction in the cost of field testing. Ten years later, studies indicated that satisfactory crosstalk reduction could be obtained at a still lower cost, by spending less time on the test splicing and supplementing it by the addition of suitable balancing condensers.

A type of balancing condenser, which had been used to a limited extent where a small number of circuits were to be balanced, is the shielded twisted pair known from its specifica-

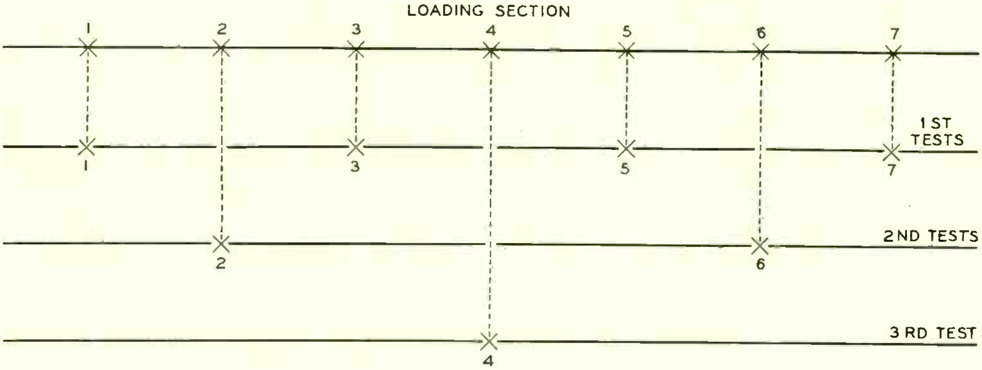

Fig. 2—*Until* 1919 *capacitance unbalance tests were made at seven points in each loading section*

tion number as the KS-1955 pair. Pairs of wires in a stub cable had also been used as balancing condensers in special cases where large numbers of circuits were involved, as in relocation work on toll cables and in the necessarily long sections of toll cable which cross water courses. In general these two types of balancing condensers behave alike: in the stub cable pair the other cable conductors and the sheath take the place of the KS-1955 shield.

In Figure 3 the three direct capacitances of such a pair are designated x, y and z. If, for example, capacitance A of the quad in Figure 1 is too low, a suitable length of KS-1955 pair can be connected between wires 1 and 3 of the quad, adding capacitance x to A. Since the shield of the KS-1955 pair is connected to the sheath of the cable, y is also added to E and z to G, and unbalances involving E and G are also changed. If the shield were left floating, the capacitances added to E and G would be uncertain, since they would depend upon the capacitance between the shield and the sheath, changing with their relative positions. To avoid these difficulties, a new unit was desirable in which the capacitance x was large compared with the capacitances from the wires of the unit to other wires or sheath.

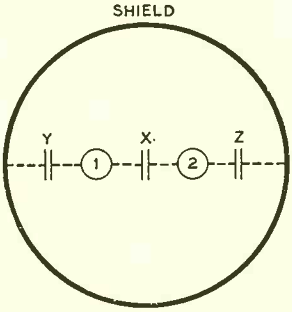

Fig. 3—In using the shielded twisted pair, pictured at the left of Figure 4, not only the capacitance between the two wires but those between each wire and the shield must be considered

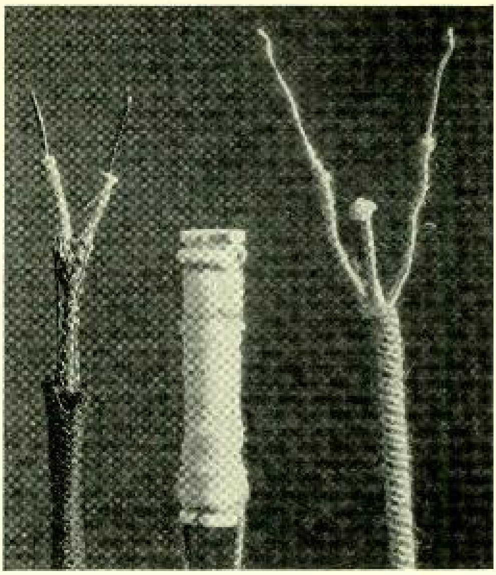

Fig. 4—To balance capacitances at cable splices, the shielded twisted pair formerly used (left) has been replaced by a pair wound spirally around an insulating core. At the end of a repeater section the unit shown in the center and in Figure 5 is used

Such a unit was developed by the Outside Plant Department and has been standardized for Bell System use (Figure 4, right). In general, it is made by wrapping two parallel insulated 22-gauge wires helically around a non-conducting core. The new unit has been designed to have an "x" capacitance of about 5 micro-microfarads per inch, a convenient value for field use. A foot of the new unit, although much smaller in diameter, is as effective as about two and a half feet of the KS-1955 pair in reducing phantom-to-side unbalances, and as effective as about seven feet of the older unit in reducing side-to-side unbalances. In order to make the y and z capacitances small enough to neglect, the units are covered with a

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cotton sleeve when they are installed. From the measurement of the three unbalances between the circuits of a quad, it is very simple to determine, with this type of unit, a set of three balancing capacitances which will reduce unbalances to almost nothing.

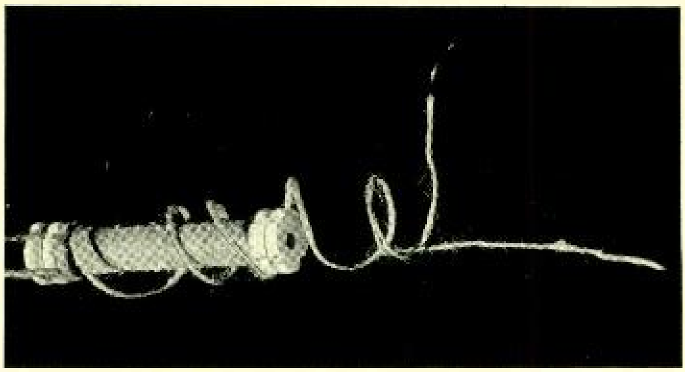

Fig. 5—The balancing condensers are adjusted when they are installed, by unwinding and removing a portion of the wire from the spool

On cables in which the larger proportion of the circuits are to be operated on a four-wire basis, only one test splice per loading section is now made. There, excessive unbalances remaining after test splicing on the small proportion of "two-wire" quads (those in which transmission takes place in both directions over each circuit) are corrected by means of the new balancing unit. On the "four-wire" quads, the test splicing is confined to the reduction of phantom-to-side unbalances alone. After the loading coils have been installed, phantom-to-side, side-to-phantom and side-to-side far-end crosstalk measurements are made on the "four-wire" quads over the complete repeater section. If the crosstalk values (average, maximum or both) are excessive, they are brought within the permissible values by installing balancing condensers at one end of the repeater section on a sufficient number of the quads having the larger crosstalk values.

Since the balancing capacitances required at the repeater station may be fairly large, the lengths of the balancing unit required would often become prohibitive. For this purpose balancing condensers like that shown in Figures 4 (center) and 5, with capacitances of about 215 and 610 micro-microfarads, were developed by the Telephone Apparatus Development group. They consist of small Isolantite spools, about an inch and three quarters long, on which are wound two parallel 30-gauge double cotton covered wires. Two 18-inch leads facilitate connections to the cable conductors. The condensers are impregnated in paraffin to exclude moisture. They are adjusted in the field by unwinding and removing a portion of the wire from the outer part of the spool. As in the case of the balancing unit already described, the y and z capacitances are kept small by covering the condensers with cotton sleeves.

For cables in which the circuits are mostly of the two-wire type, a single test splice per loading section has been found impracticable due to the large number of balancing units required at the splice. In those cables it is proposed to continue making three test splices per section, but balancing condensers may be used as an alternative to coordination at splices 2 and 6.

Balancing the capacitances of four-wire circuits by condensers at one end of a repeater section does not necessarily reduce the crosstalk to zero.

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When circuits have the same attenuation and speed of propagation, a single balancing condenser will substantially neutralize the far-end crosstalk due to capacitance unbalance. When, however, circuits have different propagation constants, as do the phantom and side circuits of the H-44-25 system,* a single balancing condenser cannot do so. Furthermore, the balancing condensers giving the greatest reduction in phantom-to-side crosstalk and in side-to-phantom crosstalk are different, and a compromise value must be used. For cases requiring more complete neutralization than can be obtained with a single condenser, there is available a method in which balancing condensers are placed at both ends of a repeater section. Another method is available in which balancing series resistances neutralize the crosstalk component at right angles to that neutralized by the balancing condensers.

These more complicated methods will probably not be required. Field trials indicate that the present crosstalk requirements can be readily met by the simpler one-end balancing. In one trial all of the four-wire loaded quads in a 39-mile cable were balanced by one test splice per loading section, supplemented by condensers placed at one end, which brought the crosstalk values well within tolerable amounts. In a second trial on a 42-mile cable, supplementary balancing condensers were used with only a sufficient number of quads to bring the crosstalk within tolerable values. It was found that only about forty-three per cent of the quads had to have balancing condensers.

This method of localized balancing neutralizes only far-end crosstalk, propagated in the direction of the signal, and does not eliminate near-end crosstalk, propagated in the other direction. Since, however, repeaters are one-way devices, near-end crosstalk is important only in circuits transmitting in the opposite direction, and four-wire circuits transmitting in opposite directions are segregated in a cable so that they do not run close enough for crosstalk to be important. This method can be used in maintenance work on four-wire circuits in existing cables as well as on new cable installations, and is rapidly replacing the older methods in the Bell System's cable plant.

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*RECORD, *September*, 1931, *p.* 6.

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A Bone-Conduction Receiver for the Audiphone

By R. C. MINER
Transmission Instruments Engineering

PROBLEMS encountered in designing aids for the hard-of-hearing are closely related to those that arise in developing telephone systems and apparatus. This close alliance between the two studies extends not only to the fundamental physical laws involved, but to an equally close historical association, since the inventor of the telephone, Alexander Graham Bell, was primarily a teacher of the deaf. His work in telephony arose largely from his intensive study and deep interest in the science of sound and hearing. It is natural, therefore, that Bell Telephone Laboratories should undertake the development of hearing aids. As a result of their studies, the highly satisfactory audiphone employing air-conduction receivers, made available through the Western Electric Company,* has recently been augmented by a new bone-conduction receiver, which may be employed interchangeably with the air-conduction receiver in connection with any of the amplifier-type audiphones.

Hearing by bone conduction is not

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*Record, *June*, 1932, *pp.* 342, 346, and 362.

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a new discovery. As early as 1509 mention was made of hearing a tuning fork by bone conduction, when the fork was pressed against the teeth or mastoid bone. More recently it has been common practice for physicians to employ both air and bone conduction to aid their diagnosis of hearing defects. That a hearing aid could be made employing bone conduction was demonstrated by the American Telephone and Telegraph Company before an Otological conference in Boston as early as 1912, and for the last twelve or fifteen years a bone conduction receiver has been available for use with Western Electric audiometers.

The development of hearing aids subsequent to 1912, however, dealt for the most part with the air-conduction type because of the inherently greater efficiency of transferring energy to the ear by conduction through the air. Under certain conditions, however, it is found that better results may be obtained by using bone conduction in spite of its lower efficiency of transmission. To extend the usefulness of the Western Electric audiphones, therefore, the 700 type bone-conduction receiver was developed. It is shown in use in the illustration at the head of this article.

This new receiver should prove valuable under a variety of conditions, but it is intended to supplement rather than to replace the widely used air-conduction receiver. The mechanism of hearing, as already described in the Record,* comprises a middle and an inner ear, the latter containing the auditory nerves, and the former containing a series of very small bones which conduct the sound vibrations from the drum to the inner ear. Hearing may be reduced either by failure of the middle ear to transmit the sound or by some impairment of the inner ear, which is generally classified as nerve deafness. Which of these two types of deafness is predominant, and the degree of hearing impairment, determine to a large extent the type of hearing aid that will be most effective. With certain cases of middle ear deafness, where the hearing mechanism no longer is able to perform its function of efficiently carrying sound to the auditory nerve, the bone-conduction receiver is more satisfactory. Since with this type of deafness the inner ear is in normal condition, all that is needed is to get the sound transmitted to it without having to pass through the middle ear mechanism. Bone conduction offers a simple and convenient method of accomplishing this result, and middle-ear deafness is thus one of the important fields of use for the new receiver.

When it is the inner ear and the nerves that are affected, however, the air-conduction receiver is frequently the most satisfactory form of hearing aid. Its greater efficiency of energy transference makes it easier to obtain the high intensities which are required to produce an adequate response of the nerves.

Although the type of hearing aid

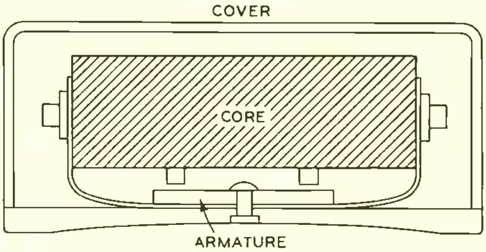

Fig. 1—Sound vibrations in the new bone-conduction receiver result from the reaction between a light armature and phenol-plastic cover assembly, and a relatively heavy magnetic structure

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*Record, *November*, 1928, *p.* 75.

that will be most effective generally depends on which of these two types of deafness is predominant, there are fields for use of the bone-conduction receiver where the type of deafness does not have so great an influence on the type of receiver employed. When the deafness is of mild form, for example, with appreciable hearing in both ears, the bone-conduction receiver has the advantage of offering a third channel of hearing to supplement the other two. Another advantage of the bone-conduction receiver under these conditions is that a single receiver helps the hearing of both ears. The vibrations are transmitted by the skull not only to the ear nearest the receiver, but across the head to the other ear with a loss of something less than ten decibels. Psychological factors also have a large influence on the type of receiver preferred. A wearer is sometimes less conscious of a receiver placed behind his ear and so converses more naturally. As a result he may express a preference for the bone-conduction receiver even though a careful test would show that he heard no better with this type of instrument than he did with the air-conduction type.

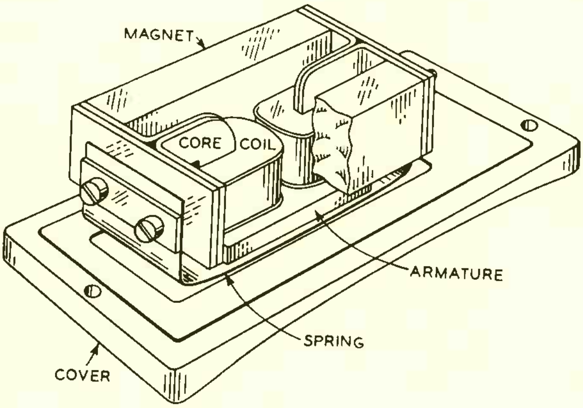

Fig. 2—Simplified sketch of the magnetic structure

Still another field for the bone-conduction receiver is with group audiphones in churches and theatres.* The air-conduction receivers, particularly when removed from the ear, radiate an appreciable amount of sound, while the bone-conduction receivers do not. As a result of this characteristic of the air-conduction receiver, those seated near people using them are sometimes disturbed by the radiated sound. The quietness of the bone-conduction receiver, on the other hand, makes it particularly suited for this service. It can, moreover, easily be employed to supplement the air-conduction receiver in most installations, since the inherent difference in efficiency may readily be overcome by increasing the amplification.

The construction of the Western Electric 700 type bone-conduction receiver is indicated in Figure 1. Sound vibrations transmitted to the skull result from the action and reaction between a very light phenol-plastic cover riveted to an armature, and a relatively heavy magnetic core structure. Voice currents pass through a winding on this core, and as the current varies, the armature is attracted with greater or less force. Both cover and core structure move as a result of these magnetic forces, since the two parts are fastened together by a phosphor-bronze spring, which is riveted to the armature and cover and screwed to the core structure as indicated in the illustration.

The cover, with overall dimensions of 1⅜ by ¾ by ⅝ inches, is in two parts: a rectangular case that encloses the electromagnets, and a front, which is slightly curved to

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*Record, *March*, 1931, *p.* 332.

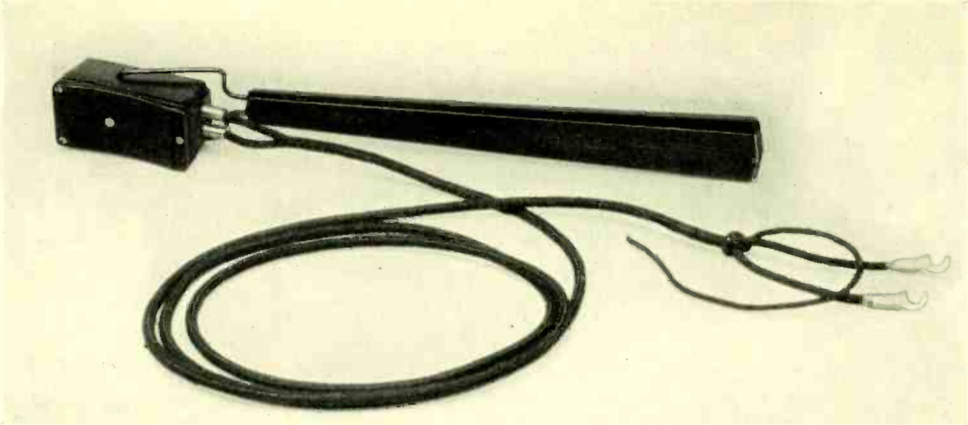

Fig. 3—A lorgnette type handle is one of the two types of mounting provided for the bone-conduction receiver

conform to the shape of the mastoid bone. Three screws fasten the front to the case. The arrangement of the magnetic system is indicated more clearly in Figure 2. It consists of two permanent bar magnets secured parallel to each other by cores which are bent in between the two magnets and extend down to form pole pieces. A coil is placed around each of these cores as shown. The permanent magnets maintain a strong steady flux through the armature and cores, which is modulated by the voice currents in the coils. The windings of the coils terminate in small spring-type jacks fastened to the cover. The entire assembly of magnetic and cover structure weighs only slightly over half an ounce.

Two methods of holding the new receiver are provided. One, shown in use in the illustration at the head of this article, is a head band of light music wire, which may be easily bent to a shape most comfortable to the user. The other, shown in Figure 3, is a lorgnette type of handle which is convenient when the receiver is to be used only at intervals. This handle is of black phenol-plastic with markings to match the embossed cover design.

Two new cords have been provided to allow the bone-conduction and air-conduction receivers to be used interchangeably with the 37 and 38 type audiphones. One of these, known as the K type, has small plugs at one end which fit into the jacks of the new receiver, while the other end has terminals for fastening to the audiphone. The other, or J type, cord—only fourteen inches long—serves as an adapter to allow the air-conduction receiver to be connected to the K type cord. It has terminals that fasten to the air-conduction receiver on one end, and small jacks, like those in the bone-conduction receiver, on the other. When the air-conduction receiver is to be used, the plugs on the end of the K type cord are inserted in these jacks on the end of the J type cord.

An important consideration in the design of all hearing aids is that they be as inconspicuous as possible. The bone-conduction receiver ranks very high in this respect. Since it rests behind the ear, and is very small in size, it is almost invisible from the front.

Studies made with the bone-conduction receiver have brought out a

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further advantage which at times may be important. The limit of assistance that can be obtained from an air-conduction receiver when the hearing loss is large is the threshold of feeling. The intensity can be raised only to more or less well-defined limits before the sound becomes painful. With a large hearing loss, the threshold of hearing may be almost as high as the threshold of feeling, so that little advantage can be obtained from the receiver. The bone-conduction receiver, on the other hand, can be operated at very high levels without unpleasant effects or any sensation of pain.

With its many advantages, the new 700 type bone-conduction receiver will prove a valuable supplement to Western Electric hearing aids. With three types of audiphones, two of which may be used with either of two types of receivers and any of several microphones, a user will have wide latitude in selecting a combination that most satisfactorily fits his particular needs.

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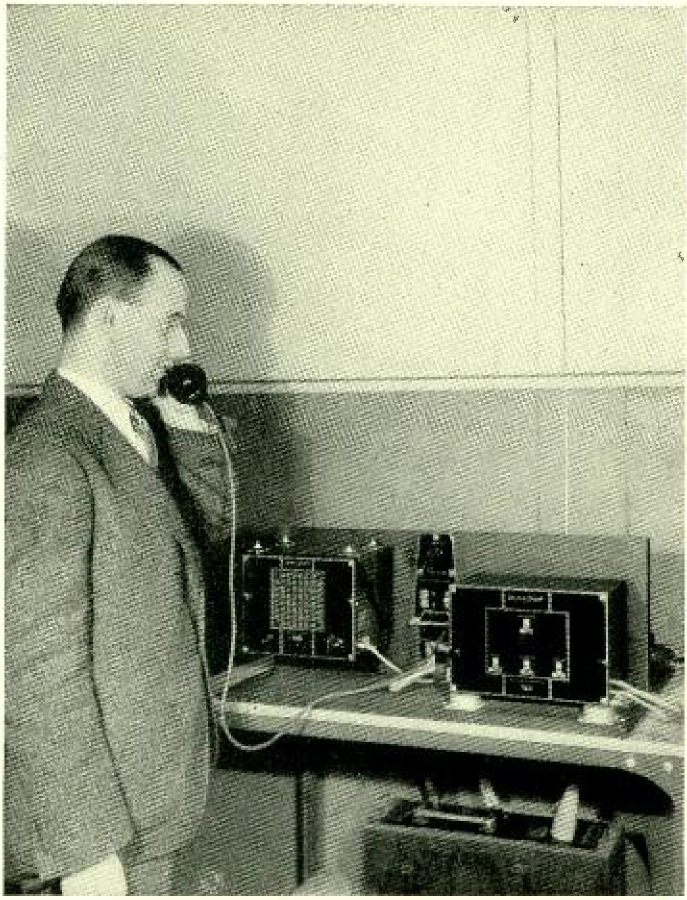

A. B. Bailey demonstrating use of ultra-short-wave police radio system for two-way communication

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The Bridged T Equalizer

By A. L. STILLWELL
Telephone Apparatus Development

ATTENUATION produced by telephone circuits varies, in general, with frequency. Where this change of attenuation with frequency is large over the transmitted band, attenuation equalizers are employed to reduce distortion of the signals. The application of attenuation equalizers to one type of service has already been discussed in the RECORD*. Except when a feed-back type of amplifier is employed, the function of an equalizer is to produce a loss which varies with frequency in a manner complementary to that caused by the circuit to be equalized. An important additional requirement generally placed is that both the input and output impedance of the equalizer should be the same at all frequencies, and that they should be pure resistances—usually of 600 ohms. These requirements have led almost universally to the adoption of the constant-resistance bridged T structure for the equalizer network.

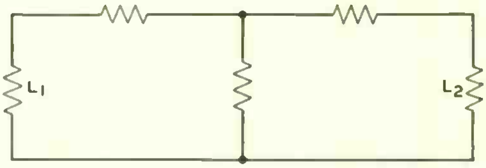

Fig. 2—A balanced T network has three impedances arranged in the form of a T, the two series ones being equal

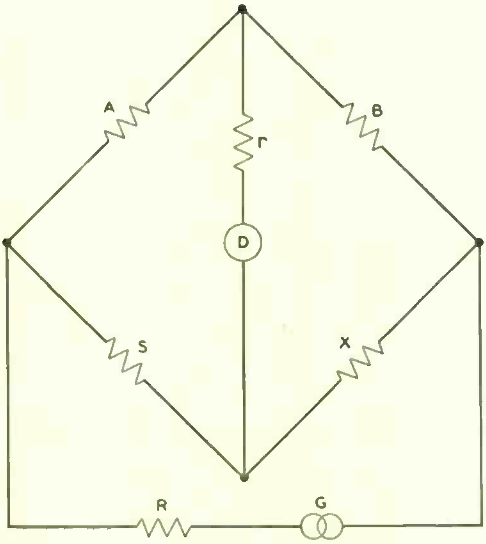

Fig. 1—A Wheatstone bridge is in balance, and no current flows through the detector, when BS = AX

The name of this equalizer derives from the T network with a bridging impedance across its series elements, but many of its desirable characteristics are exemplified in the ordinary Wheatstone bridge, shown schematically in Figure 1. A power supply G is connected to two opposite points of the bridge through the impedance R, and a detector D is connected across the other two points through an impedance r. The bridge is balanced when the ratio of the impedance of A to that of S is the same as that of B to X, or, put in a slightly different form, when BS = AX. Under these conditions no current flows through the detector.

The general form of a symmetrical T network is shown in Figure 2, where L_1 and L_2 are the input and out-

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*RECORD, *December*, 1933, *p.* 113.

put impedances respectively. The bridge circuit of Figure 1 may readily be arranged in this form as shown in Figure 3, where the lettering is the same as in Figure 1. Arranged in this manner, the T network proper is composed of A, r and B, with S as the bridging impedance, while R serves as the input impedance and X as the output impedance. If the impedances R, A, r, and X all be made alike and designated R, and the impedances S and B—for reasons discussed later—be designated Z_{11} and Z_{21} respectively, the T form of the bridge becomes as shown in Figure 4 and the condition for balance is that $Z_{11}\,Z_{21} = R^2$.

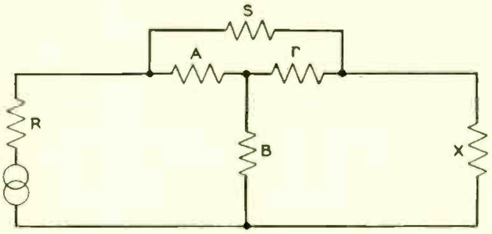

Fig. 3—A Wheatstone bridge circuit may be readily arranged in the form of a T network with X and R serving as the output and input impedances respectively

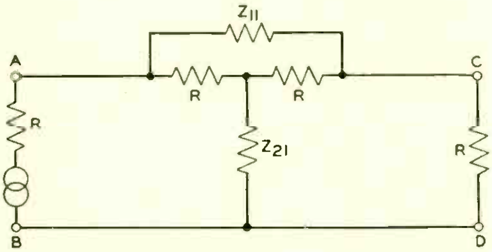

Fig. 4—A Wheatstone bridge will become a balanced T network with constant output and input impedances when R, A, r, and X are all made equal and when $BS = R^2$

If this balanced condition is maintained a simple calculation shows that the input impedance of the network—the impedance looking in at AB—is equal to R regardless of the value of Z_{11}, and that the output impedance—looking in CD—is also equal to R, which is obvious since the network is symmetrical. The loss produced by the network, on the other hand, which is a function of the ratio of the current I_1 in the output impedance before the network is inserted to I_2, the current after the network is inserted, is found to be equal to $1 \times (Z_{11}/R)$. Z_{21} is not involved in the expression since by the balance condition $Z_{21} = R^2/Z_{11}$. Thus the loss varies as Z_{11} and any value of loss may be secured without affecting the input and output impedances providing the balance condition has been maintained.

Z_{11} and Z_{21} stand for generalized impedances. They may be resistances, capacitances, inductances, or any combination of them. The only requirement is that established by the balance conditions ($Z_{11}\,Z_{21} = R^2$) which demands an inverse relationship between Z_{11} and Z_{21}. Thus if Z_{11} were a pure inductive reactance represented by $j\omega L$, Z_{21} would be $R^2/j\omega L$, which readily converts to $-j(R^2/\omega L)$. This corresponds to a capacitive reactance, $-j(1/\omega C)$, where $C = L/R^2$. If Z_{11} were a capacitance, on the other hand, Z_{21} would be an inductance, while if Z_{11} had been a resistance, Z_{21} would

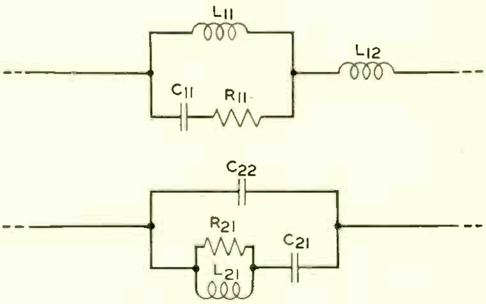

Fig. 5—One impedance is the inverse of another when their product is a constant value. The two networks shown are the inverse of each other

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also have been a resistance. When z_{11} is a network, z_{21} is a network of the same number of elements but each element is the inverse of the corresponding element of z_{11}: a series inductance in z_{11} becomes a parallel capacitance in z_{21}; a series capacitance, a parallel inductance; a parallel resonant circuit, a series resonant circuit, etc. Thus if z_{11} were the network shown in the upper part of Figure 5, z_{21} would be the inverse network shown in the lower part of the same illustration.

The reason for using a two-digit subscript for z here becomes evident. In all cases z_{11}, as evident in Figure 4, is in series with the output and z_{21} in shunt with it. The first digit of the subscript indicates whether the element belongs to the series or shunt impedance. The second digit is used to designate corresponding inverse elements of the two networks. Thus in Figure 5, C_{21} is the inverse of L_{11}; C_{22} is the inverse of L_{12}; L_{21} is the inverse of C_{11}; R_{21}, the inverse of R_{11}.

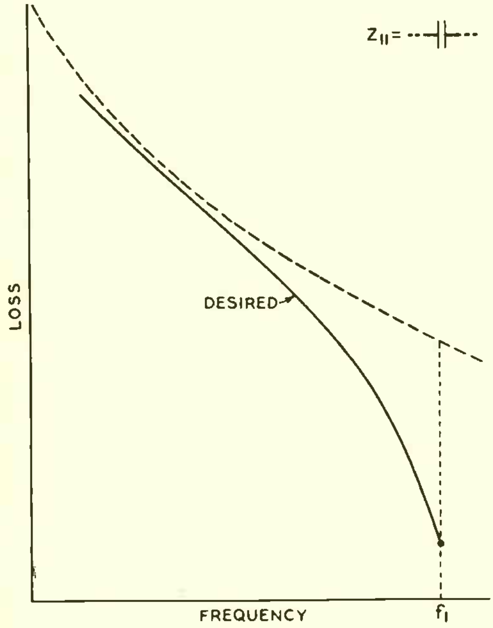

Fig. 6—Loss characteristics obtained by a capacity alone

An attenuation equalizer of this type, therefore, has a constant input and output impedance equal to R when z_{21} is the inverse of z_{11}. The loss characteristic, on the other hand, is determined by the network z_{11}.

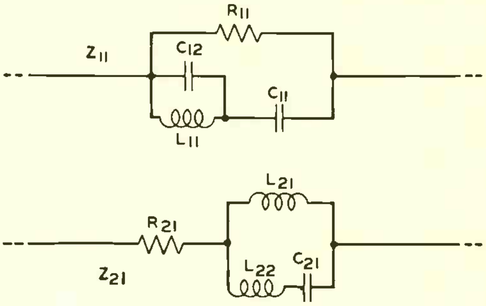

Fig. 7—The Z_{11} arm of a bridged T network giving loss indicated by solid curve of Figure 6. Below, the inverse network

To illustrate the effect of the various elements of z_{11}, a network giving the loss indicated by the solid curve of Figure 6 may be taken as an example. The network itself is shown in the upper part of Figure 7. The condenser C_{11} of this network, taken by itself, would produce a loss characteristic shown by the dotted curve of Figure 6. This shape of loss curve follows from the expression for the impedance of a condenser, which is $-j\ (1/2\pi fc)$. Since frequency occurs in the denominator of this expression, the impedance and hence the loss, decreases as frequency increases. An inductance in series with C_{11}, however, results in a series resonant circuit. Below its resonance frequency such a circuit also produces a loss that decreases with frequency, with the additional effect that the loss becomes practically zero at resonance. The effect of employing an inductance in series with C_{11} is thus to give the char-

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acteristic shown by the dotted curve of Figure 8. If we compare this with the dotted curve of Figure 6, we notice that by using an inductance to form a series resonance at approximately f_1, the loss there has been considerably reduced, while at the lower frequencies the loss has remained about the same. This is because the inductive reactance, while equal and opposite to the capacitive reactance at the resonance frequency—thus giving zero reactance—becomes decreasingly less with decreasing frequency and its effect on the increasingly large capacitive reactance is less pronounced. However, this effect does not drop out quickly enough as the dotted curve of Figure 8 is appreciably below the final characteristic at the mid frequencies of the band considered.

Fig. 8—Loss characteristics obtained by forming a series resonance circuit

It is thus apparent that an inductance is needed which is considerably smaller than that of L_{11} of Figure 8 over most of the frequency range but equal to it at f_1. This can be simulated by a parallel resonant circuit which has its parallel resonance frequency somewhat above f_1, since, up to the resonance frequency, a parallel resonant circuit behaves like an increasing large inductance. Thus L_{11} can be selected so that it is small enough to give approximately the right loss at the lower and mid frequencies, then by shunting a condenser around it, forming a parallel resonant circuit, the effective inductance of the parallel resonant combination at f_1 may be made equal to that of the former L_{11}. In this way the low impedance, and hence low loss, is preserved at f_1 and the loss is increased at the middle and lower frequencies. This gives the characteristic shown by the dotted curve of Figure 9, and the addition of the shunt resistance R_{11} modifies the loss below f_1, giving the solid curve.

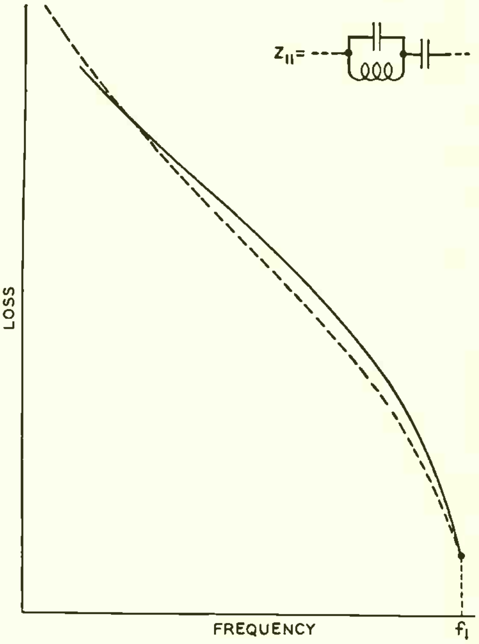

Fig. 9—Loss characteristics obtained by replacing the inductance of Figure 8 by a parallel resonant circuit, keeping the series resonance the same

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Z_{21}, shown in the lower part of Figure 7, is the inverse of z_{11}. The shunting resistance R_{11} becomes a series resistance R_{21}; the series condenser C_{11} becomes a shunt inductance L_{21}; while the parallel resonant circuit C_{12} and L_{11} becomes a series resonant circuit L_{22} and C_{21}. The two networks, connected to form the complete network, are shown in Figure 10.

The flexibility of the loss characteristic of an equalizer network depends on the complexity of the z_{11} and z_{21} arms. In general, for every component in either the z_{11} or z_{21} arm of the above type of equalizer there is a point through which the equalizer loss characteristic may be made to pass. Thus, if the z_{11} arm contains say four impedance elements, in general it is possible to make the equalizer loss pass through four predetermined points on any given loss characteristic. The degree of perfection with which a given characteristic can be matched by an equalizer therefore depends on the number of coils, condensers, or resistances it is considered economical to use.

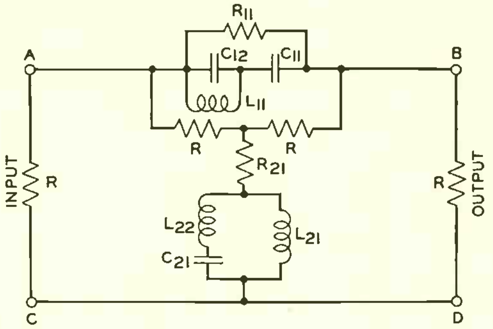

Fig. 10—*The balanced bridged T network to give constant input and output impedances and the loss characteristics shown by the solid curve of Figure* 9

Electrical Research Products, Incorporated

The Electrical Research Products, Incorporated, wholly owned subsidiary of the Western Electric Company, Incorporated, began business January 1, 1927. Its principal activity has been in the field of sound pictures.

It has paid the research and development expenses incident to its business (approximately $5,700,000) *and has paid its own way in every other respect. From its business of furnishing apparatus and technical services and licensing, it has earned net profits of* $9,450,000, $4,000,000 *of which has been paid to the Western Electric as cash dividends. In addition it has collected and paid over to the American Telephone and Telegraph Company* $5,700,000 *as royalties under patent licenses.*

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A High-Quality Broadcast Transmitter of Medium Power

By R. E. POOLE
Radio Development

THE installation of a number of high power broadcast transmitters should not be allowed to hide the fact that the backbone of the nation's broadcasting system is of necessity composed of smaller stations. To meet the need of these stations for high quality transmitters of moderate capacity, the Laboratories has designed a transmitter that may be used at any power up to 5 kw., and by minor modifications can be made to operate at 15 kw. The many desirable features of this transmitter have been widely recognized, and several stations employing this type of apparatus are under construction at the present time.

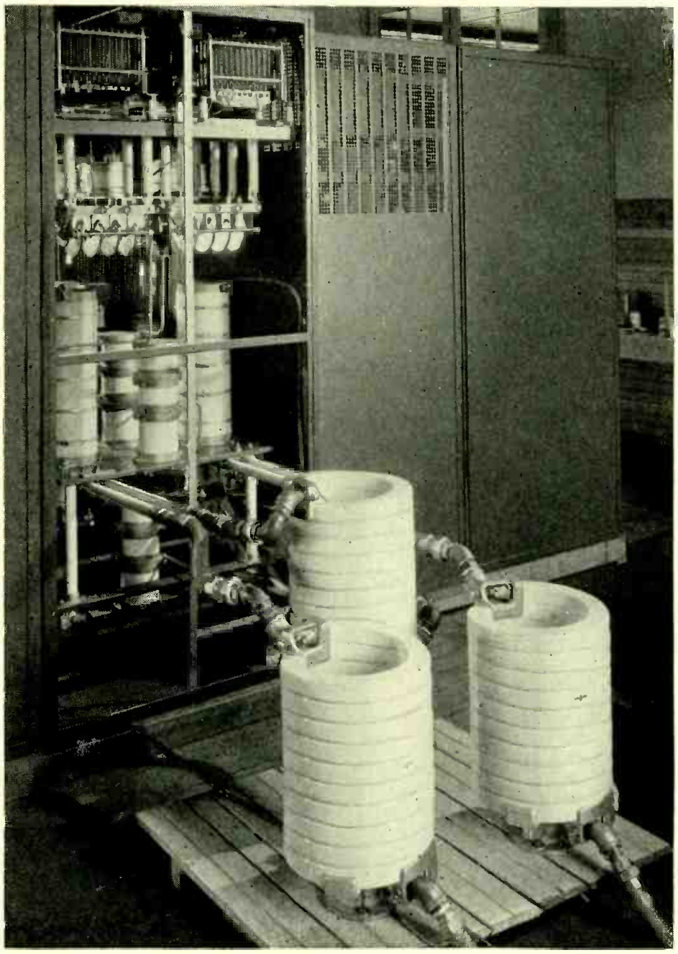

Fig. 1—Extruded porcelain coils serve as insulators in the cooling water line between the power amplifiers and the pumps

The new transmitter is shown in Figure 6 as set up for final test at the Whippany laboratory. The main assembly consists of five panel-type units. From left to right, these are the oscillator-modulator, the first power amplifier, the second power amplifier, the output tuning unit, and the control unit. The first two also contain their associated power supply ap-

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paratus. The plates of the second power amplifier are energized from a separate hot-cathode mercury-vapor rectifier which is mounted behind the panel assembly or in some other convenient location. Control of the transmitter is fully automatic; the operation of a single toggle switch energizes the entire equipment in proper sequence. Provision has also been made for reducing the antenna power from 5 kilowatts to 1 kilowatt without break in the program. This is accomplished by a remotely controlled switch, which throws the antenna from the second to the first power amplifier.

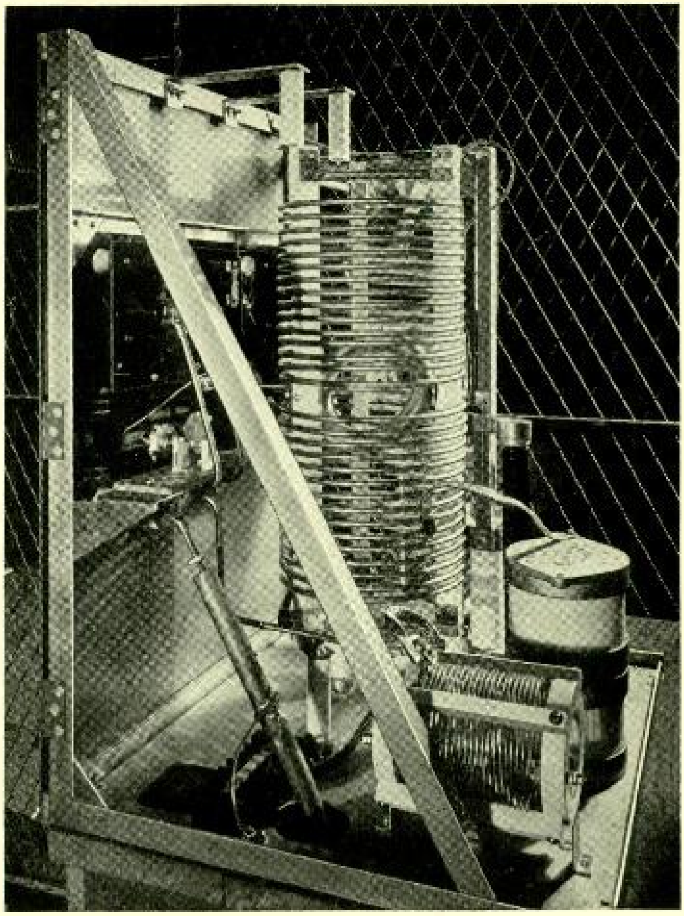
Fig. 2—Antenna coupling unit showing entering concentric conductor in lower left hand corner

One of the interesting features of this new transmitter is the control unit, at the extreme right in the illustration, which stands guard continuously over the operation of the transmitter. In this unit are the control switches, relays, water gauges, thermometers, and other accessory equipment required to start the various elements in proper sequence and to keep watch over their continuous operation. Centrally located on the upper half of the unit is a double row of pilot lamps, which show when all the important parts of the transmitter are working properly, and indicate the location of trouble when it arises. Should a flash-over occur at any of the tubes, the power will be tripped off, and a pilot lamp will be extinguished. Even though the power is immediately reapplied, this lamp will remain out to indicate where the trouble occurred so that it may be investigated at leisure and necessary repairs made. This indication of the location of trouble by the lamps makes a tube replacement or other repair a matter of comparatively short delay.

The oscillator-modulator unit, the first at the left in the illustration, is in reality a complete 100-watt transmitter. This unit includes the quartz-crystal oscillator, which is tempera-

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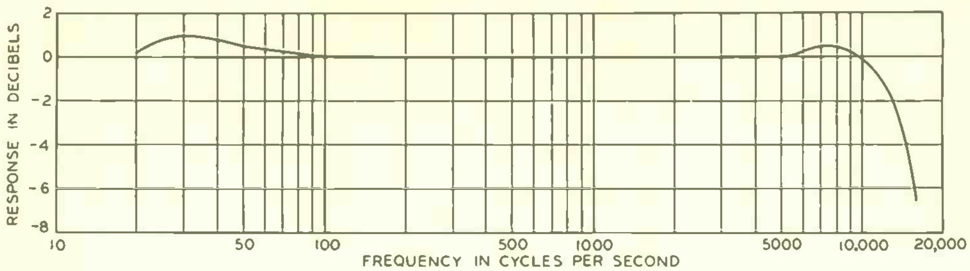

Fig. 3—Overall frequency characteristic of the 5-kw. broadcast transmitter

ture controlled and maintains its nominal frequency to better than 50 cycles. The temperature control circuit is independent of the operation of the rest of the transmitter, and holds the crystal at its operating temperature whether the transmitter is in service or not. Besides the oscillator, this unit also houses the modulating amplifier, employing two tubes in push-pull, and the output amplifier. The tube capacities are such that full 100% modulation of the carrier is obtained without appreciable distortion. The associated power supply apparatus is contained in the lower part of the cabinet.

The second unit from the left is a linear amplifier that increases the output of the first unit to 1000 watts. Rectifiers for grid and plate potential are in the lower part of the cabinet as in the oscillator-modulator unit. The output is delivered to a tuned circuit inductively coupled to a load circuit, which may be switched either to the antenna, through a concentric transmission line, or to the input circuit of the second power amplifier—the adjacent unit on the right. The switch that accomplishes this transfer is electrically operated and controlled by a toggle switch on the control unit. This transfer may be made almost instantaneously during a slight break in the program, so that there is no loss of program time. Both of the first two units have control switches on the front of their cabinets, and may be controlled either by these switches or from the main control unit.

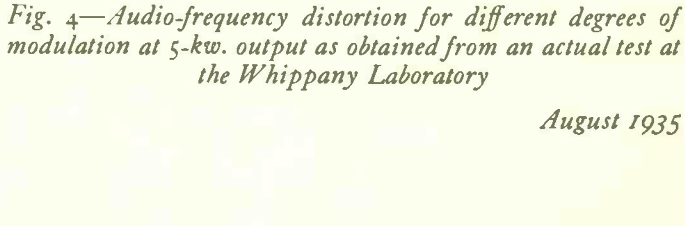

Fig. 4—Audio-frequency distortion for different degrees of modulation at 5-kw. output as obtained from an actual test at the Whippany Laboratory

The next two cabinets house the 5-kw. amplifier and its tuning apparatus. The power amplifier employs two water-cooled tubes. A view from the rear of the power amplifier, while the set was undergoing final tests at Whippany, is shown in Figure 1. The cooling water must necessarily be brought into contact with the anodes of the tubes which are at high potential, and the extruded porcelain coils in the foreground are employed as insulators in the water supply lines. The center coil insulates the two tubes from each other and the other two insulate the tubes from the pump and cooling equipment. To facilitate the

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replacement of a tube, a switch is provided which when operated to the "tube change" position stops the water pumps, and removes the filament voltage of the amplifier tubes without opening the filament supply of the high voltage rectifiers. These rectifiers require several minutes to reheat, and by keeping them in operation an appreciable reduction can be effected in the time that is required to change a tube.

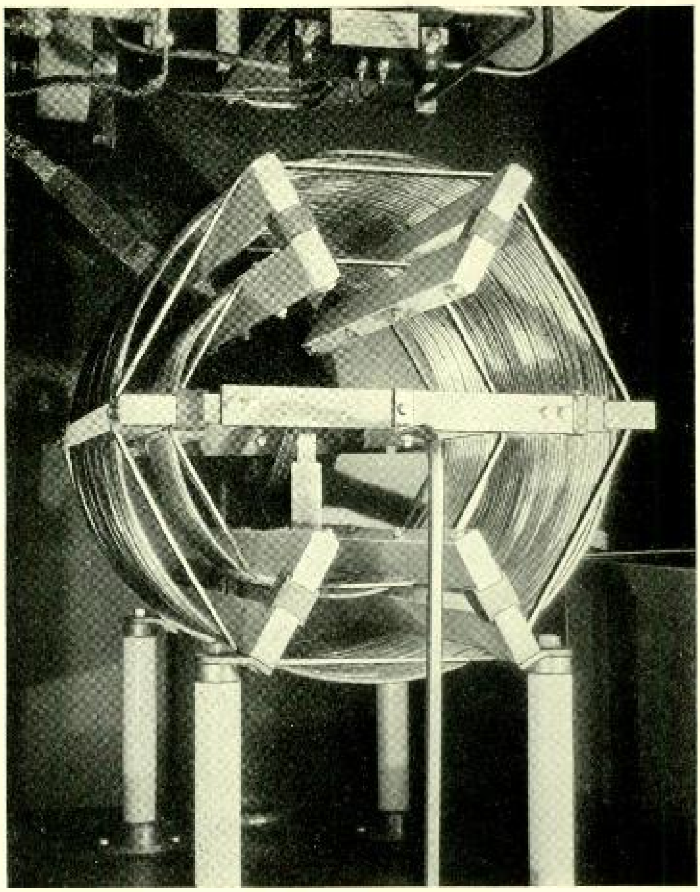

Fig. 5—Tuning unit of final amplifier

Besides these major units there is the antenna coupling unit, shown in Figure 2, which is generally mounted at the foot of the antenna. This illustration also is of a test set-up at Whippany. Two antenna-current meters are mounted on the front of the panel, one for use when operating at 1 kw., and one for operation at 5 kw. A single antenna-current meter with a double scale is mounted on the tuning unit, so that the antenna current may be read at either point.

High-voltage plate supply for the power amplifier is furnished by a six-tube mercury-vapor rectifier, which may be installed to suit the station layout. A spare tube position is provided so that a tube is heated and ready for replacing a faulty tube at all times, thus avoiding the usual heating delay. Each tube is provided with a reverse-current relay for protection, and indicating lamps associated with each relay are mounted on the front of the control unit.

One of the distinctive features of this new transmitter is the high quality of its radiated signal. The overall frequency response, as shown in Figure 3, is flat to within 1 db. from 20 to 10,000 cycles. As shown in Figure 4, audio-frequency distortion in the equipment is under 4 per cent even at 100 per cent modulation, and is approximately 1 per cent for 30 per cent modulation, which represents the average level of most broadcast programs. The weighted noise level, due to residual components introduced by rectifiers and filament supplies, is

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more than 75 db below the signal at 100 per cent modulation. The carrier frequency is maintained well within fifty cycles of the assigned frequency, and harmonic radiation on any multiple of the carrier frequency is at least 75 db below the carrier.

The high quality of this new transmitter, its automatic control features, and its flexible adjustment of power output make it suitable for a wide variety of broadcast stations. Into its design and construction has gone the fruit of many years of radio development and research. The division of the apparatus into equal size units makes it possible to meet almost any station layout. The first and second units should be adjacent to each other, and the third and fourth should also be kept together, but except for this requirement, no particular order of the units is mandatory. The control unit, for example, instead of being as shown in Figure 6, could be at the other end of the row, and other rearrangements are possible, not only of the transmitter units but of the power supply and water cooling equipment as well. Another advantage of this design is that the five-kilowatt amplifier and the control unit, together with certain modifying parts, may be added to existing one-kilowatt installations to produce a high quality transmitter of the increased output.

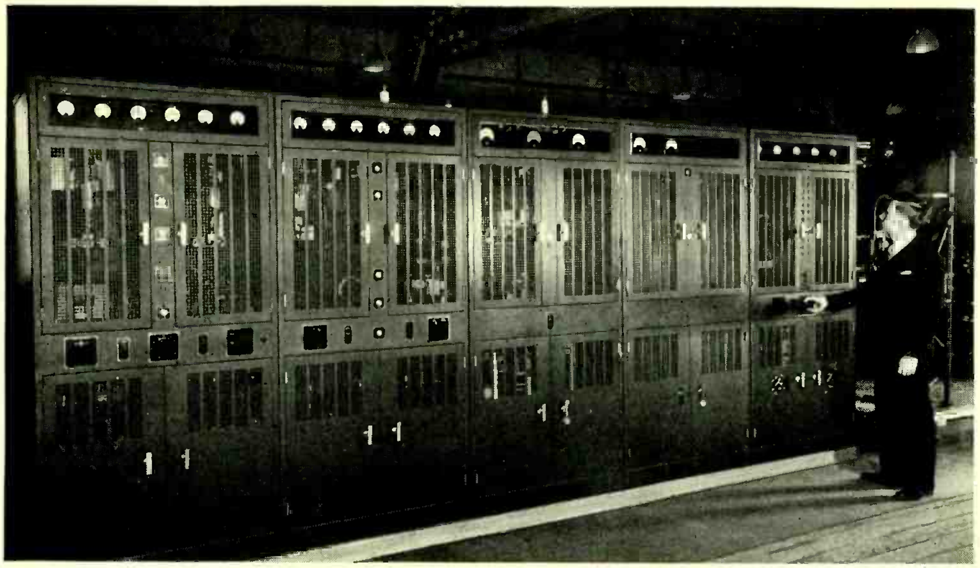

Fig. 6—Overall view of the transmitter

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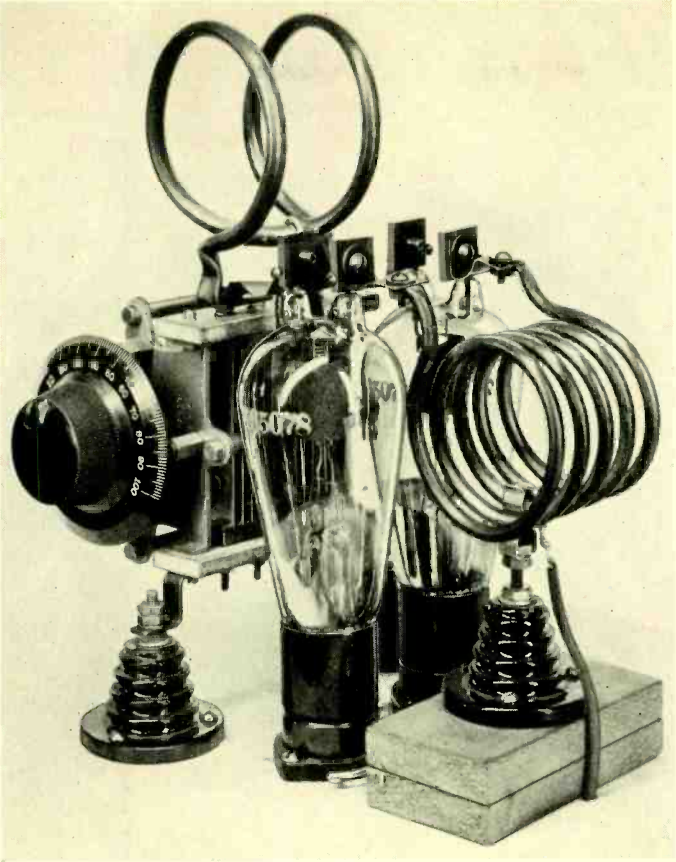

A New Vacuum Tube for Ultra-High Frequencies

By C. E. FAY
Vacuum Tube Development

IN the past few years, considerable attention has been devoted to the development of radio communication at the ultra-high frequencies. As the available channels in the lower frequency range become assigned, this high-frequency portion of the radio spectrum, including frequencies higher than thirty megacycles, offers an abundant supply of additional channels. Developments have not progressed sufficiently, however, to give knowledge of the full possibilities of these higher frequencies. To be able to carry on studies of communication systems that may employ them, it has been necessary to develop vacuum tubes that will oscillate and amplify in this ultra-high frequency region.

Difficulties have been encountered in operating the conventional vacuum tubes at these higher frequencies. One of these is a reduction in efficiency as the operating frequency is in-

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creased. For ordinary tubes, the efficiency does not decrease to any great extent for frequencies below fifteen megacycles. At frequencies somewhat above thirty megacycles, however, it begins to fall off rapidly, until a point is finally reached where the maximum allowable energy must be dissipated in the tube elements to produce any detectable output power. This is known as the frequency limit of the tube.

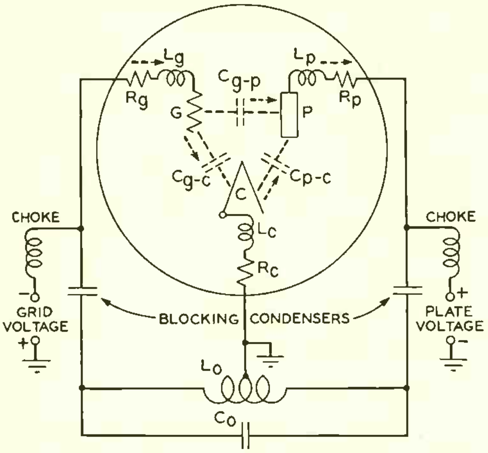

Fig. 1—Diagrammatic arrangement of vacuum tube showing external and internal reactance

One of the causes of this decrease in efficiency with increasing frequency is that the charging currents to the inter-electrode capacitances increase in proportion to the frequency. Since these charging currents must flow through the tube leads, which are not ordinarily designed to carry heavy currents, a considerable energy loss results, which decreases the useful output. These capacitances and charging currents are indicated by the dotted lines of Figure 1.

Besides its reduction caused by excessive charging current, the efficiency of a vacuum tube falls off very rapidly as the time of a period of oscillation approaches the time required for electrons to travel from the cathode to the anode. Reduction in efficiency due to this effect begins to be noticeable for most tubes at frequencies between thirty and sixty megacycles. Its most obvious cause is a lagging in phase of the plate current with respect to the plate voltage, although other and more involved effects are present.

Still another difficulty in the operation of the ordinary tubes at very high

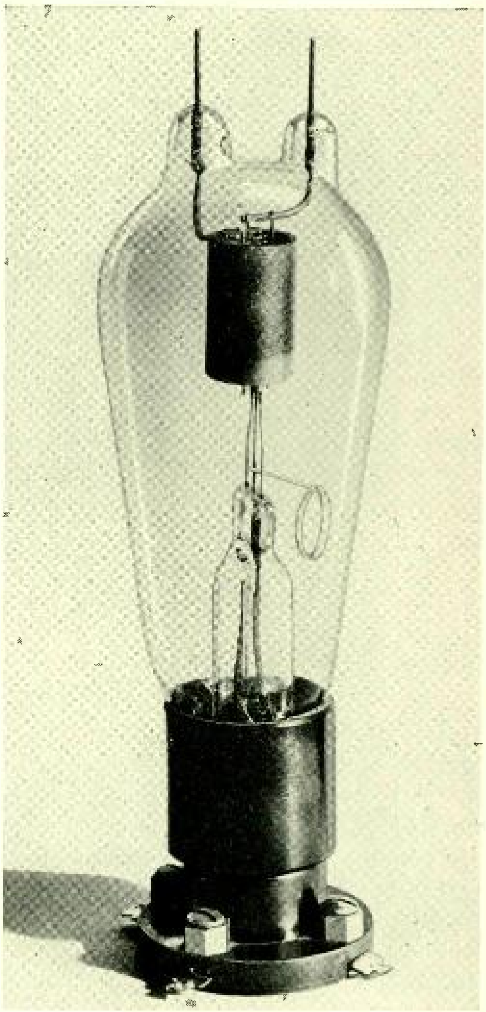

Fig. 2—The Western Electric 304A vacuum —designed for frequencies from thirty to three hundred megacycles

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frequencies is the magnitudes of the inductances and capacitances within the tube relative to the external tuning reactances. The capacitances and inductances within the tube itself are fixed in magnitude, but at ordinary frequencies are very small compared to the external reactances, such as L_0 and C_0 of Figure 1. To tune the circuit to a higher operating frequency, however, L_0 and C_0 must be made smaller, and a frequency is ultimately reached at which they become small compared to the inductances and capacitances of the tube. In extreme cases the tube reactances themselves control the oscillating frequency.

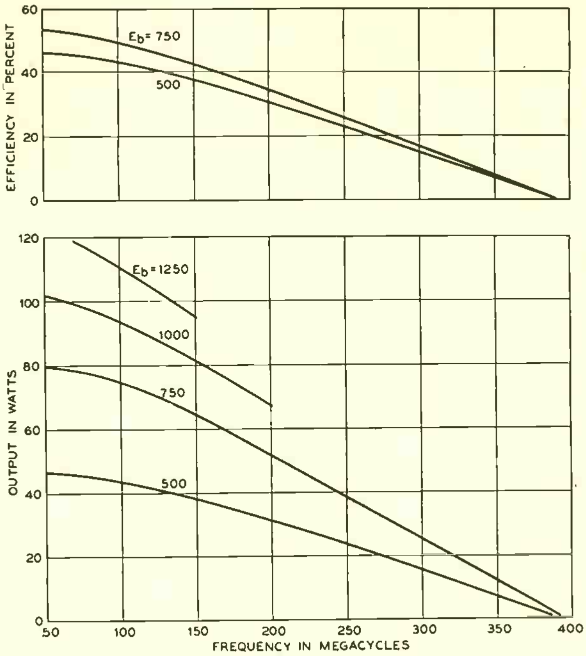

Fig. 3—Output and efficiency characteristics of 304A tube at various plate potentials

To avoid these difficulties that arise when ordinary tubes are operated at ultra-high frequencies, a tube has been recently developed in which these frequency limitations have been eliminated to such an extent that the tube is suitable for operation in the range from thirty to three hundred megacycles. This tube, known as the Western Electric 304A, and shown in Figure 2, is a low power triode suitable either as an oscillator or an amplifier. Its characteristics and ratings are given in the tabulation of Figure 4. At frequencies up to one hundred megacycles it may be operated at full rating, but with higher frequencies the output is gradually reduced. The power output and efficiency of this tube in the range from fifty to four hundred megacycles is shown by the characteristic curves plotted in Figure 3.

Several modifications have been incorporated in this new tube to make it suitable for operation at the higher frequencies. Dissipation of energy in the leads due to excessive charging current is avoided both by decreasing the inter-electrode capacitances and by decreasing the resistance of the leads. The grid and plate electrodes are supported by short heavy wires which pass through the top of the hard glass envelope. These serve both as supports and lead-in wires, and provide a construction giving the low inductance and resistance essential to the operation of the tube at ultra-high frequencies. This construction has the further advantage of eliminating any solid dielectric other than the glass envelope. In a high-frequency field, solid dielectric absorbs energy and may break down, so that its elimination is desirable.

Besides these modifications, the charging currents themselves have been made very small by employing smaller electrodes. In general, the size

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of the anode is determined by the amount of heat that must be radiated, which for any given material is a function both of its operating temperature and its radiating area. By employing graphite for the anode, which is a much better radiator than molybdenum, the material commonly employed as an anode material, it has been possible to radiate the desired amount of heat with a smaller plate. The plate is cylindrical in shape, and thus a smaller surface area makes possible a smaller diameter. The smaller diameter, in turn, results in a shorter electron transit time, and thus increases the frequency at which the phase lag of the current with respect to the voltage becomes appreciable.

With these many advantages the new tube is proving highly satisfactory for a variety of ultra-short wave circuits. A typical application is the push-pull oscillator operating at sixty megacycles shown in the illustration at the head of this article.

Filament Voltage	7.5 Volts
Filament Current	3.25 Amperes
Maximum D.C. Plate Voltage—Unmodulated	1250 Volts
Maximum D.C. Plate Voltage—Modulated	1000 Volts
Maximum D.C. Plate Current	0.10 Ampere
Maximum Continuous Plate Dissipation	50 Watts
Maximum R.F. Grid Current	5 Amperes
Maximum D.C. Grid Current	0.020 Amperes
At a Plate Voltage of 1000 Volts D.C. and Plate Current of 0.050 Amperes;	
Amplification Factor	11
Transconductance	2300 Micromhos
Plate Resistance	4800 Ohms
Inter-electrode Capacities	
Grid to Plate	2.5 Micro-microfarads
Grid to Filament	2.0 Micro-microfarads
Plate to Filament	0.7 Micro-microfarads

Fig. 4—Characteristics and ratings of the 304A vacuum tube

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Contributors to This Issue

After receiving the E.E. degree from Lehigh University in 1915, M. A. Weaver spent about six months with the Bethlehem Steel Company, and then joined the Long Lines Department of the American Telephone and Telegraph Company. Here he worked in the Long Lines Plant and Engineering Departments on equipment maintenance, toll cable testing, and toll cable engineering, until 1923, when he transferred to the D. & R. Department. Here, his work was concerned with cross-talk problems on all types of cable circuits, particularly design features and methods of installation, balancing, and testing. As a member of the Noise Prevention Department in the Laboratories he is handling problems of interference in cables at voice and carrier frequencies.

Clifford E. Fay received his B.S. degree in Electrical Engineering from Washington University in 1925, and an M.S. degree in 1927. Immediately after he joined the Technical Staff of Bell Laboratories. Here, with the Vacuum Tube Development group, he has been engaged in the development of power tubes for use in high-frequency radio transmitters.

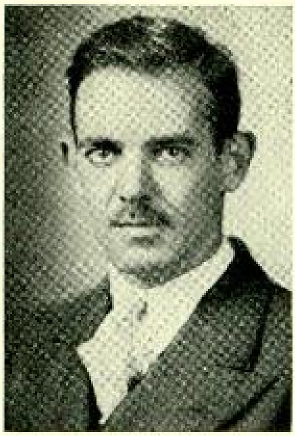
Upon receiving the B.S. degree from the University of California in 1929, R. C. Miner joined the Research Department of the Laboratories. Previous to this, however, he had gained some practical experience in the telephone industry by working for the Pacific Telephone and Telegraph Company in and around San Francisco during college vacation periods. At the Laboratories he has been engaged in the development of moving coil receivers and microphones and in methods of making acoustical measurements.

A. L. Stillwell graduated from Cambridge University in 1925 with a B.A. degree (engineering tripos). After a year with the Standard Telephones and Cables Ltd. in London, he came to the Laboratories and joined the Filter group. Since then he has been working on the development of equalizers for carrier circuits. He received the M.A. degree from Cambridge in 1929.

M. A. Weaver

C. E. Fay

R. C. Miner



A. L. Stillwell

R. E. Poole

F. B. Llewellyn

After receiving the M.E. degree from Stevens Institute of Technology in 1921, Robert E. Poole remained there as instructor in Electrical Engineering until 1924. He then joined the Radio Development Department of the Laboratories where he engaged in the design of high-power radio broadcasting transmitters. With the establishment of the Whippany Laboratory, Mr. Poole transferred his headquarters to Whippany where he has been in charge of a group testing high-power radio transmitters.

Between the years 1915 and 1922 F. B. Llewellyn alternated the study and the practice of electrical engineering. During several years he acted as a radio operator, in ship-and-shore communication, for the Marconi Company, the Navy, and the Independent Wireless Telegraph Company. In 1922 he received the M.E. degree from Stevens Institute of Technology, and after a year with the Vreeland Laboratory, he joined the Engineering Department of the Western Electric Company, now these Laboratories. Here he has since been occupied with research problems in the radio field, notably those concerned with detection, constant frequency oscillators, and ultra-high-frequency electronics. In 1928 he received the Ph.D. degree from Columbia University for studies conducted there on a part-time basis. Mr. Llewellyn was recently awarded the Morris Liebman Memorial Prize by the Institute of Radio Engineers.

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