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TELEVISION

Statement *by David Sarnoff, President of Radio Corporation of America, made on his return to New York from Europe, September 25, 1937.*

DURING my five weeks stay abroad, I studied the latest developments of television in Europe. While interest is shown everywhere in this new branch of the radio art, greater progress has been made in England than elsewhere in Europe.

Nevertheless, the experience to date with television in England has only served to emphasize the formidable nature of the problems which must be solved before a satisfactory service of television to the public can be rendered, and a new industry soundly established.

The question is often asked: "Is England ahead of the United States in television?" I shall try to answer this question by stating the facts as I have now observed them on both sides of the Atlantic.

The B.B.C. (British Broadcasting Corporation) has been operating its television transmitter, located at Alexandra Palace in London, for about a year. The range of this transmitter is more than 25 miles and covers all of London and its immediate vicinity. The system employed is known abroad as the Marconi E.M.I. Television System, which is fundamentally based on the RCA Television System first developed in the RCA Laboratories in the United States. Under an exchange of patent licenses, this British Company may use RCA patents in England and, in turn, RCA and its American licensees may use British patents in the United States.

Each side is therefore in a position to benefit from developments and improvements made by the other.

For nearly a year the B.B.C. has been broadcasting television programs to the public on a regular daily schedule of one hour in the afternoon and one hour in the evening.

Some fifteen British radio manufacturers have been offering television receiving sets to the public at prices ranging between $200 and $500 each. At the Olympia Radio Show, which I visited while in London, all the manufacturers exhibited their latest television sets, and the B.B.C. arranged special programs so that the public could view the actual operations of television while visiting the radio show. From a technical standpoint the results were highly satisfactory. The public filled the television booths and showed great interest. But while hundreds of thousands of ordinary broadcast receivers were sold during the show the public bought less than 100 television receivers in total.

During one year's operation of a public television service in England, less than 2,000 receivers in all have been sold to the trade and less than 1,000 are actually in the hands of the public. There is but one television transmitter in London, and I was informed that it will probably be two years more before a second transmitter is erected in any other part of England.

The foregoing represents the present status of television in England, despite the fact that geographically its problem is simple compared with that imposed by the vast area to be served by a television service in the United States. Also it is to be noted that in England the cost of erecting a television station, the establishment of a special organization, and the furnishing of television programs, have been paid by the Government out of license fees paid annually by the public for the privilege of listening or seeing by radio.

The range of the RCA television transmitter atop the Empire State Building, now operated by the NBC from its television studios in the RCA Building in New York City, is approximately the same as that of the B.B.C. station in London. The television receivers installed in the homes of our experts, who have been carrying on field tests during the past year, are likewise of the same order of performance as those in use in England.

The major problem of television, in both countries, is to provide a program for the home that will meet public requirements and maintain public interest.

To place television on a commercial basis in the United States, it is necessary to establish a sufficient number of sending stations, which must be interconnected and able to furnish a regular service at least to the population residing within the principal market areas of our country. The erection of such stations, the provision of necessary interconnecting facilities, and the establishment of a regular program service that would meet public requirements and hold public interest, call for vast financial expenditures before any returns can reasonably be expected.

I firmly believe in the American system of private enterprise, rather than Government subsidy; of free radio to the home, rather than license fees paid to the Government by owners of receiving sets; and I have no doubt that in due time we shall find practical answers to the practical problems which now beset the difficult road of the pioneer in television. The road calls for faith and perseverance as well as ingenuity and enterprise, but it is a road that holds great promise for the public, for artists and performers, and for the radio industry.

A NOVEL RELAY-BROADCAST MOBILE-UNIT DESIGN

By

M. W. Rife

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As a general rule the term "Mobile Unit" is applied to a vehicle carrying mobile short-wave transmitting and receiving equipment for relay broadcast work. Very often the vehicle is drafted to additional services such as the transportation of other equipment, studio properties, chairs, music stands, instruments, etc. Under these conditions it is practically impossible to maintain a mobile unit as an emergency device suitable for handling emergency special event broadcasts at a moment's notice. Other limitations suggest themselves, such as low cruising speed, and inability to drive on streets restricted to passenger vehicles.

Late in 1936 the National Broadcasting Company purchased a heavy duty passenger coach to be used as a mobile unit. This car was particularly suited to this type of work due to its striking appearance, its large roomy body and trunk, heavy chassis and speedy and powerful engine. As this is a passenger car and normally has three seats to accommodate nine persons, the two rear seats were removed at the factory to permit equipment installation. Other modifications made at the factory included strengthening of the trunk compartment to accommodate a gasoline engine-driven generator, inclusion of a special hatch cover over the front seat so that an announcer might be able to stand up and make observations above the roof of the car, side mounts for spare tires in both front fenders, addition of two concealed roof antennas for receivers, strengthening of the top at various points to permit the installation of lead-in bowls and microwave directional antenna systems, installation of additional lighting fixtures for either bright or dim interior lighting, special paint and lettering job in accordance with NBC specifications and bonding of all parts of the chassis and body where this was necessary.

The design and installation of the equipment was made by NBC personnel in Chicago where this particular unit is stationed. The equipment might be divided into three sections; namely, the front control console, the rear equipment console and the trunk power compartment. The equipment installed in these three sections is as follows:

Front Control Console

Ultrahigh-frequency superheterodyne receiver (31-41 mc).
High-gain audio amplifier and power supply.
Control panel.
Monitoring amplifier.
Automatic audio-gain control unit.
Spare power supply for receivers.
Low-voltage power pack for receivers, and
miscellaneous equipment.

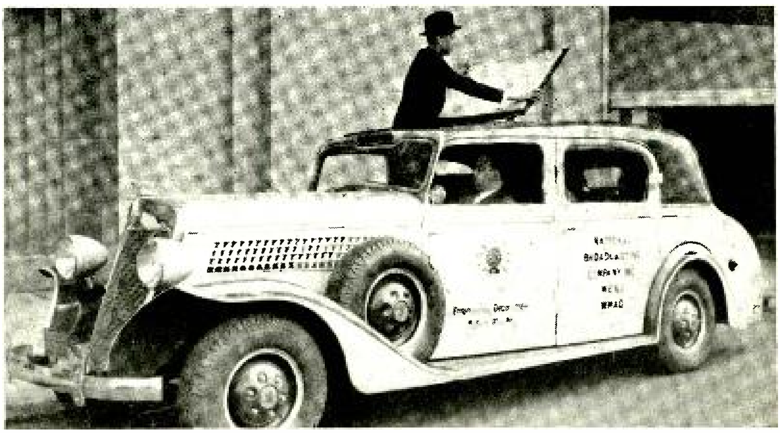

Fig. 1—The silver and blue NBC Chicago mobile unit is equipped with a hatch through which the announcer may make observations in all directions.

Rear Equipment Console

50-watt intermediate-frequency crystal-control relay broadcast transmitter (1600-3000 kc).
40-watt ultrahigh-frequency crystal-control relay broadcast transmitter (31-41 mc).
Intermediate-frequency receiver (550-30,000 kc).
High-voltage power pack for both transmitters.

Trunk Power Compartment

1000-watt, 110-volt, 60-cycle single-phase gasoline-engine-driven generator (Onan type 10L).
12-volt starting battery for gasoline-engine-driven generator.

In addition some control and monitoring equipment is mounted on the dash of the car so that it is possible for the driver to operate the equipment and announce a program while driving if an emergency requires that this be done. Propagation measurements of ultrahigh-frequency transmissions can be made by the driver without help from additional personnel.

Special consideration was given to the appearance of the equipment installation and the finishes used. In general the rear equipment and front control consoles are finished in a fine black wrinkle to match the leather upholstering, with all radio and audio equipment including the front console panel finished in a fine gray wrinkle. All trim and fixtures are black nickel-plated and a sparing amount of polished chrome is used to liven up the appearance. This combination proves

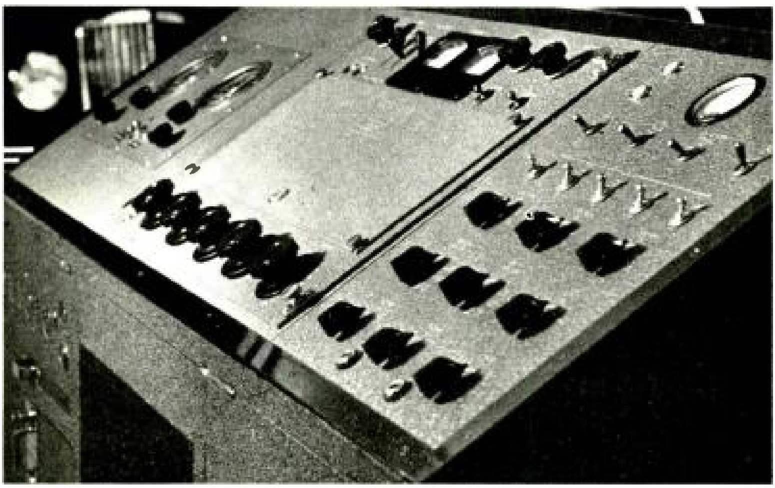

Fig. 2—The sloping panel console contains a superheterodyne ultra-high-frequency receiver, a high-gain audio amplifier and the control panel. Other equipment is mounted in the drawers below.

to be particularly effective in setting off the equipment proper, and allows the color scheme of the consoles to blend in with the upholstering and floor covering.

In order that the weight of the front control console and rear equipment cabinet be kept as low as possible, all metal framework and facing is of Dowmetal, this metal being a third lighter than aluminum. Dowmetal is easily machined and is not easily scratched due to its hardness and hence was ideal for this application. Drawers for the consoles are constructed of plywood with facings of Dowmetal sheet. Special rubber compression stops are included to prevent the drawers from rattling when the car is in motion.

All power requirements are met by the 1000-watt gasoline-engine-driven generator located in the trunk compartment of the unit. An auxiliary battery supply is included, however, for receivers and miscellaneous equipment, the filament-heating supply being obtained from the regular car battery, and the plate supply from dry B batteries. As the high-gain audio amplifier is a standard battery-operated job, no modifications were made to permit its operation from an a-c power pack as this did not seem desirable.

The front control console is mounted just behind and facing the driver's seat. Its sloping panel contains the ultrahigh-frequency

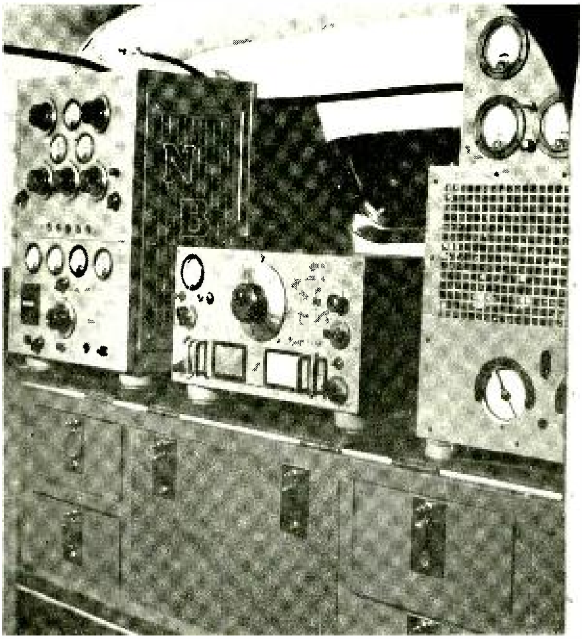

Fig. 3—The 40-watt ultrahigh-frequency transmitter, intermediate-frequency receiver and 50-watt intermediate-frequency transmitter are located on the rear equipment console. The large drawer directly under the receiver contains the high-voltage power pack for the transmitters.

superheterodyne receiver, the high-gain audio amplifier and the control panel. The receiver and high-gain audio amplifier were specially designed by the NBC laboratory in New York and are the last word in this type of equipment. This equipment is flush mounted and the top section of the console is hinged so that servicing of these units is a simple matter.

The control panel is the nerve center of the mobile unit. Here, all control of the various units takes place. Included are controls for turning on or off equipment, lights and power supplies, and switching of inputs, outputs and monitoring channels of the various equipment components. Standard toggle and ganged rotary switches are used so the use of relays was not found necessary. This is particularly desirable in a mobile unit where vibration is present, and where it is advisable to keep the amount of wiring to a minimum. A feature of the control panel is the use of rotary-type switches to switch the monitoring amplifier or headphones to the output of any of the equipment units, including the r-f monitor rectifier of the ultrahigh-frequency transmitter. A similar switch is used to place the automatic audio-

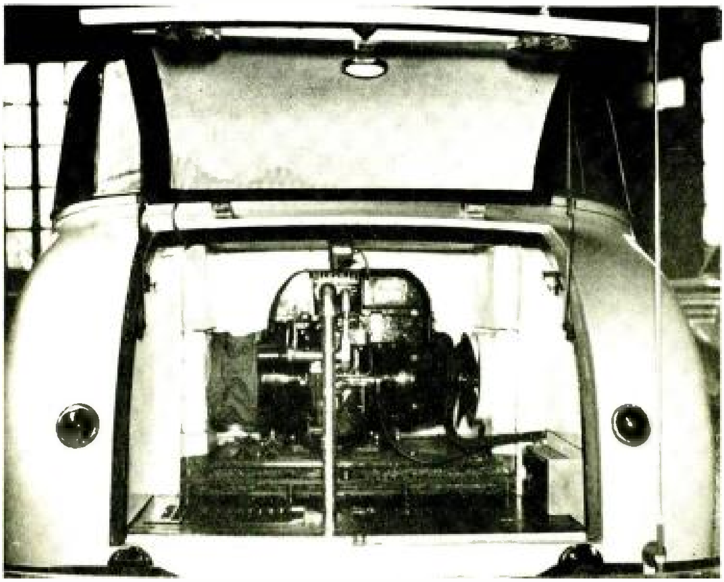

Fig. 4—The 1-kw gasoline-engine-driven a-c power plant is located in the trunk compartment. Intake ventilating ducts are located at each end of this unit with exhaust ducts in the floor. The trunk is lined with rock wool held in place by perforated metal sheet.

gain control unit in the input of either transmitter. Toggle switches are provided to switch the interior lights to dim or bright, or from the a-c supply (through transformers) to the car battery. Two different size bulbs are used in the lighting fixtures to obtain this effect.

The six drawers in the front console contain the monitoring amplifier, local battery telephone set, auxiliary battery supply, automatic audio-gain control and low-voltage a-c power pack for receivers and miscellaneous equipment. All of this equipment is mounted on sub-bases bolted to the bottom of the drawers which makes this gear easily accessible. Inputs, outputs and power connections are made through twist-lock plugs. Drawer space in which equipment is not mounted is utilized by providing additional sub-bases with sockets for spare tubes and crystals. This eliminates the usual drawer full of loose tube cartons and facilitates keeping inventory of these items. Coils for the intermediate frequency receiver are contained in a special rack mounted in one of the drawers and two additional drawers are available for miscellaneous cords, plugs, microphone extensions, headphones and other loose equipment. A writing shelf similar to those used in most desks is included above the drawer space and just below the sloping panel. The seat for the engineer operating this equipment is a swivel stool with back rest. This seat is normally bolted to the floor, but can be easily removed if desired.

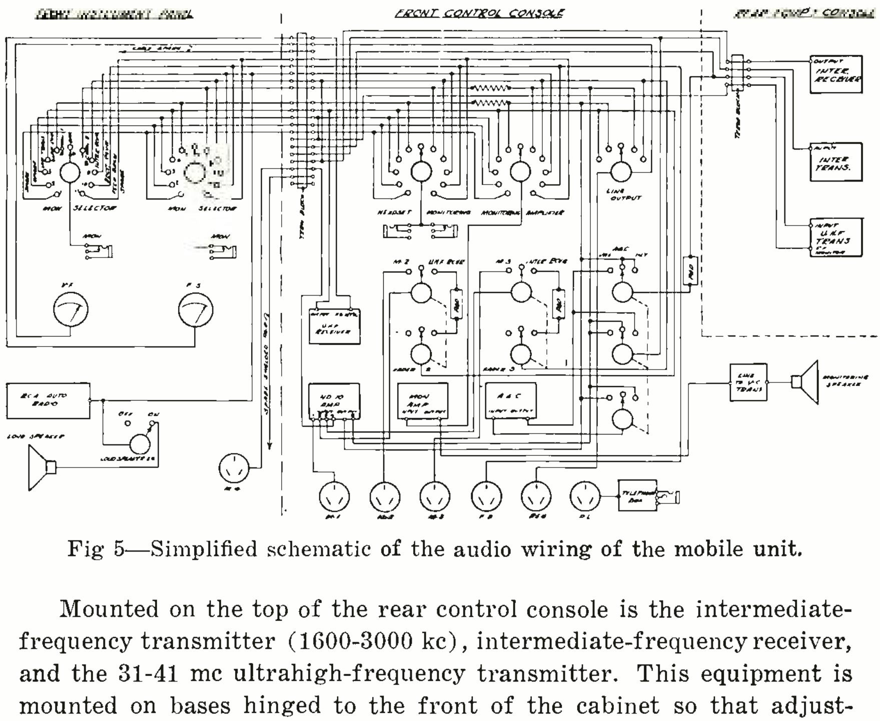

Fig 5—Simplified schematic of the audio wiring of the mobile unit.

Mounted on the top of the rear control console is the intermediate-frequency transmitter (1600-3000 kc), intermediate-frequency receiver, and the 31-41 mc ultrahigh-frequency transmitter. This equipment is mounted on bases hinged to the front of the cabinet so that adjustments or servicing might easily be accomplished merely by tilting each unit forward. The top of the cabinet is covered with black battleship linoleum and trimmed with narrow, white-metal beading. The largest of the five drawers contained in this cabinet houses the high-voltage power pack from which both transmitters may be operated simultaneously. This unit is adequately shielded with iron to prevent any electromagnetic disturbance from interfering with equipment located on the top of the cabinet. The other four smaller drawers are used as storage space for miscellaneous equipment.

The design of the trunk power plant compartment was more difficult than it would appear. Soundproofing, isolation of the gasoline engine generator from the chassis to prevent engine vibration reaching the interior of the car and removal of heat generated by the air-cooled engine were the three major problems involved. The sound-proofing problem was taken care of by lining the interior of the trunk with rock wool blanket held in place by perforated metal sheets. This was found to be more than adequate, and the unobjectionable noise in the interior of the car caused by the car engine was found to be greater than that generated by the power plant. Considerable experimental work was necessary to remove vibration caused by the one-

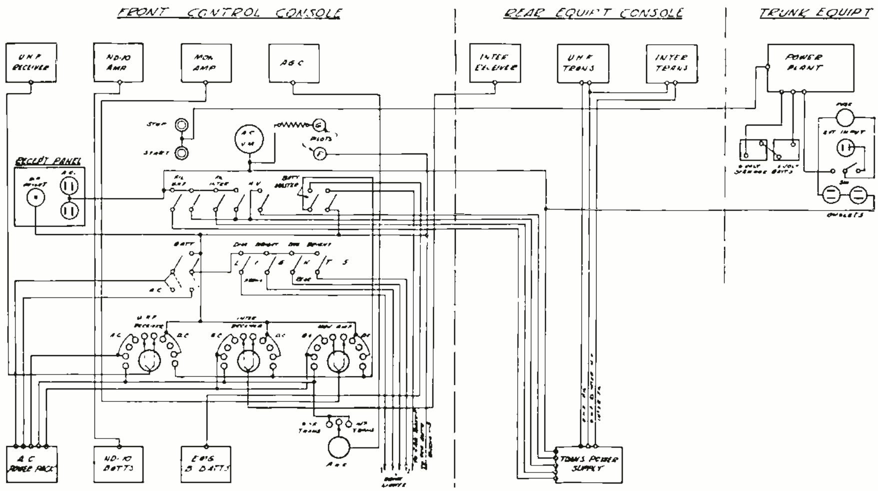

Fig. 6—Simplified schematic of the power wiring of the unit.

cylinder engine reaching the chassis and carrying through to the interior of the car. Various commercial isolating devices were tried, but the type in which the engine is supported in rubber suspension mounts was found to be the most satisfactory. These units were mounted directly to a heavy framework bolted to the chassis of the car. Adequate cooling for the air-cooled engine was provided for by including a forced filtered system of ventilation in which all intake and exhaust ducts go through the bottom of the trunk compartment, thereby eliminating unsightly grills or ventilators in the top or sides of the trunk compartment. Spun glass filters are used to prevent road dust from getting into the engine, these being blocked by baffles when the power plant is not in use. The muffler used on the power plant is mounted under the bottom of the trunk compartment and is separate from that used for the car engine. Provision is also made for the use of external a-c power obtained from city feeders when this supply is available and assignment conditions permit.

Included on the dash of the car are rotary monitoring selector switches and jacks for headset monitoring for announcer and driver, automobile radio controls, a meter for field-strength measurements, six-volt d-c electric clock and a compass. An RCA automobile radio of the latest type is mounted under the dash on the firewall. Headphone monitoring is provided as well as a switch for the loudspeaker. A special arm for holding an RCA inductor microphone is provided for the announcer which may be removed when not in use.

Four antennas are included on the unit at the present time. A concealed roof antenna is used for the intermediate-frequency receiver; a running-board antenna is provided for the automobile radio; while a quarter-wave vertical whip antenna mounted on the rear bumper is used for the ultrahigh-frequency transmitter. The transmitting antenna is insulated from the body of the car and fed by standard ⅜-inch, 72-ohm concentric cable. A similar whip antenna mounted on the front bumper is used with the ultrahigh-frequency receiver. No permanent antenna is provided for the intermediate-frequency transmitter; however, installation of a collapsible top-loaded vertical antenna for this purpose is contemplated in the near future.

Removal of interference to receivers caused by other electrical gear in the mobile unit will not be covered here as each problem was treated separately and nothing of an unusual nature presented itself.

Although this mobile unit has been in service but a short time it has proven its worth on several occasions. The advantages of using a speedy unit having permanently mounted equipment set up in a manner that provides the utmost flexibility in operations is readily apparent. As an emergency device that must be called upon to function at a moment's notice this type of mobile unit meets those requirements of relay broadcast work where public interest, convenience, and necessity must be met.

SOME UNCONVENTIONAL VACUUM TUBE APPLICATIONS

By

F. H. Shepard, Jr.

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I. Introduction

IT IS becoming more generally known that vacuum tubes have many applications as means of doing things which, without tubes, have been done with difficulty, inconvenience or not at all.

It is also becoming generally realized that, when vacuum tubes are used conservatively with due regard for the tube characteristics, and when the same factors of safety are used as are generally used for other mechanical or electrical devices, the reliability of the vacuum tube will in many cases surpass the reliability of the mechanical or electrical device.

It seems needless to say that, to be entirely successful and economical, the electronic device should be at least as simple, and as cheap, as the electrical or mechanical device which it replaces, or that it should do the job more efficiently. In spite of this, we sometimes find ourselves working out complicated electronic circuits to perform a function that can be done by some relatively simple electrical or mechanical device.

The following paragraphs show some devices that have been used to advantage in many applications:

II. A Self-Balancing Capacity-Operated Relay

The author has described several capacity-operated relays[1, 2] which operate on absolute values of the antenna-to-ground capacity. These circuits are sensitive and stable, and are suitable for use where it is desired to operate on a definite value of capacity.

There is, however, a great field of application where it is desired to operate the relay only on a change of capacity occurring within a

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[1] Miscellaneous Applications of Vacuum Tubes, June, 1935 *Proceedings of the Radio Club of America.*

[2] Application of Conventional Tubes in Unconventional Circuits, *Proc. I.R.E.*, December, 1936.

period of several seconds duration. In such applications the actual antenna-to-ground capacity may drift considerably over long periods of time due to temperature and humidity changes.

The circuit shown in Figure 1 will automatically rebalance itself after a short period of time, to compensate for the slow variations of the antenna-to-ground capacity within a certain range.

In this circuit operating on the a-c line, the sensitive element consists of a pentode oscillator, the feed-back of which is determined by the difference in the ratio between the inductance of the two parts of the oscillator coil L_1 and L_2 and the ratio between C_1 and the antenna-to-ground capacity. Because the cathode of the oscillator is at an r-f

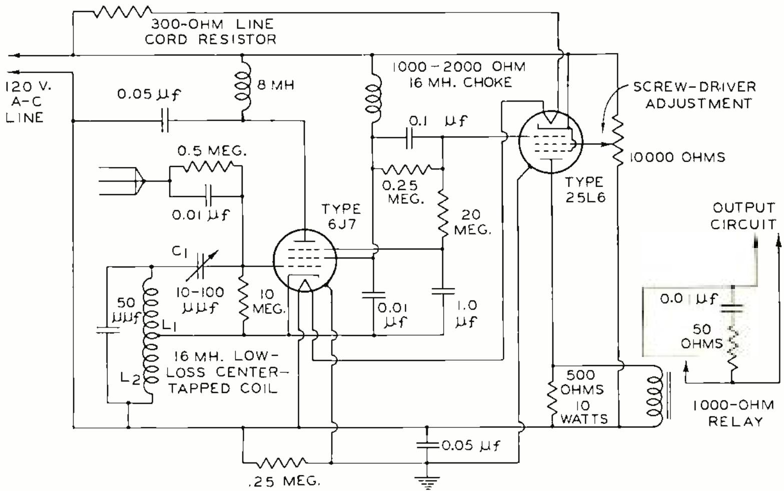

Fig. 1—A self-balancing capacity-operated relay.

potential, and because the control grid of the output tube is by-passed for high frequencies through suitable by-pass condensers to the cathode of the oscillator, a negative d-c voltage equal to the peak r-f voltage on the cathode of the 6J7 is built up across the grid-leak and condenser of the 25L6 due to the rectifying action of the grid. The 6J7 oscillator oscillates at high frequency on one-half of the a-c cycle and builds up the above-mentioned negative charge on the grid of the output tube. During this time, the output tube has negative plate and screen voltages and so is non-conducting. On the other half of the a-c cycle, the 6J7 oscillator has negative plate and screen voltages and so ceases oscillating. The negative charge built up on the grid of the output tube does not leak off during this interval and, hence, is effective in controlling the plate current of the output tube during its positive plate-voltage interval.

The sensitivity of the unit is an inverse function of the quality of the oscillator coil: that is, the lower the losses in the oscillating circuit, the greater will be the sensitivity. Theoretically, as the oscillator plate impedance approaches infinity and the losses in the antenna and the tuned circuit approach zero, the sensitivity approaches infinity. Sensitivity gained in this manner is much more stable than that gained by regeneration.

The sensitivity of this circuit to small changes in capacity can be increased regeneratively by increasing the resistance of the choke coil feeding the screen of the 6J7. When this resistance is more than about 15,000 ohms, the circuit will become unstable, i.e., the relay will relax.

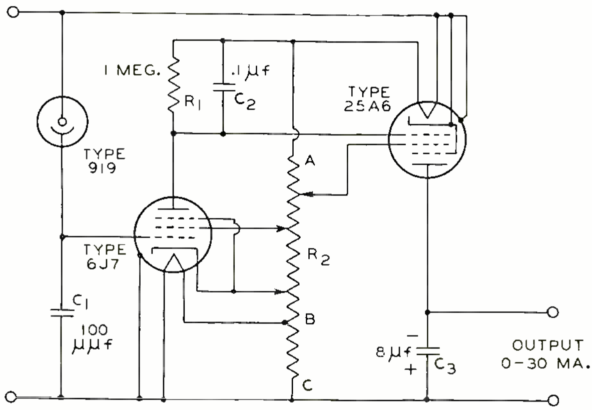

Fig. 2—A simple two-stage a-c operated photo-amplifier relay.

The increase in sensitivity is caused by the fact that as the intensity of oscillation of the 6J7 increases, its plate and screen currents decrease. This causes the screen voltage and, hence, the mutual conductance of the 6J7 to increase; the increase, in turn, helps to increase the intensity of oscillation. Chattering of the relay can be effectively eliminated by the damping resistor placed across the relay coil. If desired, the resistor can be replaced by an electrolytic condenser large enough to keep the relay from chattering, (8 microfarads for most relays).

The intensity of oscillation is automatically controlled by feeding a negative potential to the No. 3 grid of the 6J7. This potential makes the grid negative with respect to the cathode, reduces the mutual conductance of the tube, and reduces the plate impedance of the tube which is effectively shunted across the lower part of the tuned circuit. Both of these actions tend to reduce the amplitude of oscillation for a given feedback, or to make the oscillator require a greater feedback for a given intensity of oscillation. As the negative potential applied to the No. 3 grid of the 6J7 is derived through a time-delay circuit from the grid of the output tube, this grid, as explained above, assumes a direct potential equal to the peak high-frequency voltage appearing across the lower half of the oscillator coil, L_2. Thus, the intensity of oscillation is maintained substantially constant for wide variations in antenna-to-ground capacity (feedback). The above is true for slow capacity variations; however, sudden variations cause the intensity of oscillation to vary rapidly before the corrective voltage can be applied through the condenser-resistor time-delay circuit to the No. 3 grid of the 6J7. Hence, short-time capacity variations are effective in operat-

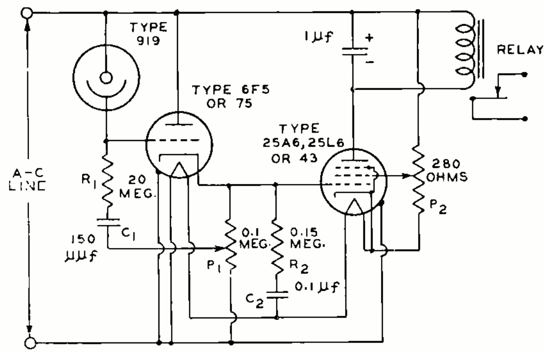

Fig. 3—A quick-acting a-c operated photo-amplifier relay functioning in a positive direction on the received light.

ing the relay while long-time variations within limits are automatically compensated for.

This type of circuit finds application in connection with door openers, counters, burglar alarms, advertising displays, etc., and has even been used as a foul-line indicator for bowling alleys.

III. A Simple Two-Stage A-C-Operated Photo-Amplifier Relay Circuit[2]

Figure 2 shows a simple two-stage photo-amplifier relay circuit operating directly on the a-c line. The simplicity of the circuit is illustrated by the fact that the complete list of circuit parts includes only one voltage-divider resistor, one plate-load resistor, and three condensers. The circuit shown in Figure 2 consists of a high-impedance phototube feeding through a voltage amplifier or buffer stage into a power-output stage. The filament voltage of the buffer stage has been lowered to reduce the temperature of and, hence, the electron emission from the grid of the buffer tube. The plate current of the buffer stage is kept at a minimum in order to reduce the electron bombardment of the gas molecules within the tube and, hence, the gas current to the grid. The bias to the grid of the buffer stage is obtained by means of the rectifying action of the grid itself. This method of obtaining the grid bias keeps the effective bias and, hence, the plate current of the tube constant, regardless of large fluctuations in contact potential between the grid and the cathode. The impedance of the condenser C_1 acts as a load impedance for the phototube. Condenser C_1 is charged up to a definite negative potential on one-half of the a-c cycle and is allowed to discharge through the phototube on the other half of the cycle. The amount that it is discharged by the phototube determines

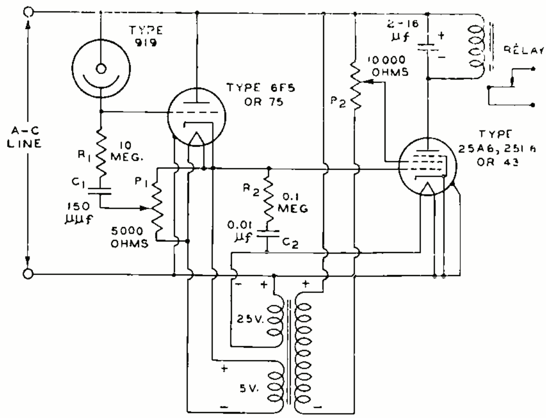

Fig. 4—A quick-acting non-degenerative a-c-operated photo-amplifier relay functioning in a positive direction on the received light.

the working potential on the grid of the buffer stage. The size of C_1 can be set to any desired value to determine the desired sensitivity range of the relay.

IV. Quick-Acting A-C Operated Photo-Amplifier Relay

Figures 3, 4, and 5, show three variations of an a-c-operated photo relay that will respond to a pulse of light having a duration as short as one-sixtieth of a second, or 1/120th of a second if the pulse is properly phased with the power-supply voltage.

The operation of these circuits is essentially as follows:

On the negative half of the power-supply cycle with no light on the photocell, the cathode of the 6F5 goes negative with respect to the grid, passes current to the grid, and thus during a period of several cycles charges C_1 to a value equal to the peak value of the a-c voltage applied between the grid and cathode. This voltage added to the instantaneous a-c voltage applied between the grid and cathode is sufficient to reduce the plate current of the output tube to zero. When light is received by the phototube, the phototube current has two effects, first, an instantaneous drop appears across the phototube load R_1, and second, the phototube current into condenser C_1 opposes the current fed into C_1 from the grid of the amplifier. This action causes the potential across C_1 to balance at some negative value between zero and a value equal to the peak of the a-c voltage between grid and cathode. The potential across C_2 is fixed in like manner by a balance between the charging grid current and the discharging buffer-tube current. In Figures 3 and 4, an instantaneous flash of light occurring on the

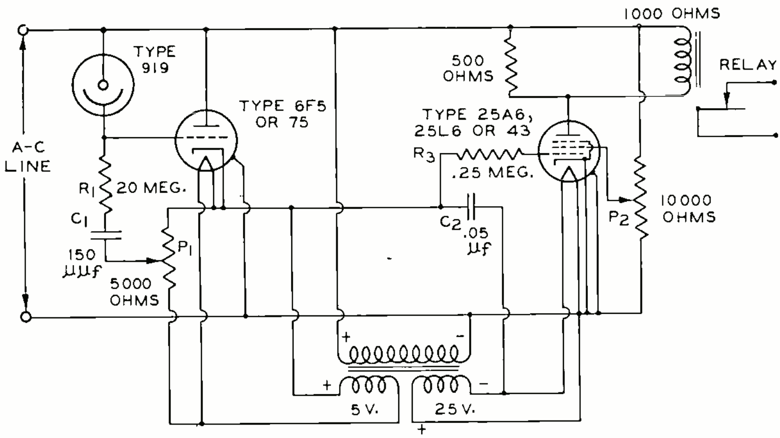

Fig. 5—A quick-acting slow-releasing a-c-operated photo-amplifier relay functioning in a positive direction on the received light.

positive half of the cycle will cause an instantaneous drop across R_1, an instantaneous change in the plate current of the buffer stage, and an instantaneous change in grid voltage of the output tube. In Figure 5, due to the current taken by the buffer stage, there is a rapid loss of potential across C_2. This loss is slowly restored through R_3 over a period of several cycles. Thus the pulse of output current will have sufficient duration to operate a sluggish mechanical relay, even though the pulse of light be of extremely short duration.

The circuit in Figure 4 is identical to the circuit shown in Figure 3 except that the heater voltages are supplied by suitable transformers. The use of a transformer supply for the heaters eliminates the degeneration or loss of gain that is inherent in the circuit shown in Figure 2, due to the fact that the d-c potential between the lower side of C_1 and the cathode of the buffer tube does not remain constant. In this circuit care should be taken to keep the instantaneous transformer polarities as shown.

These simple photo relays find their application for use in moderately high-speed sorters, counters, register controls, shooting galleries, etc.

V. A Sensitive Light-Intensity Indicator[2]

Figure 6 shows a sensitive photo-amplifier circuit that can be used for accurately matching the intensities of amounts of light. With this circuit, it is easily possible to indicate light differences or changes which may amount to small parts of one per cent. In this circuit arrangement, the high-impedance 954 pentode acts as a load impedance for the 919 high-impedance vacuum-type phototube. It can be seen by reference to Figure 7 that the potential of the common connection between the 919 and the 954 is determined by the intersection of the 954 and 919 characteristics. It is also evident that a small change of light on the phototube will result in an output of several volts. This output voltage is applied to the grid of a 38 output tube, the plate

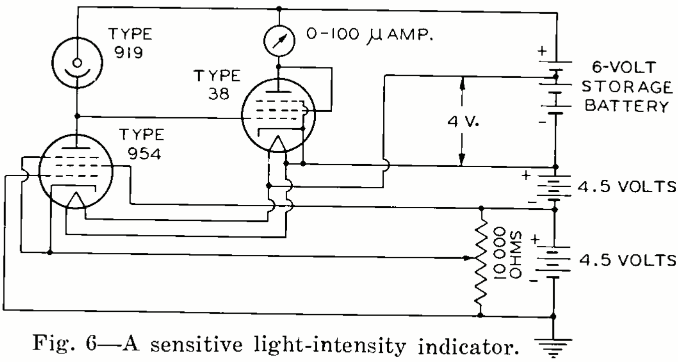

Fig. 6—A sensitive light-intensity indicator.

current of which is indicated on a 100-microampere meter. Because the phototube with the 954 load has an extremely high output impedance, it is necessary to operate the 38 so that its grid-input impedance is extremely high. To reduce the grid emission to a minimum, the voltage to the heaters of the 38 and the 954 is reduced to 4 volts. The possibility of emission from the heaters to the grid is eliminated by operating the heaters at a potential positive with respect to the plate of the 954 and the grid of the 38. Gas current to the grid of the 38 is kept at a minimum by keeping the potentials within the 38 low so as to minimize the ionization of any gas that may be in the tube. Because all of the high-impedance external connections are made to electrodes brought out from the tops of the tubes, external leakages are reduced to a minimum. External leakage can be greatly reduced by carefully cleaning the tubes and coating them with a non-hygroscopic wax. This can be done by dipping the tubes in hot ceresin wax and holding them under the surface of the wax until the greater part of the moisture on the glass is boiled off. Care should be taken not to scorch the wax. Ceresin wax has long been used by research men for reducing the effects of moisture leakages in high-impedance d-c circuits. Dr. Rentschler of the Westinghouse Lamp Company very kindly made available to the writer information on the use of ceresin wax.

This circuit finds application where it is desirable to indicate very small percentage variations of an amount of light. For instance, it can be used to indicate the absorption of light by a fluid and, consequently, to indicate or control the concentration of certain chemicals in suspension or solution. The use of monochromatic light can be used to advantage when it is desired to isolate a particular constituent. This circuit also finds application in color matchers and in indicating small changes of small amounts of light. For instance, when measuring small changes of small amounts of light, it has been demonstrated

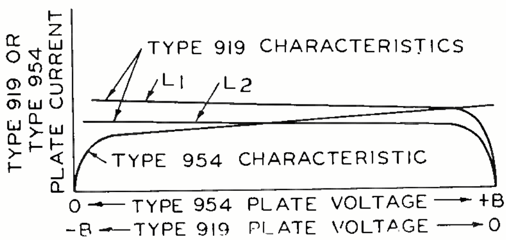

Fig. 7—Plate characteristics of a vacuum-type phototube with a pentode as a plate load.

that a change of light intensity on the order of two-millionths of a lumen is sufficient to swing the output meter over its full scale.

VI. A Variable-Range Variable-Sensitivity Light-Variation Indicator[2]

It is sometimes desirable to make the sensitivity of the light-intensity meter indicator less for small percentage changes of light. The sensitivity can be reduced to any desired degree by varying the plate characteristics of the 954 between those of a pentode and those of a triode. This variation is produced in the arrangement shown in Figure 8 by properly adjusting P_2 and P_3 to control the relative potentials on the control grid and the screen grid. When the No. 2 grid of the 954 is positive with respect to the cathode, the 954 has a high-impedance pentode characteristic. Changing the No. 1 grid bias changes the height of the characteristic as shown in Figure 9 by the curves No. 1 and No. 2. As the potential of the No. 2 grid is made more negative, the characteristic of the 954 changes to that shown by curve No. 3. With zero bias on the No. 1 and the No. 2 grids, the characteristic curve is as shown by curve No. 4. When a negative bias is placed on the No. 2 grid, the shape of the characteristic is unchanged, but it is shifted along the voltage axis as shown by curve No. 5. If the No. 1 grid is biased properly, the slope of the characteristic is increased. Curve No. 6 shows the effect of positive bias on the No. 1 grid (bias on the No. 2 grid for this curve is zero). Placing a negative bias on the No. 2 grid shifts this characteristic along the axis as shown by curve No. 7.

From this analysis it can be seen that the 954 phototube load can be adjusted to give practically any desired positive impedance load at any desired current and at any desired voltage across the tube. This means that the full-scale reading of the output meter can be made to cover a fraction-of-a-per cent light variation, a 100 per cent light variation, or any desired amount of variation between these two

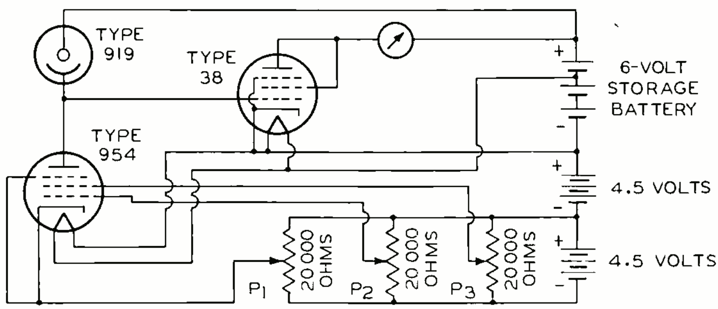

Fig. 8—A light-intensity indicator having variable range and variable sensitivity.

extremes. A photo-amplifier such as this finds application as a densitometer for use in connection with the analysis of photographically recorded spectra, and for use in connection with a suitable monochrometer or light filter as a means of measuring the absorption lines or the concentration of certain chemicals in solution. These are but two of a large number of possible applications for this type of circuit.

VII. A Photo-Electric Exposure Control

Figure 10 shows a circuit that will act to operate a relay or a solenoid when the phototube has received a certain predetermined amount of light after switch S_2 has been closed. This action is independent of the time over which the light has been received, providing the time is greater than about ½ a second. In practice the phototube can be arranged to receive the light reflected from or transmitted through a part or the whole of the plate, film, or paper in the camera.

When switch S_2 is open the plate and screen voltages to the output tube are zero, and the output relay is de-energized. One side of the a-c line is applied through one of the timing condensers to the grid, while the cathode is connected to the other side. Due to the rectifying action between the grid and cathode of the buffer stage the condenser assumes a charge, the direct potential of which reaches a value approaching the peak of the a-c line voltage. On closing switch S_2 the d-c voltage or the major portion of this voltage is removed from between the grid and cathode of the buffer stage, thus allowing the above mentioned d-c potential on the grid to cut off the plate current. This in turn allows the voltage drop across the plate load of the buffer stage to drop to zero. As this drop across the buffer stage plate load is both signal and bias for the output stage, and as the plate and screen voltages of the output tube are applied by closing switch S_2, the plate current of the output tube will rise to operate the output relay or

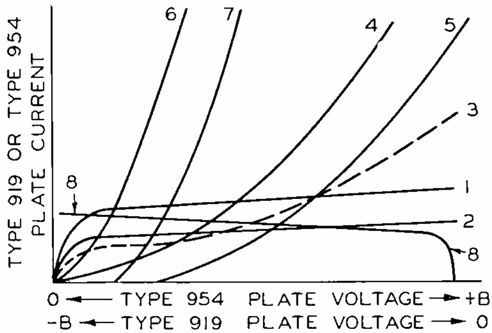

Fig. 9—Plate characteristics of a vacuum-type phototube with a pentode having a controlled plate impedance as the phototube load.

solenoid. This in turn lights the exposure lamp or opens the camera shutter. After the charge across the timing condenser is dissipated through the phototube to a low enough value so that the grid of the buffer stage permits the buffer stage to conduct plate current, a voltage will be developed across the buffer-stage plate-load resistor of sufficient value to cut off the plate current of the output tube. This action releases the relay or solenoid, and thereby turns off the exposure lamp or closes the camera shutter.

Theory:—The current passed by a vacuum-type phototube is directly proportional to the intensity of the light received by the phototube, and is practically independent of the voltage across the tube. The quantity of light received is intensity times time, and is in general a measure of the proper exposure. The quantity of electricity passed by the photocell is proportional to its current times time, hence, the quantity of electricity passed is a measure of the quantity of light received. As it takes a definite quantity of electricity to discharge a condenser from one potential to another, the size of the condenser and the voltage through which it has to be discharged can be used as a measure of the desired exposure. It should be noted that judgment of the operator is still necessary to determine the setting of P_1 which compensates for the percentages of light and dark areas in the picture.

This particular circuit finds its application in all types of copy work, and in all types of picture work where time exposures are used. For instance the exposure of zinc plates in newspaper work, the exposure in making microphotographs, in portrait work, in photostat work, in printing, etc., can all be controlled by this device.

VIII. A Sensitive Light-Balance Indicator

Where it is desired to indicate accurately the balance between two amounts of light, the circuit shown in Figure 11 is of value. This circuit is capable of indicating a light unbalance of less than ¼ of 1 per cent when the light on the phototubes is as low as 0.0001 lumens; the accuracy of balance is somewhat better with larger amounts of

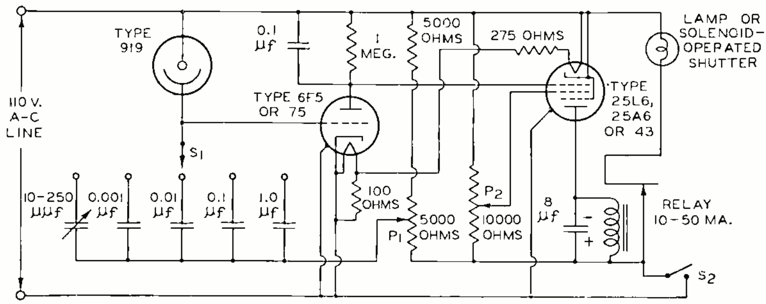

Fig. 10—A light-quantity metering and control circuit.

light. All of this is done by using the unregulated a-c power lines for the power supply.

The circuit uses a 6E5 electron-ray tube as the balance indicator, a 38 as an a-c operated buffer stage and electrometer, and a 6H6 as a rectifier to supply direct current to the 917 and 919 phototubes. One phototube acts as the high-impedance load for the other, so that relatively high voltage outputs are available, even for very small percentage variations of light. The high-impedance output of the phototube bridge circuit is fed directly to the grid of the 38. Because the cathode of the 919, the anode of the 917, and the grid of the 38 come out of the tops of the tubes, the connecting wires touch nothing but tube caps and the tubes can be coated with a suitable non-hygroscopic wax such as white ceresin wax to reduce to a minimum all external leakages. Low leakage is essential when measuring small amounts of light. The type 38 acts on the alternating voltage supplied to its plate and screen in such a manner that the average direct potential built up on its plate is negative, and is of such magnitude that it can be used as bias and signal to the grid of the 6E5 indicator tube. As a means of increasing the input resistance of the 38, the heater is operated at reduced voltage, the screen at about 10 volts, and the plate at about 18 volts. To avoid the effects of emission from the heater to the grid, the heater is operated at a potential at all times positive with respect to the control grid.

In actual use a system of mirrors or reflecting surfaces are arranged to take the light from a common source and pass it through or reflect it from the surface of the sample under test. A calibrated shutter, a pair of rotatable calibrated polarized discs, or a calibrated runway for the light source, can be used to determine accurately the change in balance intensity as various samples are placed between the light source and one of the phototubes.

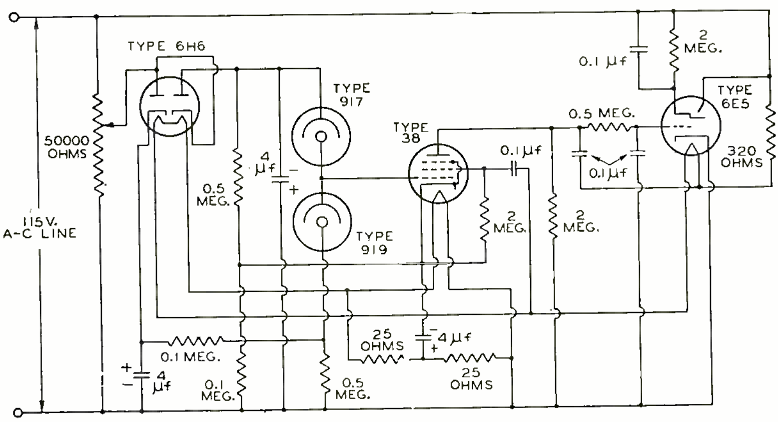

Fig. 11—A sensitive light-balance indicator.

With a suitable light source having the proper wavelengths, this device can be used for color matching, turbidity measurements, reflectance measurements, and the absorption analysis of solutions. It is generally known that all substances in solution have certain spectral absorption lines, and that the measurement of the amount of light of a particular wavelength that is absorbed by a solution, is an indication of the concentration of a particular constituent in the solution. With a source of monochromatic light of the desired wavelength, the above circuit has been successfully used for accurately determining the sugar concentration in beverages, also the concentration of various syrups and flavorings. This arrangement has also been used commercially[3] for determining the vitamin concentrations of vitamin bearing oils.

[3] Ronald L. McFarland, J. Wallace Reddie, and Edward C. Merrill, A New Photo-electric Method for Measuring Vitamin A. Presented before the American Chemical Society at Chapel Hill, North Carolina, March 22, 1937, and repeated in the July 15, 1937, analytical edition of "Industrial and Engineering Chemistry."

FIELD STRENGTH OBSERVATIONS OF TRANS-ATLANTIC SIGNALS, 40 TO 45 MEGACYCLES

BY

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Summary.—*The results of daily observations at Riverhead, N. Y., since the middle of January, 1937 are reported. Some of the schedules of London and Berlin television transmitters are reported as being heard, and measurements of field strengths are summarized. The vertical angle of arrival was measured, and by means of a reversible directive antenna it was determined that the signal at times arrives from the reverse direction over the longest way around the world.*

THIS paper will report briefly on the results of a series of observations started at Riverhead, N. Y., January 11th, 1937 on the frequencies of the television transmitters at Alexandra Palace, London, and which later included also the frequencies of the television transmitters at Berlin.

London was understood to have a sound channel on 41.5 megacycles per second, with a power rating of 3 kw and a vision channel on 45 megacycles with a rating of 5 kw. The Berlin transmissions consisted of a sound channel on 42.5 megacycles and a vision channel on 44.3 megacycles. The transmitting antennas were vertically polarized. The distances involved were 3400 miles for London and 3900 miles for Berlin.

Most of the observations took place between 1000 and 1100 E.S.T. Observations were, however, also made at other hours between 600 and 1700 E.S.T. The observations were at first made at the Frequency Measuring Laboratory of the Riverhead Station. They were later extended to another site where special antennas could be erected.

To facilitate the design of an antenna some measurements of the vertical angle of arrival were made. For these measurements, three horizontal dipoles were erected at 16.7 feet, 27.3 feet and 50 feet above ground. Figure 1 shows how these antennas were arranged. By comparing the strengths of the signals picked up on each of these dipoles the vertical arrival angle was determined, according to the method

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described by Friis, Feldman and Sharpless.[1] In order not to introduce errors due to transmission-line losses and standing-wave patterns, the transmission lines from the dipoles were made of equal lengths. A receiver was mounted in the survey car shown, which could be parked near the antennas. The three transmission lines passed to the receiver through a plug and jack arrangement providing rapid change from one antenna to another.

Fig. 1—Three horizontal dipoles and survey car containing receiver used for vertical arrival angle measurements.

A number of measurements made showed that the vertical arrival angle of the signals heard was close to 7.5 degrees. A horizontal rhombic antenna was then constructed so as to have its maximum lobe towards England at this angle. Its effective height was about 8 meters.

As the observations made at the Frequency Measuring Laboratory indicated that possibly the signal was arriving along paths other than the shorter arc of the great circle from England to Riverhead, it was decided to arrange the rhombic antenna in such a fashion that its direc-

[1] "The Determination of the Direction of Arrival of Short Radio Waves", H. T. Friis, C. B. Feldman, and W. M. Sharpless, *Proc. I.R.E.*, Jan. 1934.

tion of reception could be reversed. This was done by installing at each end of the antenna remotely controlled double-pole double-throw switches. From the blades of these switches transmission lines of equal length were run to another remotely controlled double-pole double-throw switch and from the blades of this latter switch a transmission line was run to a receiver. Figure 2 shows a diagram of the antenna and the way in which these switches were connected. The control circuits of the switches were connected so that by operating a toggle switch at the receiver it was possible to connect the receiver to either end of the

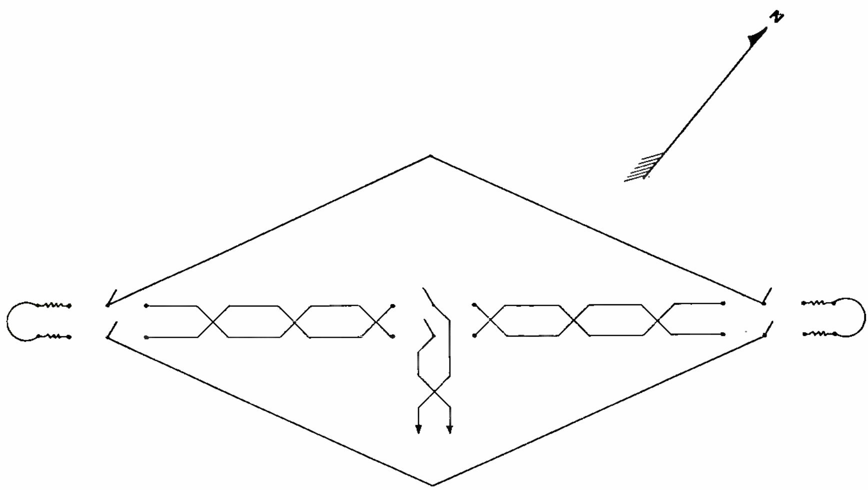


Fig. 2—41.5-Megacycle horizontal rhombic antenna fitted with remotely controlled switches to reverse directivity.

antenna and simultaneously connect a damping network to the other end. This made it possible to "listen" in either a northeasterly or southwesterly direction. It was found that the damping network reduced the back-end signal sensitivity about 28 db. Figure 3 shows the receiver and measuring equipment used.

The measurements were made by dividing the observations into five-minute periods and alternately measuring the 41.5 and 45-megacycle emissions. During each period maximum and minimum signal strengths were recorded.

Figure 4 shows the results of one day's measurements. The solid lines indicate the maximum and minimum values obtained on the London 45-megacycle channel. The broken lines indicate the maximum and minimum values obtained from the 41.5-megacycle channel. It shows the maximum signal received at the terminals of the receiver to have reached a peak value of about 700 microvolts on the 45-megacycle channel. It is also evident that there is a fairly constant ratio of fading of about 25 to 30 db on this channel. This phenomenon was observed on several occasions, but was not evident on the 41.5-megacycle signal.

Figure 5 gives a summary of all daily observations made on 41.5 and 45 megacycles between 1000 and 1100 E.S.T. The solid lines represent the daily ranges of the 41.5-megacycle field strength and the dotted lines represent the same for the 45-megacycle channel. Dates on which

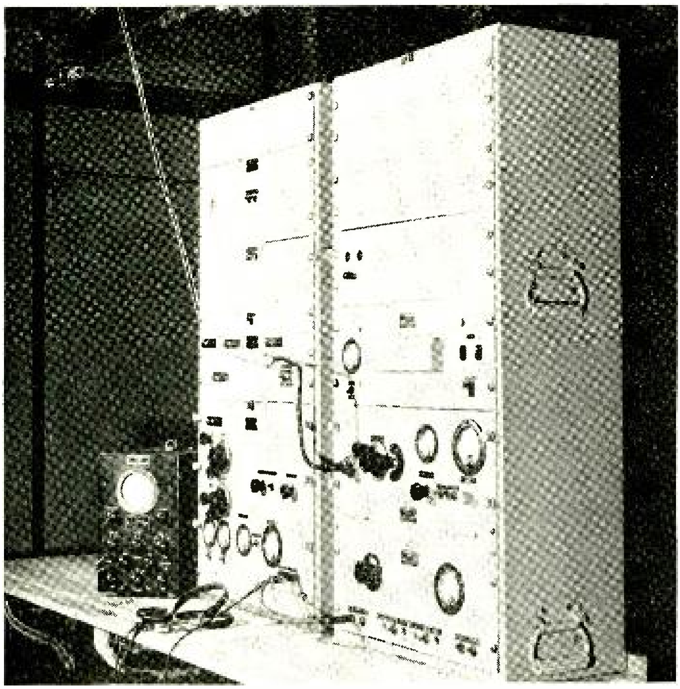


Fig. 3—Ultra-high-frequency receiver and signal generator used for field strength measurements.

no observations were made are indicated by "X". It will be noted that the signals were first heard January 21st. Conditions for this form of propagation seemed to be at their best during February, falling off badly in March. Whilst the data have not been plotted for the Berlin signals, these were also heard on a number of occasions in February.

Since a possible explanation for long distance propagation at these frequencies is that perhaps they are reflected by the F_2 layer, an examination of the F_2 critical frequencies for vertical incidence is of interest. Figure 6 shows a plot of monthly averages of the F_2 critical frequency for noon, Eastern Standard Time, as measured at Washington, D. C., by the National Bureau of Standards[2] over a period of years. It will be noted that the tendency has been toward higher values of F_2 critical frequency. It seems this tendency is in phase with the increase of sunspot numbers on the present eleven-year cycle of solar disturbances, which is due to reach a maximum about 1939.

Figure 7 shows F_2 critical frequencies as measured at Washington, D. C., by the Bureau of Standards,[3] each Wednesday between the hours of 1000 and 1600 E.S.T. plotted along with data relative to conditions observed on the 41.5-megacycle and 45-megacycle channels on the same

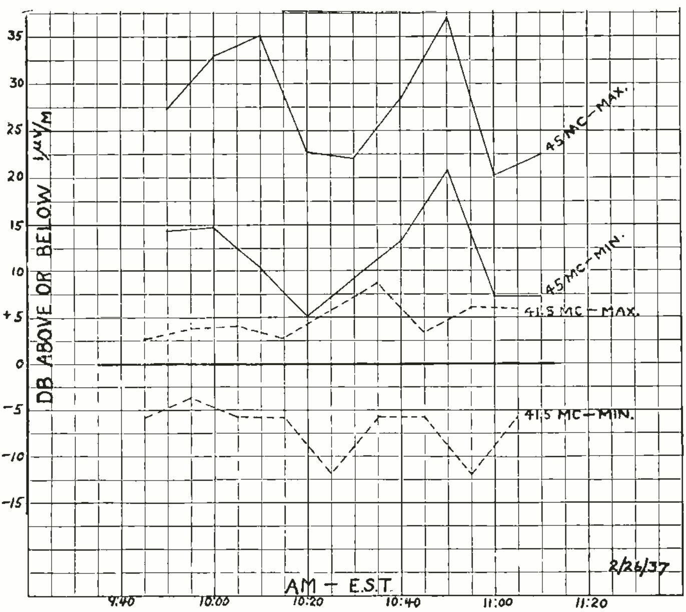


Fig. 4—Plot of single day's observation of the two transmitters at Alexandra Palace showing maximum and minimum values for each.

days. It is noted that the correlation is not perfect. Perhaps better correlation could be had if data were used for critical frequency measurements made more nearly on the path of propagation. Such data, however, were not available at the time.

[2] "Averages of Critical Frequencies and Virtual Heights of the Ionosphere, observed by the National Bureau of Standards, Washington, D. C., 1934-1936," by T. R. Gilliland, S. S. Kirby, N. Smith, and S. E. Reymer, in Terrestrial Magnetism and Atmospheric Electricity, Dec. 1936, Johns Hopkins Press, Baltimore, Maryland.

[3] The critical frequency data for Jan., Feb. and March of 1937 were kindly furnished by Dr. J. H. Dellinger of the Bureau of Standards.

The types of fading observed on the 41.5 and 45-megacycle channels differed greatly. Usually the 41.5-megacycle channel faded rapidly and deeply while the 45-megacycle channel was quite steady for several minutes at a time and would then slowly fade to a new signal level or pass through a shallow dip. The maximum field strength observed on the 45-megacycle channel was about 37 db above 1 microvolt per meter.

Normally the schedule of operation of the English transmitters was from 9:45 a.m. and again from 4 to 5 p.m. Eastern Standard Time. The 4 to 5 p.m. schedule so far has not definitely been heard at Riverhead.

For the week of February 8 the 41.5-megacycle transmitter was kept in operation until noon, Eastern Standard Time, but no definite improvement in field strength was observed during the additional hour. On March 31 the 41.5-megacycle transmitter was operated continuously from 6:30 a.m. until 1:00 p.m. Eastern Standard Time, but during this run the signal was unheard at Riverhead.

Observations were made simultaneously at LeRoy, Indiana from March 3 to March 31 inclusive. The 41.5-megacycle channel was heard on four occasions at LeRoy. On these four occasions, the signal was also heard at Riverhead, the field strength being somewhat higher at Riverhead. Apparently conditions favorable to transmission affect large areas at the same time.

The measurements made at the Frequency Measuring Laboratory consisted in observing both the 41.5 and 45-megacycle signals on various antennas. There were available several short wave fishbone antennas directed toward Europe, South America, the West Coast, etc. All of these were tried and it was frequently noted that when the signal was weak, best reception could be obtained by using an antenna directed toward the West Coast. On several occasions, the signal was inaudible on antennas directed toward Europe, but of reasonable strength on the West Coast antenna. However, during periods of strong signal the European antennas gave the best results. In general the reversible rhombic antenna gave similar results except that at no time did this antenna show an improved signal from the southwesterly direction. Usually during periods of weak signal the normal direction gave from 6 to 12 db better signal than the reverse direction. However, on two occasions for a period of several minutes each, the signals from both directions were of equal strength.

A possible explanation for the failure of the reversible rhombic antenna to show a good signal from the reverse direction is that the signal may have been coming to Riverhead over some path other than the great circle one. If this had been the case the rhombic antenna, being rather sharply directive, would show a good back-end response

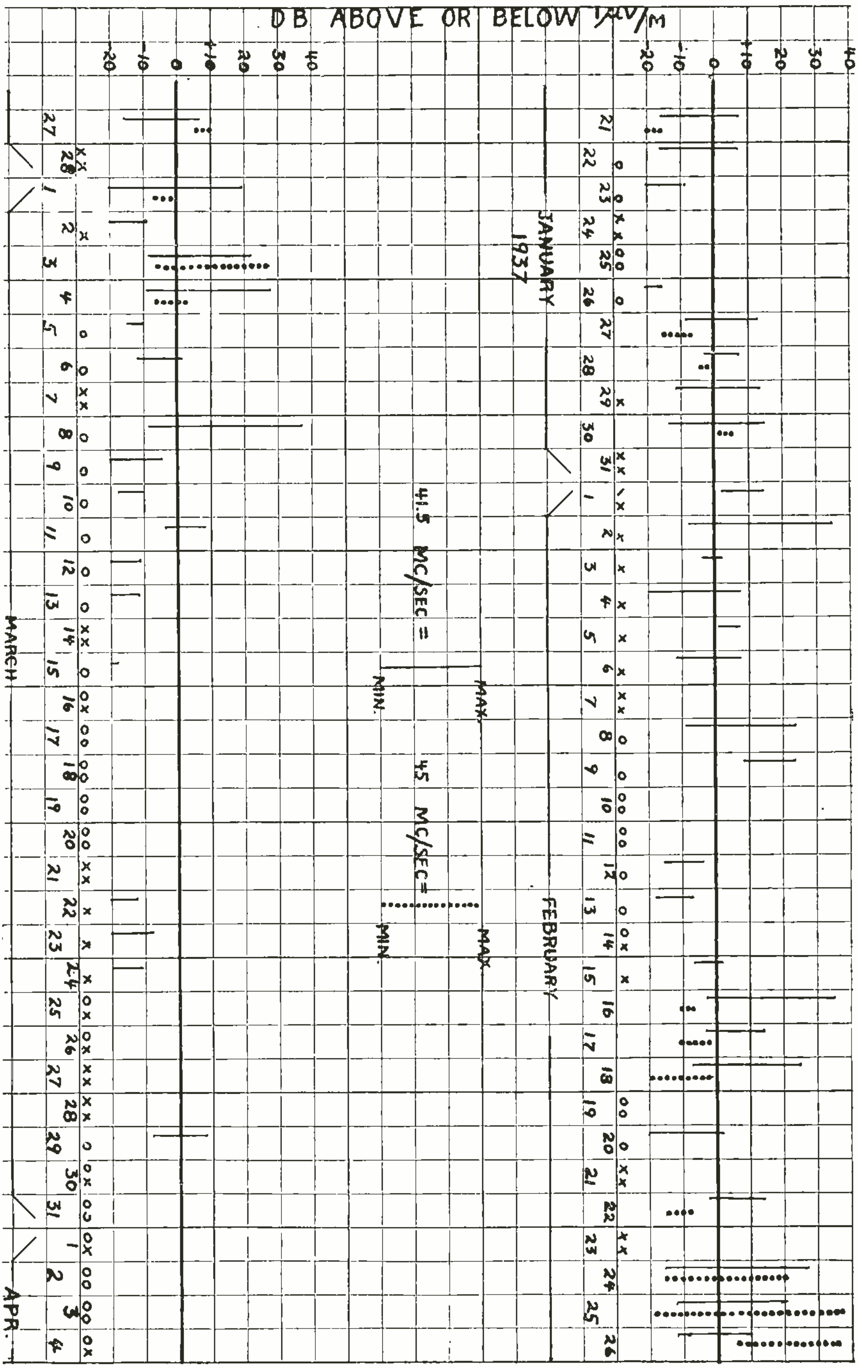


Fig. 5—Daily maximum and minimum field strengths. "X's" indicate no observation made. "O's" indicate signal unheard.

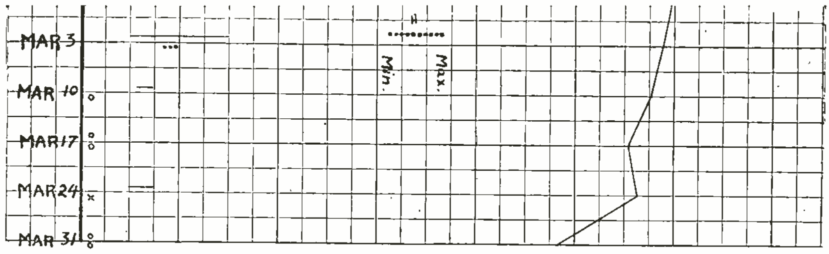
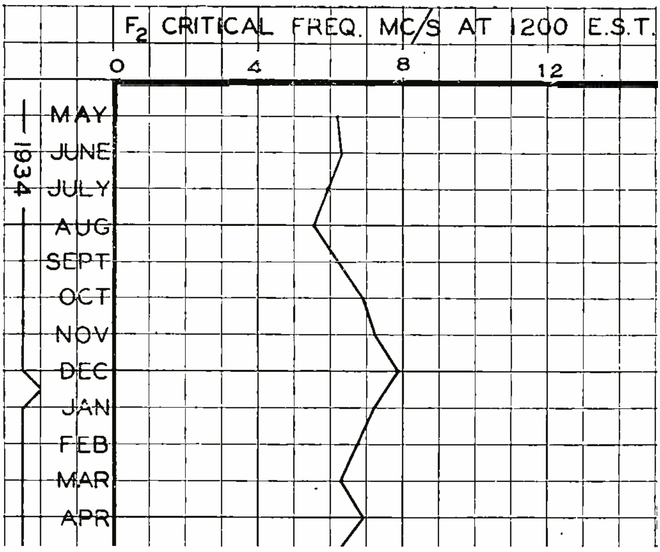


Fig. 7—Average F_2 critical frequency 10 A.M. to 4 P.M. E.S.T. each Wednesday at Washington, D. C., plotted with field strength ranges on same days. "X" indicates no observation made. "O" indicates signal unheard.

compare this new antenna with the rhombic indicated no instances of better results with the vertically polarized receiving antenna. The fact of the polarization of the received signal being independent of the polarization of the transmitting antenna supports the conclusion that propagation was by refraction phenomena in the ionosphere much the same as in the case of frequencies on the order of 10 to 20 megacycles.

Some additional vertical arrival angle measurements were made in the 29-megacycle amateur band. Figure 8 shows a table of the amateurs



observed on March 4, 1937. The vertical arrival angle together with the distance between Riverhead and the transmitter location allows a figure for the reflecting layer height to be computed if an assumption is made as to the number of reflections. In the calculations made to determine the column on the right a single reflection was assumed in all but two cases, in which two reflections were assumed. The average apparent layer height derived by this method on these assumptions was 346 kilometers. The average minimum F_2 layer height as measured by the Bureau of Standards at Washington, D. C., on March 3 during approximately the same time of day was 240 km. The difference may be due either to the method of measurement or to an error in the assumption made as to the number of reflections.

LAYER HEIGHT DETERMINATIONS AT RIVERHEAD, N. Y.
29 MC—AMATEURS

Call	*Location*	*Distance KM*	*Arrival Angle Degrees*	*Layer Height KM*
W5FHJ	Ruleville, Miss.	1800	17.2	360
....	Dallas, Tex.	2340	8.9	282
W9YUD	Fremont, Neb.	2040	14.1	378
W9DHO	Wisner, Neb.	2090	13.7	358
W5CZZ	Terrill, Tex.	2280	11.7	357
W5EYV	Refugio, Tex.	2600	9.0	357
W9JWI	Independence, Mo.	1960	17.2	402
W6LUL	Los Angeles, Cal.	4050	11.6	302
W9GND	Grand Falls, N. D.	2070	12.6	331
W9DHQ	Wishek, N. D.	2200	9.5	290
W5FNH	Kerrville, Tex.	2720	8.1	353
W5CZZ	Terrill, Tex.	2280	10.6	332
....	Kansas City, Mo.	1950	16.2	378
W7CKZ	Aberdeen, Wash.	4050	12.4	318
W9LKD	Wichita, Kans.	2180	10.9	332

MEAN LAYER HEIGHT — 346 KM

Fig. 8—Layer height and vertical arrival-angle determinations made with setup shown in Fig. 3.

ANALYSIS AND DESIGN OF VIDEO AMPLIFIERS

BY

S. W. SEELEY AND C. N. KIMBALL

License Laboratory, Radio Corporation of America

NATURE OF THE PROBLEM

THE amplification of the wide band of frequencies which constitute the video modulating signals in television transmission presents a special problem in amplifier design, since the requirements differ considerably from those encountered in audio amplifiers, in which only flat frequency response and freedom from harmonic generation are usually sought. Video amplifiers must be designed with particular reference to the maintenance of constant gain over the entire video frequency band, and attention must also be given to phase characteristics as affecting the time delay in transmission of the signals through the amplifier.

The high frequencies involved and the necessity for the maintenance of definite time-delay characteristics are the factors which require the most attention, and we propose to indicate means for attaining the desired amplifier characteristics through expedients which are easily applied in practice.

The present RMA standards of 441-line interlaced scanning, with a field frequency of 60 cps. and a frame frequency of 30 cps., impose severe requirements on the video amplifiers used in television receivers. The amplifiers must be capable of passing, with constant gain, all frequencies from 60 cycles to at least 2.5 megacycles, and the time delay must be substantially independent of frequency.

The necessity for constant time delay over the video band may be explained from consideration of the effect upon picture detail of using a video amplifier with phase characteristics which cause the high-frequency end of the video band to be delayed with respect to the low-frequency end in transmission through the amplifier. (This is generally the manner in which the time delay varies as a function of frequency in typical video amplifiers.) With 441-line scanning and a twelve-inch tube (ten-inch picture) the spot on the "Kinescope" screen moves at a rate of approximately 1.5 x 10^5 inches per second, that is, it takes about 7 microseconds to move one inch horizontally. (These figures are based on a return time in the horizontal sweep of ten percent of a scanning cycle.) Thus, consider the situation existing when the transmitted picture consists of a pattern half white and half black, with the vertical center line of the screen separating the two halves. The video signal is a square wave, containing a fundamental frequency of 13,230 cycles per second (441 lines and 30 frames), and all its odd harmonics. The maintenance of this wave form in transmission through the video amplifier requires that the time delay be constant for all frequencies. If the delay decreases with frequency, the higher harmonics of the square wave will be retarded less than the lower frequencies, and the resulting pattern on the "Kinescope" screen will not have the sharp line of demarcation between black and white as contained in the original picture. A difference in time delay of one microsecond between the high and low ends of the video band will cause a horizontal shift in the higher frequency components of the picture of about .14 inches with respect to the low-frequency components.

Similar results are obtained from an analysis of the situation on a phase shift basis, since the total time delay at any frequency is equal to the quotient of the total phase delay in the amplifier and the angular frequency. (Note that the phase reversal of 180 degrees which occurs in each stage of the amplifier due to tube action does not constitute a phase delay. We are concerned here only with the phase and time delays due to the presence of reactance in the plate circuit loads, and shall confine our remarks to these quantities.) The square wave generated by scanning the pattern described above may be expressed in a Fourier series of sines and cosines of the fundamental frequency (13,230 cycles) and its harmonics. The maintenance of the square-wave form requires that the total phase delay in the video amplifier vary linearly with frequency (as can be seen by analysis), and linearity of the phase characteristic implies a constant time delay.

Generally the patterns scanned by the "Iconoscope" beam are not as geometrically precise as that used here for discussion of the video amplifier requirements, but are made up of random variations of light and dark shading. The necessity for constant time delay is no less important in this case, for the picture will be distorted in the event of non-uniform time delay, especially if the pattern contains considerable detail.

It follows, then, that both constant time delay and flat frequency response are equally important in video amplifiers, and that anything done to bring about correction of one should not affect the other adversely. It is assumed here that the signal input will be held below the level which causes harmonic generation in the amplifier, so that harmonic distortion need not be discussed further.

ANALYSIS

In a well designed resistance-coupled amplifier, the top frequency which can be amplified without material loss in gain is determined by the effect of the reactance of the tube and circuit capacitances in shunting the resistive plate load.

Obviously the upper limiting frequency may be extended in any case by using a low value of plate-load resistor, so that the reactance of the shunting load-circuit capacitance is large in comparison with that load resistance. It is seen that the frequency range may be increased extensively if the load resistance is made sufficiently small, but the gain drops off at all frequencies as R_L is decreased.

One way to diminish the shunting effect of the load-circuit capacitances is to insert a properly proportioned choke in series with the output-load resistor. This causes the plate-circuit load of the stage to have very nearly constant impedance over a wide band of frequencies, the top frequency being determined by R, L and the total capacitance C from plate to ground.

The compensated stage, with its constant impedance load circuit, as shown in Figure 1, has a gain which is approximately constant and equal to $g_m R$ at all frequencies up to and including f_o, the top frequency which the stage must amplify. (See Appendix I for derivation.)

R and L for a given value of f_o are determined by the load-circuit capacitance C. To fulfill this condition R must be made equal to the reactance of this load capacitance at the top frequency, f_o, that is,

$R = \dfrac{1}{2\pi f_o C}$, and the reactance of the compensating choke at f_o must be equal to half the load resistance, i.e. $2\pi f_o L = \dfrac{R}{2}$

The gain of $g_m R$ per stage due to the use of this compensated load circuit is equal to the gain which would be experienced with zero load-circuit capacitance and no compensating choke; hence, the compensation for flat-frequency response is seen to be adequate.

With only R and C in the load circuit, and with no compensating choke, the gain is equal to

$$\frac{g_m R}{\sqrt{1 + C^2\omega^2 R^2}} = \frac{g_m R}{\sqrt{1 + f^2/f_o^2}} \text{ if } R = \frac{1}{2\pi f_o C}$$

Here the gain at the top frequency f_o is only .707 $g_m R$, a loss of approximately 30 per cent with respect to the gain of the compensated stage.

It is seen that, even in a compensated stage, the limiting frequency f_o can not be increased indefinitely, for the output-load resistance must be decreased as f_o is increased. Since the gain falls off inversely with f_o (gain $= g_m R = \frac{g_m}{2\pi f_o C}$), the limiting frequency is reached when $\frac{g_m}{2\pi f_o C} = 1$. At frequencies higher than $f_o = \frac{g_m}{2\pi C}$ the amplifying properties of the stage disappear, and the output voltage becomes less than the input voltage.

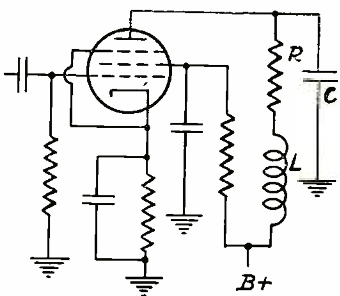


Fig. 1

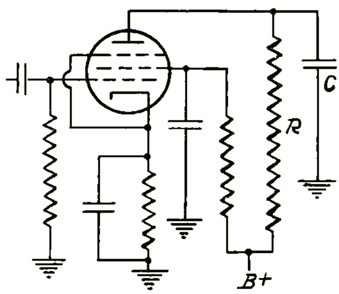


Fig. 2

There is a simple method for determining the load-circuit capacitance of each stage in the video amplifier, which depends upon the fact that the gain of an uncompensated stage falls to 70.7 per cent of its low-frequency value at the frequency f' for which $R_L' = \frac{1}{2\pi f' C}$. (Here R_L' is the output-load resistance and C is the load-circuit capacitance to be measured.) The procedure is as follows: In the plate circuit of the first stage of the amplifier insert a load resistor of about 3000-5000 ohms. Place a fixed bias on the grid of the second tube just sufficient to produce cathode current cut-off. Apply a low-frequency signal (about 10 kc) to the grid of the first tube and adjust its magnitude to produce a second-tube cathode current of some predetermined value, say .1 ma. Now determine the frequency f' at which the input voltage to the first tube must be increased to $\sqrt{2}$ times its low-frequency value to maintain the cathode current of the second tube at .1 ma (i.e., f' is the frequency at which the stage gain is down 30 per cent).

This frequency f' is used to calculate C by $C = \frac{1}{2\pi f' R'_L}$. (Note that this value of C includes all the tube and circuit capacitances effective during operation of the amplifier.) This frequency f' will generally be lower than the top frequency which the compensated stage is intended to amplify. With this value of C next determine R_L (to be used in the compensated circuit) to satisfy the equation $R_L = \dfrac{1}{2\pi f_o C}$ where f_o is the top frequency to be passed by the amplifier. The compensating choke to be inserted in series with R_L should have a reactance at this top frequency f_o of half the value of the load resistance, that is $2\pi f_o L = \dfrac{R_L}{2}$.

This procedure can be repeated stage by stage throughout the entire amplifier, by connecting, in each case, the signal generator to the grid of the stage whose load-circuit capacitance is desired, and by using the following tube as a vacuum-tube voltmeter.

The determination of C for the last stage may be made in this manner by utilizing the "Kinescope" as a vacuum-tube voltmeter. Bias its control grid back to a point which permits the cathode-ray tube to act as a plate-circuit detector, and repeat the procedure previously outlined. The use of the "Kinescope" in this manner permits the measurement of C for the last tube under actual operating conditions, and C therefore includes the input capacitance of the Kinescope control grid.

Phase and Time Delay

I. *Uncompensated video amplifier stage:* The gain is equal to $g_m R \Big/ \sqrt{1 + \dfrac{f^2}{f_o^2}}$ if $R = \dfrac{1}{2\pi f_o C}$. The phase shift due to passage through the stage of a signal of frequency f is $\phi = -\tan^{-1} 2\pi f C R = -\tan^{-1} \dfrac{f}{f_o}$. (See Appendix II for derivation), where the negative sign means a greater phase *delay* for the higher frequencies than for the lower ones.

Note that this delay is due only to the presence of reactance in the plate-circuit load. There is no delay in the tube at these frequencies, for the tube merely reverses the phase of its input voltage.

The actual time delay in seconds corresponding to a frequency f is $\Delta t = \dfrac{\text{phase delay in radians}}{2\pi \times \text{frequency in cycles/second}}$

The phase shift and time delay for several stages in a video amplifier are additive; three similar stages cause three times the time delay of a single stage, whereas the gain of a three-stage amplifier is the product of the individual stage gains.

Figure 3 shows curves of phase delay and gain vs. $\frac{f}{f_o}$ for a single uncompensated stage. Note that the phase delay does not increase linearly with frequency, hence the time delay is not constant over the frequency band.

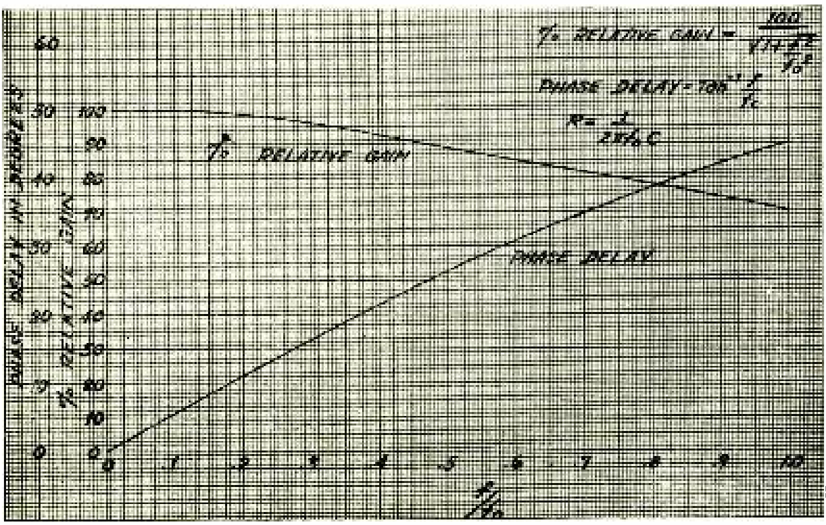


Fig. 3—Uncompensated stage of video amplifier.

The quantitative effect of non-uniform time delay will be discussed in detail in the section dealing with compensated video amplifiers.

II. Choke compensated amplifier stage: As noted previously, the condition for flat-frequency response out to a frequency f_o is

$$R = \frac{1}{2\pi f_o C} = 4\pi f_o L$$

The phase delay for a single compensated stage is

$$\phi = +\tan^{-1} 1/4 \left[\left(\frac{f}{f_o} \right)^3 + 2 \left(\frac{f}{f_o} \right) \right]$$

and the time delay in seconds is

$$\triangle t = + \frac{1}{2\pi f} \tan^{-1} 1/4 \left[\left(\frac{f}{f_o} \right)^3 + 2 \left(\frac{f}{f_o} \right) \right]$$

(see Appendix I for derivation of these expressions). Note that the phase and time delays for a given frequency f are dependent only upon f and the top frequency f_o, regardless of the numerical values of R, L and C which are used to attain flat-frequency response out to a frequency f_o. Here, again, the total phase and time delay for several stages is equal to the algebraic sum of individual stage delays.

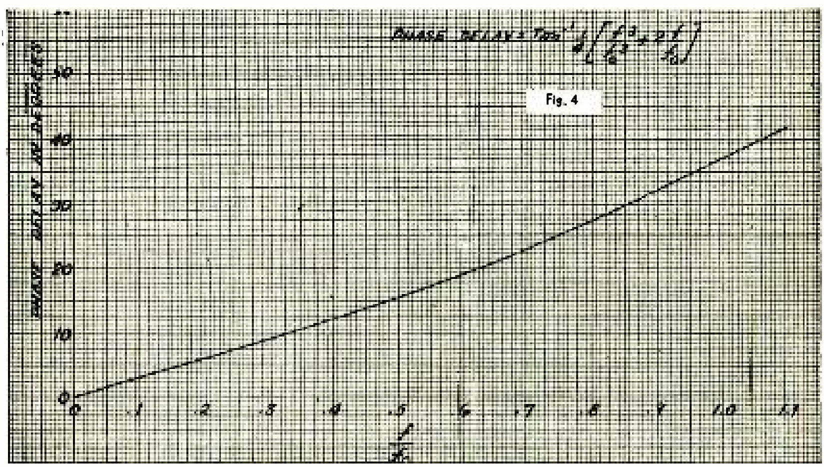


Fig. 4—One compensated stage of video amplifier.

The phase delay of a compensated stage is plotted vs. f/f_o in Figure 4 and it is seen that the non-linear phase characteristic will result in a non-uniform time-delay curve.

The elements in the load circuit of a video-amplifier stage can be proportioned to produce a constant time delay throughout the video band but this generally results in a non-uniform gain characteristic.

As a quantitative indication of the magnitude of anticipated time delay and its effect upon the displacement of picture elements on the "Kinescope" screen, there is plotted in Figure 5 the total time delay due to a three-stage video amplifier compensated for constant gain to a frequency of 2.5 megacycles, and a curve of the actual horizontal displacement of picture elements corresponding to the different frequencies in the video band is shown. Since the video-detector load will generally be compensated for constant impedance to f_o, its contribution to the total time delay has been included. Therefore, the net time delay is equivalent to that of a four-stage video amplifier fed from an uncompensated detector load. The calculations of element displacement are based on 441-line horizontal scanning, a ten-inch picture on a twelve-inch tube, and ten percent return time in the horizontal sweep.

Figure 5 shows that the delay increases with frequency. The total time delay is not significant, as it is the difference in time delays for the various frequencies in the video band with which we are concerned. These differences are the cause of the relative displacements of the various frequency components in the picture. Constant time delay

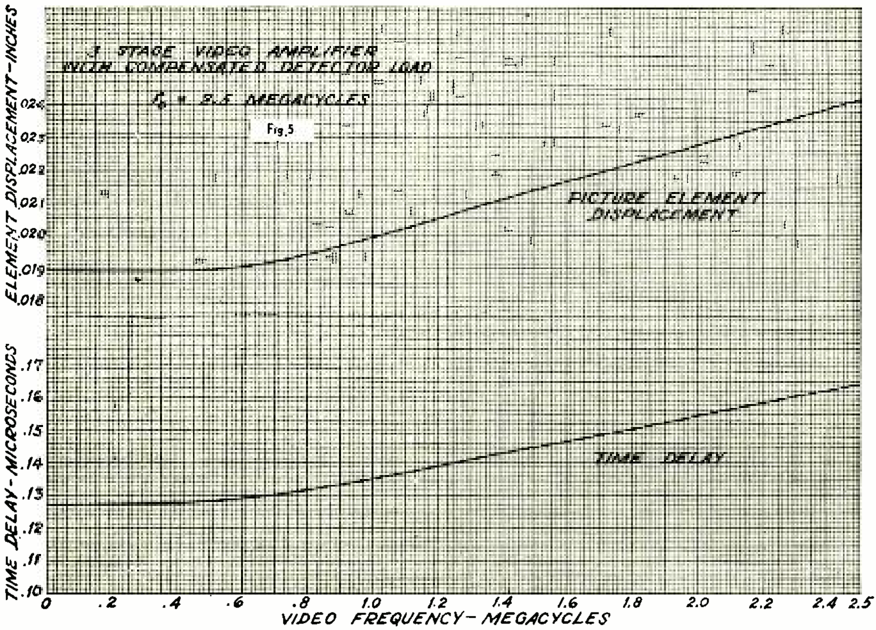


Fig. 5—Three-stage video amplifier with compensated detector load.

would result in all the picture elements being displaced by the same amount, regardless of the video frequency with which they are associated, and no picture distortion would result. As it is, the curve of displacement vs. frequency shows that picture elements corresponding to the two extremes of the video band (60 cycles and 2.5 megacycles) will be displaced by approximately .019″ at the low end and .024″ at the top end. The relative displacement, or the effect which causes the relative shift in picture elements and the corresponding distortion, is only .005″, and this is small in relation to the width of a scanning line.

Number of Stages

The choice of an even or odd number of stages in the video amplifier depends upon the video detector circuit and the method of transmission. With negative modulation, as specified in the current RMA standards, a positive pulse of modulation on the television carrier occurs when the scanning beam in the "Iconoscope" passes through a black portion of the picture. With any type of detector circuit in which the cathode end of its load resistor becomes positive for the video signal during modulation peaks, it requires an even number of tubes in the amplifier to reproduce on the "Kinescope" screen the same polarity of shading as that in the transmitted picture. This is true when the detector cathode is grounded. If the negative end of the detector load is grounded an odd number of amplifier tubes is required; this connection may not be so favorable to uniform video-frequency response in an uncompensated detector-load circuit because of the shunting effect on the detector load of the heater-cathode capacitance of the detector tube.

Low Frequency Considerations

The maintenance of the proper gain and phase-delay characteristics at the low-frequency end of the video band (60 cycles) requires that attention be given to the coupling circuits between successive stages, since the cause of non-uniform characteristics in this part of the band will generally be due to insufficient interstage coupling.

An extremely small departure from linearity in the phase delay vs. frequency characteristic at very low frequencies can be quite serious. One degree at sixty cycles corresponds to 46.2 microseconds, and a .1 $\mu fd.$ coupling condenser in conjunction with one-megohm grid leak produces a phase shift of 1.5 degrees or 69.3 microseconds. Consideration of the reproduced pattern under these conditions shows that if a solid white screen were being transmitted a 1.5 per cent change in intensity from top to bottom for each such coupling unit would result.

For this reason the low-frequency characteristic of each stage must be compensated by the use of a plate-circuit load impedance which becomes capacitive at low frequencies. This is accomplished by including a second load-circuit resistance at the low (v.f.) potential end of the main resistor. This additional resistor is by-passed by a condenser such that the phase delay in the total load circuit (at low frequencies) just compensates for the phase advance caused by the preceding grid-coupling circuit.

The tendency toward motor-boating in video amplifiers is sometimes prominent because of the maintenance of normal gain at low frequencies. This is best avoided by using as small a coupling condenser as possible, consistent with proper 60-cycle performance, and by maintaining the output impedance of the power supply at a very low level for frequencies at which motor-boating is liable to occur (10-30 cycles). Separation of the screen supplies for the different amplifier tubes by means of high inductance (500-henry) chokes and heavy by-passing of all screen leads with 8 $\mu fd.$ electrolytic condensers generally suppresses all tendency toward motor-boating.

Measurement of Gain and Phase Delay

The gain characteristic under actual operating conditions is best determined by utilizing the "Kinescope" as a vacuum-tube voltmeter, since the input capacitance of its control grid is then present across the output circuit of the last stage in the video amplifier. The tube should be biased back to act as a plate-circuit detector, and the gain characteristic is determined from measurements of the input voltage to the amplifier required to maintain the cathode current of the "Kinescope" at some constant value.

The phase-delay characteristic of a compensated video amplifier can be determined directly from calculation or from the curve of Figure 4 if it is known that the gain is constant for all frequencies up to the frequency which represents the top of the desired video band. If, however, the gain is not constant, due possibly to intentional over-compensation to produce an increase in high frequency gain, the phase-delay characteristic is not as easily calculated, and may best be determined experimentally.

This measurement is most effectively made with the aid of an oscilloscope whose horizontal and vertical amplifiers are identical and capable of amplifying at least up to 2.5 megacycles. The application of this instrument to the measurement of phase delay makes use of the fact that two voltages of the same frequency, applied to the separate pairs of deflecting plates in the cathode-ray tube (through amplifiers, if the voltage level is so low as to preclude direct application of the voltages to the plates), cause the trace on the oscilloscope screen to assume a definite pattern, dependent upon the relative amplitudes and the phase relation of the voltages under observation. The oscilloscope trace is linear when the voltages are 0 degrees or 180 degrees out of phase, and becomes a circle with 90 degrees phase angle and equality of amplitude of the voltages.

Any other phase relation causes the trace to be elliptical, and the desired phase angle can be determined graphically from measurements on the screen of the major and minor diameters of the ellipse, or, more accurately, by employing *R-C* circuits to shift the phase of one of the voltages until a linear trace is made to appear. The unknown angle is then computed from ω and R and C of the phase-shifting network.

The presence of capacitive reactance in the plate circuits of the video amplifier causes the output voltage to lag the input voltage in time phase; hence the *R-C* circuits used for phase shifting in the phase-angle measurements must be so arranged as to cause the phase of the input voltage to be delayed before it is applied to the oscilloscope, or, conversely to advance the phase of the output voltage. Figures 6 and 7 show typical circuits for use in this work. A linear trace is obtained when the phase shift in the *R-C* network is equal to the phase shift in the video amplifier. As noted below $\phi = (90° - \theta)$, where θ is the angle between e_{gen} and i through R and C. Note that C must include the additional capacitance occasioned by connection to the oscilloscope; i.e., either the input capacitance of the horizontal amplifier or the

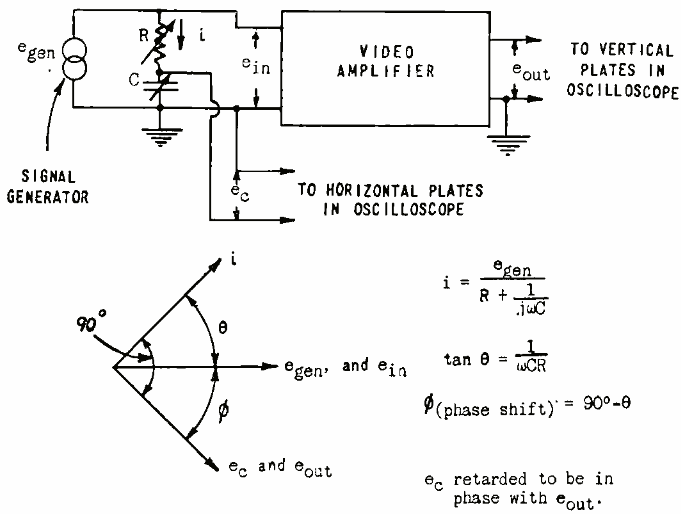


Fig. 6

capacitance between horizontal deflecting plates, depending upon whether or not the voltage is applied directly to the plates. The input capacitance of the vertical-deflection system will also add to the output capacitance of the video amplifier, and this must be taken into account in determining overall performance.

The necessity for the horizontal and vertical amplifiers to be identical applies only to their phase characteristics, which may have any arbitrary shape so long as they are the same. Similar gain characteristics are not necessary.

An alternative arrangement for phase-angle measurement is shown in Figure 7.

The measurement is facilitated considerably, and no doubt is left as to the relative phase characteristics of the two oscilloscope amplifiers, if an additional amplifier (whose phase and gain characteristics are arbitrary) is interposed between the signal source and the input to the video amplifier. A capacitance attenuator may be used to prevent overloading due to excessive input to the video amplifier, and the voltage derived from the phase-shifting network may be used for direct application to one pair of plates, while the video amplifier's output voltage is applied directly to the other pair.

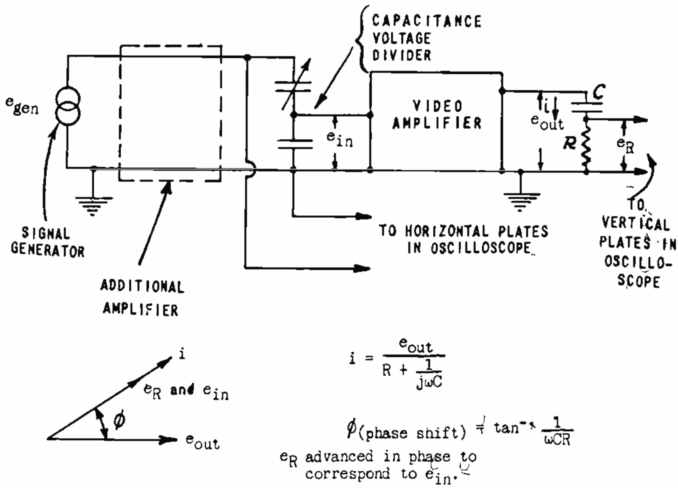


Fig. 7

APPENDIX I.

Gain, phase and time delay of a compensated stage in a video amplifier.

Let $r_p >> Z_L$, hence the gain $= g_m Z_L$, and the phase shift is equal in degrees to the phase angle of the complex impedance $Z_L = R_L \pm jX_L$.

$$Z_L = \frac{(R + j\omega L)\,\dfrac{1}{j\omega C}}{R + j\left(L\omega - \dfrac{1}{\omega C}\right)} = \frac{R + j\,(L\omega - L^2 C\omega^3 - R^2 C\omega)}{R^2C^2\omega^2 + (LC\omega^2 - 1)^2}$$

Substituting $R = 2L\omega_o = \dfrac{1}{C\omega_o}$ as the condition for constant gain to $\dfrac{\omega_o}{2\pi}$ cycles.

$$Z_L = \frac{R\left[1 - \dfrac{j}{4}\left(\dfrac{f^3}{f_o^{\,3}} + 2\dfrac{f}{f_o}\right)\right]}{\left(\dfrac{f}{f_o}\right)^2 + \left(\dfrac{f^2}{2f_o^{\,2}} - 1\right)^2}$$

$$\text{Gain} = \frac{g_m R \sqrt{1 + \left(\frac{1}{4} \frac{f^3}{f_o^3} + \frac{1}{2} \frac{f}{f_o} \right)^2}}{\left(\frac{f}{f_o} \right)^2 + \left(\frac{f^2}{2f_o^2} - 1 \right)^2}$$

The phase delay in the stage is

$$\phi = + \tan^{-1} \tfrac{1}{4} \left(\frac{f^3}{f_o^3} + 2 \frac{f}{f_o} \right)$$

and the time delay

$$\triangle t = \frac{\phi}{2\pi f} = + \frac{1}{2\pi f} \tan^{-1} \tfrac{1}{4} \left(\frac{f^3}{f_o^3} + 2 \frac{f}{f_o} \right)$$

APPENDIX II.

Gain, phase and time delay in an uncompensated stage of a video amplifier.

Let r_p be very large in comparison to Z_L, so that the gain may be written as $g_m Z_L$.

$$Z_L = \frac{\frac{R}{j\omega C}}{R + \frac{1}{j\omega C}} = \frac{R(1 - jRC\omega)}{R^2C^2\omega^2 + 1} = \frac{R}{\sqrt{R^2C^2\omega^2 + 1}}$$

Let $\frac{\omega_o}{2\pi}$ be the frequency at which $R = \frac{1}{2\pi f_o C}$, then

$$Z_L = \frac{R}{\sqrt{1 + f^2/f_o^2}} \text{ and the gain} = \frac{g_m R}{\sqrt{1 + \frac{f^2}{f_o^2}}}.$$

With $r_p >> Z_L$ constant current flows through Z_L, and the phase shift in voltage due to the presence of C is equal to the phase angle of the complex impedance $Z_L = R_L - jX_L$.

Phase delay $\phi = \tan^{-1} \left| \frac{X_L}{R_L} \right| = + \tan^{-1} RC\omega$ or, substituting

$$R = \frac{1}{2\pi f_o C}.$$

The time delay at any frequency f is $\triangle t = \frac{\phi}{2\pi f}$, which here equals

$$+ \frac{1}{2\pi f} \tan^{-1} \frac{f}{f_o}.$$

THE DESIGN OF INDUCTANCES FOR FREQUENCIES BETWEEN 4 AND 25 MEGACYCLES

By

DALE POLLACK

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***Summary**—The results of a study of the resistance of small single layer inductances for use at frequencies between 4 and 25 megacycles are given. From experimental and analytical work a procedure has been deduced for the optimum design of such inductances.*

INTRODUCTION

WHILE it has long been recognized that radio receiver performance is dependent upon the merit of the inductances employed in the tuned circuits, few data beyond 4 megacycles are available upon which to base the design of such inductances. Because of the lack of this information, design calculations at high frequencies have in general been limited to obtaining the correct inductance values. While some attention has been paid to obtaining low effective resistance, the efforts in this direction have usually been misdirected, with the emphasis on factors which actually are of little consequence in determining coil resistance.

A tuned circuit is represented to a first degree of approximation by the equivalent circuit shown in Figure 1. For the small coils considered in this study, C_o is very small[1], less than $5\mu\mu f.$, and it may be combined with the capacitance of the tuning condenser, C, with which the coil is associated, since R_c is always small. The effective resistance of the inductance is usually considerably larger than the effective resistance of the condenser, but, at high frequencies, the condenser resistance may also be of importance.

The numerical value of the inductance, L, is determined by circuit considerations, and, consequently, is fixed by specification. This paper is concerned with the value of the effective resistance, R, of the coil, and with the design of coils to have a minimum effective resistance. It is usual to compare tuned circuits on the basis of the ratio of their reactance to their total effective resistance, $\frac{\omega L}{R}$ or $\frac{1}{\omega CR}$, denoted by Q. The highest Q consistent with practical limitations—such as those

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of cost and space—is the desideratum. The importance of high Q circuits in modern receiver design is well known.

The increase in resistance of an air core solenoid at high frequencies over its d-c resistance is the result of[2]:

1. Skin effect in the wires, caused by the non-uniform current distribution over the cross-section of the wire, caused by the internal flux of the wire itself.
2. Proximity effect, that is, the non-uniform current distribution over the cross-section of the wire caused by the flux from neighboring turns of wire.
3. Dielectric losses, caused by hysteresis in the dielectric between turns of wire, or located near the wire.
4. Eddy current losses in metallic bodies in the field of the coil.
5. Electromagnetic radiation losses.
6. Self-capacitance loss.

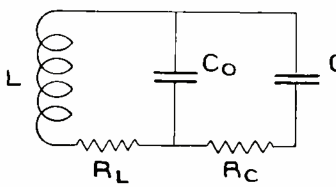

Fig. 1. Approximate equivalent circuit of coil and tuning condenser.

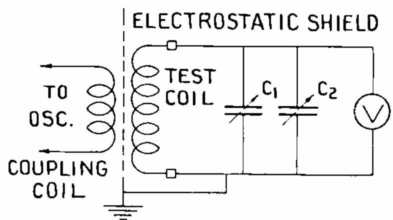

Fig. 2. Measurement circuit. A vacuum tube voltmeter is used.

As a consequence of these losses, the effective resistance of a coil may be many times its direct current resistance. The magnitude of these losses is determined by the construction of the coil, the factors which must be considered in designing a coil being:

1. Coil dimensions: length and diameter.
2. Wire: material, insulation and size.
3. Coil form: material, insulation and size.
4. Location of coil with respect to metallic and dielectric bodies.

The design problem is to proportion these factors, so that the resistance of the coil will be a minimum.

Early Work

The most extended analytic work on coil resistance at high frequencies which can be applied to the types of coils used in radio equipment is that of Butterworth[3]. Butterworth's solution treated both skin effect and proximity effect in short solenoids and multilayer coils.

The mathematical analysis made in this paper, in Appendix II, is an extension and simplification of Butterworth's conclusions, as applied to short solenoids at frequencies above 4 megacycles. Butterworth's treatment is useful, in that the assumptions which were found necessary did not place any restriction on the frequency. However, his solution to this difficult problem was complicated, and has not received much practical attention. The conclusions were later reduced to a form better adapted to practical computations[4], but the design method was still involved and has not been generally applied.

Published experimental information on coil resistance above 4 megacycles is meagre. Hund and DeGroot[5] carried their measurements as far as 5 megacycles. Hall[6] describes tests at frequencies up to 6 megacycles, and Morecroft[7] presents data on measurements made up to 4 megacycles. A more recent paper, by Schwarz[8], describes tests made at frequencies up to 10 megacycles. None of these data have been correlated for design purposes. References to work above 10 megacycles is almost entirely lacking, only one published paper having been found, by Barden and Grimes[9], describing measurements at the single frequency, 15 megacycles.

Experimental

Tests were performed on two sets of coils, having inductances of approximately 1.1 and 3.6 microhenrys. A large number of coils were wound, with groups in which one design parameter at a time was varied. The extremes over which the measurements were carried were:

Frequency: 6 to 22 megacycles.
Wire size: 14 to 36, B&S gauge.
Coil diameter: 2.5 to 6.5 centimeters.
Coil length: 0.5 to 3.5 centimeters.

In addition, tests were made for a variety of coil form materials. and for several depths of winding grooves for the wire. Except where otherwise noted enameled copper wire was used.

The circuit resistance was measured by the reactance variation method[10], using two condensers in parallel, as in Figure 2, and a vacuum tube voltmeter as the indicator. The condenser, C_1, is the tuning condenser, while C_2, with a total capacitance variation of only 6 $\mu\mu f$. was employed to obtain the reactance change. The condensers were calibrated by a substitution method at a low radio frequency. The voltmeter was calibrated by comparison with a thermocouple, also at a low radio frequency.

Each of the coils was mounted on a pair of brass pins about 4 cm. long, which fitted into mercury contact cups mounted in the center of a cubical copper compartment about 20 cm. on each edge. The frequency of the oscillator used for a power source was maintained constant to within about 0.02 per cent by tuning to zero beat with harmonics of a crystal oscillator and maintaining a continuous audible check on the beat note with a radio receiver. In this way any frequency drift or reaction of the tuned circuit upon the oscillator source would have been detected, if it had occurred.

In the application of the reactance variation method of measurement, the total of all the circuit losses is obtained, including condenser and lead losses, tube input losses and the coil losses. In this study, the data, except where otherwise noted, include the circuit and condenser resistance, but a correction has been made for the input impedance of the vacuum tube voltmeter, as mentioned below. Despite

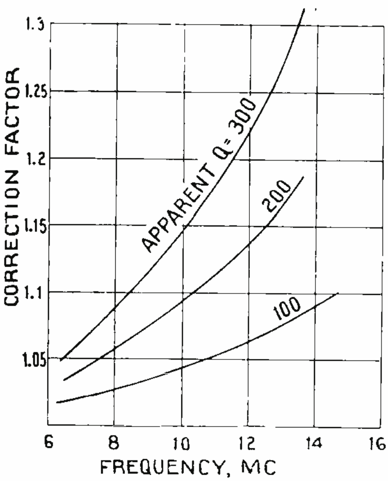

Fig. 3. Correction factor to correct for the input resistance of the '24A tube in the vacuum tube voltmeter.

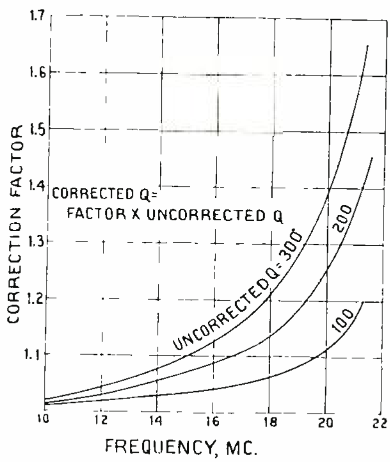

Fig. 4. Correction factor to correct for the input resistance of the triode connected 954 tube in the vacuum tube voltmeter.

the fact that some of the corrections have not been made, at any frequency, for a given coefficient of self-inductance, the various experimental curves are comparable, because under such conditions the circuit and condenser losses are constant.

As a matter of interest, an attempt was made to determine the magnitude of the loss resulting from the input resistance of the vacuum tube voltmeters. In the tests of the 3.6 microhenry coils, a '24A tube was used in the voltmeter circuit, while, in the tests of the 1.1 microhenry coils a 954 acorn tube was employed, in a triode connection. The tube losses were determined by shunting the circuit of Figure 2 with a second identical vacuum tube voltmeter, and remeasuring the Q. From the change in the overall Q of the circuit, the effective resistance of the tube can be calculated. This method assumes that the condenser losses are independent of the small capacitance change necessary to retune the circuit to resonance, when the second vacuum tube is added[12]. From the data obtained for the effective input resistance of the voltmeters, correction factors for any Q and any frequency have been plotted. These are shown in Figures 3 and 4. While, as has been pointed out previously, it is not essential that these corrections be made, since the information was available, it was employed, and the experimental results presented in this paper have been corrected for the input resistance of the voltmeter.

RESULTS

In Appendix II expressions are derived for the effective resistance, or the Q, of a coil at high frequency, taking account of the wire losses, i.e., the skin effect and the proximity effect only. When the frequency is well below the natural frequency of the coil, it is found, after certain assumptions are made, that the Q of an inductance coil is

$$Q = \frac{f^{1/2}\, LdS^2D}{10^{3/2}\, \rho^{1/2}\, (S^2D^2N + 2N^3d^2)} \tag{20}$$

or, the Q is directly proportional to the square root of the frequency. For the coils employed in these tests (20) has been found to agree with Butterworth's equations, from which it is derived, by better than 10 per cent.

An exact experimental check on this equation is difficult because of the impracticability of separating precisely the skin and proximity effects from the circuit and coil losses which the equation does not include. Moreover (20) was not used for design purposes and it consequently was not deemed essential to make a precise check. For illustration, however, a typical set of data are shown in Figure 5, in which an attempt is made to break down the various loss components of the measurement circuit. Dielectric losses in the coil form and wire insulation and vacuum tube losses were corrected for by measurement, as described elsewhere in this paper. The distributed capacitance correction was made by (12). This correction is necessary, because the tuning condenser in the reactance variation method is calibrated without a coil in the circuit. Effectively, therefore, the addition of the distributed capacitance of the test coil constitutes an error in the calibration of the tuning condenser, and a correction must be made to the measured value. The approximate lead losses were calculated from the d-c lead resistance by applying the usual skin effect formulas.

The condenser losses were estimated by assuming a condenser resistance of 0.02 ohm at 10 mc., increasing to 0.045 ohm at 20 mc. This is a conservative estimate; the condenser resistance was probably in excess of these figures, but could not be measured conveniently.

The only loss components properly chargeable to the coil, besides the skin and proximity effects, are the dielectric losses and the unaccounted losses.

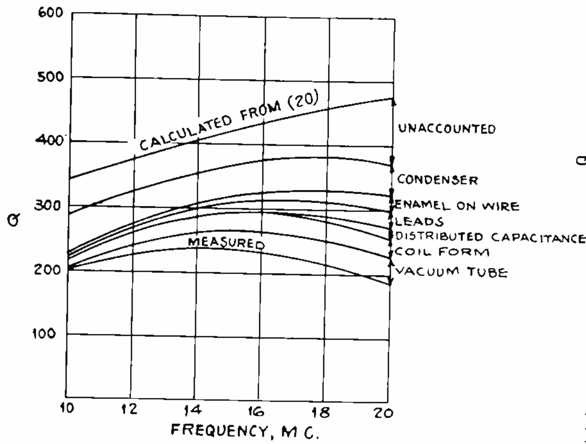

Fig. 5. Comparison of calculated and measured Q. Coil dimensions for this figure: $L = 1.1$ microhenrys, $D = 2.5$ cm., $b = 3.4$ cm., wire size = No. 18 B&S.

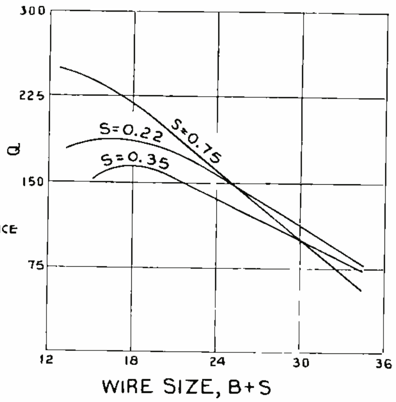

Fig. 6. Experimental curves illustrating variation in Q with wire size. For this set of curves, $f = 20$ mc., $L = 1.1$ microhenrys, $D = 2.5$ cm. Curves similar to these were used in preparing the data of Table I.

Effect of Wire Size

An experimental set of curves illustrating the variation in Q with wire size is shown in Figure 6. The optimum wire diameter, as obtained in Appendix II, is

$$d_o = b \sqrt{\frac{D}{2L\,(102S + 45)}} \tag{16}$$

or

$$d_o = \frac{b}{\sqrt{2}\,N} \tag{17}$$

In Table I[18] the values of the optimum wire size calculated from either of these expressions is compared with experimental values obtained from curves similar to those of Figure 6. The test coils were wound on bakelite forms with enamelled wire, typical of the construction used in practice. In most cases the agreement is better than the difference between diameters of successive wire sizes, namely, 12 per cent. The probable experimental error is about 8 per cent. It may be concluded that there is an optimum wire size for any given coil, and that this optimum size can be calculated from (16) or (17).

Coil Dimensions

The determination of the proper ratio of length to diameter for a solenoid is more difficult. Some confusion has resulted from the wide variations in conclusions which have been reached by various investigators[3,9,14]. The apparent inconsistencies result from the fact that different variables are considered in different studies. The solution which is reached must always depend upon the geometrical or eco-

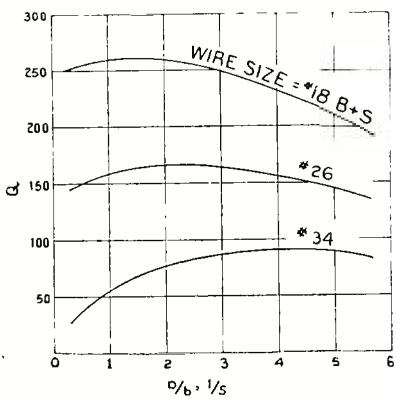

Fig. 7. Experimental curves illustrating the variation in Q with coil diameter:length ratio, when the diameter is constant. For this set of curves, $f = 17$ mc., $D = 2.5$ cm., $L = 1.1$ microhenrys.

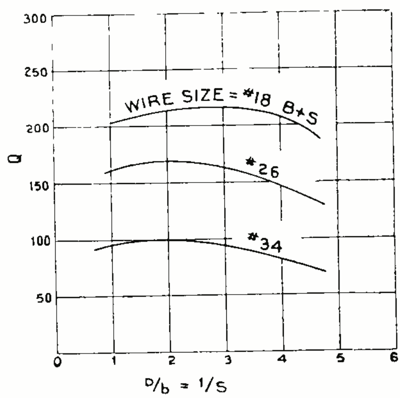

Fig. 8. Experimental curves illustrating the variation in Q with coil diameter:length ratio, when the length is constant. For this set of curves, $f = 13$ mc., $b = 1.5$ cm., $L = 3.6$ microhenrys.

nomical limitations which are assumed. The coil volume, surface area, wire length or any of several other factors may be assumed constant, depending upon the nature of the problem. It should also be noted that either the wire size may be kept constant, or the optimum value may be employed, although, for design purposes, the latter is preferable.

The effect of changes in ratio of coil length to diameter, for a certain group of coils in which the coil diameter, wire size, inductance and frequency are held constant, is shown in Figure 7. A similar set of curves, except that the coil length is held constant, instead of the diameter, is given in Figure 8. For the specifications for which these curves are obtained, namely, the wire size, inductance, frequency and either the diameter or length constant, it appears that the optimum ratio of length to diameter is between 0.5 and 0.3. Fortescue[14] arrived at optimum values of the length:diameter ratio greater than unity, from an analytical treatment based upon similar assumptions, but Butterworth has shown that an error must have been contained in the calculations. The curves also illustrate that the maxima obtained are flat, and values within a reasonably wide range may be employed without departing greatly from the maximum value of Q.

For design purposes the variation in Q with the coil dimensions, when the optimum wire size of (16) or (17) is used, is of greater importance, since higher Q's may be obtained in this way. In Appendix II it is shown that, if the wire size is maintained at the optimum value, the figure of merit is given by,

$$Q_m = \frac{f^{1/2} Db}{\sqrt{2000}\ \rho^{1/2}\ (102b + 45D)} \qquad (22)$$

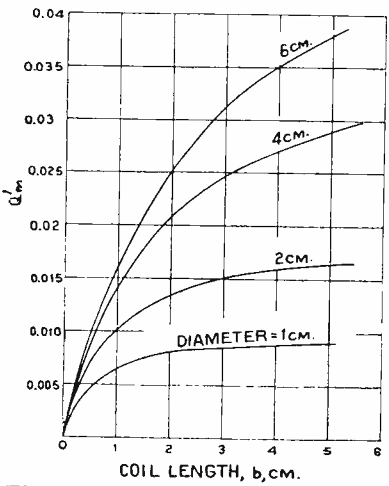

Fig. 9. Plot of (22a), showing the variation in Q with length and diameter of coil, when the wire size is kept at the optimum value.

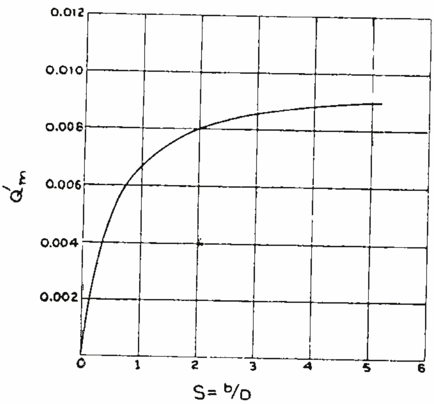

Fig. 10. Plot of (22b), showing the variation in Q with the length:diameter ratio, when the wire diameter is optimum and the coil diameter is constant.

For convenience in investigating the effect of changes in coil shape (22) is rearranged, giving

$$Q'_m = \frac{Q_m}{\dfrac{f^{1/2}}{\sqrt{2000}\ \rho^{1/2}}} = \frac{Db}{102b + 45D} \qquad (22a)$$

$$= \frac{D}{102 + 45/S} \qquad (22b)$$

Q'_m is, of course, directly proportional to Q_m. These relations are plotted in Figures 9 and 10. It is noted that, for a constant diameter of coil, the Q increases as the coil length is increased, but the increase is less rapid as large values of b/D are attained. Consequently, if an economic or a practical limitation is placed upon the design, there will be an optimum ratio.

Experimental data are shown in Figure. 11. The curve has been plotted from an equation of the form

$$Q_m = \gamma \frac{D}{102 + 45/S} \tag{22c}$$

where γ has been selected empirically to cause the curve to pass

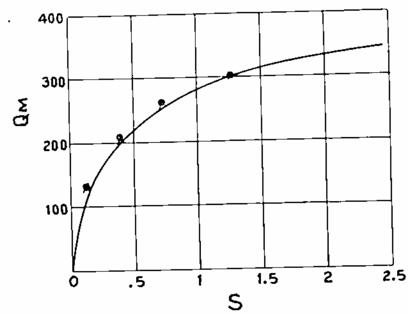

Fig. 11.—Relation between the length:diameter ratio and Q. when the wire size is kept optimum. The points are from experimental data; the curve shown was drawn from (22c), after γ had been selected to pass the curve through one of the experimental points. For this curve, $D = 2.5$ cm., $L = 1.1$ microhenrys, $f = 11$ mc.

through one of the experimental points. The curve, therefore, shows the agreement of the experimental data with the shape of (22b). If (22b) has been plotted directly, it would have shown exactly the same shape as (22c), but displaced in the positive ordinate direction, since it does not take account of the circuit losses, radiation losses and dielectric losses, which are all included in the experimental data.

Frequently, the maximum coil dimensions are fixed by the size of the coil shield, which, in turn, is limited in size by the space available in the apparatus. The presence of a shield about a coil reduces the Q, in general, due to the increased eddy current loss. It has been shown by several experimental and mathematical[15] studies that, if the coil diameter is less than half the shield diameter, and the ends of the coil are separated by at least a coil diameter from the ends of the shield, the Q of the tuned circuit is not reduced by more than 5 to 8 per cent. This restriction may be employed to determine the size of a coil enclosed in a shield.

It is of interest to note that, if the wire size, inductance and length: diameter ratio are kept constant, an optimum diameter of coil can be obtained for which the Q is a maximum. This is shown mathematically in Appendix II, and the optimum diameter is found to be

$$D_o = \sqrt[3]{\frac{10d^2L(102S + 45)}{S^2}} \tag{24}$$

A set of coils was wound which checks this relation very closely. The experimental results are plotted in Figure 12, from which the maximum Q is seen to obtain for a diameter of 4.0 centimeters. From (24) a value of 3.9 centimeters is obtained.

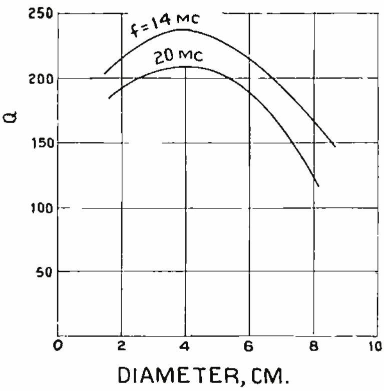

Fig. 12. Plot of Q against coil diameter, when the length:diameter ratio and the wire size are constant. For these experimental data, $b/D = 2.8$, $L = 1.1$ microhenrys, $d = 0.081$ cm.

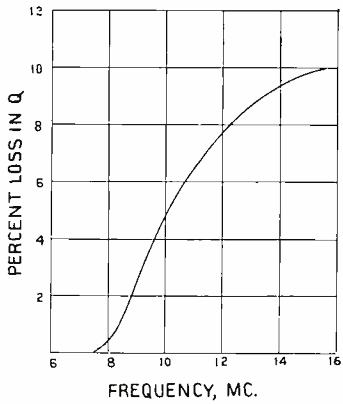

Fig. 13. Percent loss in Q resulting from grooved bakelite form. For this figure, $L = 3.6$ microhenrys, $D = 2.5$ cm., $b = 1.5$ cm., wire size, # 18 B&S.

Calculation of Number of Turns

The number of turns, for a given inductance, when the diameter and length of the coil are known, can be calculated from

$$N = \sqrt{\frac{L(102S + 45)}{D}} \tag{14}$$

which is derived in Appendix II. This equation will usually give the correct value within about 5 per cent, and the coil, when wound, can be adjusted to the required value in any of the usual ways.

Dielectric Losses

The magnitude of the dielectric losses in the coil form and in the enamel wire insulation for typical coils at high frequencies was investigated. Examples are given herewith. To measure the dielectric losses in the coil form 2 similar coils, of 2.5 centimeter diameter, were wound, one using a grooved bakelite form with a groove about 0.04

centimeters deep, the other wound self-supporting, except for 3 narrow celluloid strips to which the wire was fastened with collodion. The percentage loss in Q resulting from the grooved bakelite form is shown in Figure 13. At 14 megacycles the loss is less than 10 per cent.

The curves of Figure 14 are for a set of three similar coils, of 5-centimeter diameter, one wound without any supporting material—similar to the coil described in the preceding paragraph—one wound on a grooved bakelite form, and the third on a grooved cardboard form about 1 centimeter in thickness—worse than any case one would expect to find in practice. The groove in each case was about 0.04 centimeter deep. At 13 megacycles the reduction in Q resulting from the heavy cardboard form was 27 per cent. A wood form, used for another coil. gave results similar to those obtained for the cardboard. The loss in

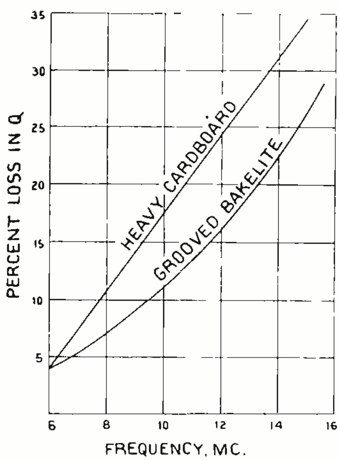

Fig. 14. Percent loss in Q resulting from heavy cardboard form and grooved bakelite form. For this figure, $L = 3.6$ microhenrys, $D = 5$ cm., $b = 1.5$ cm., wire size, No. 32 B&S.

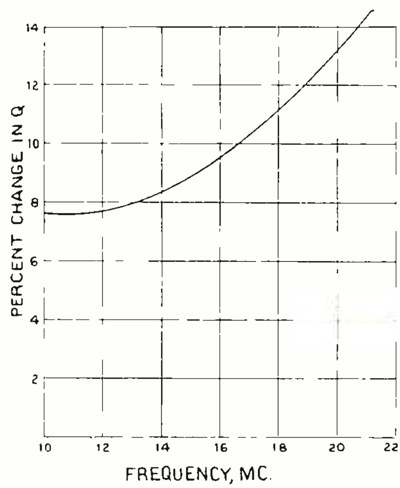

Fig. 15. Percent loss in Q resulting from groove in bakelite form. No measureable loss resulted from the use of a smooth bakelite form. For this figure $D = 2.5$ cm., $b = 1.9$ cm.. wire size, No. 14 B&S.

Q resulting from the grooved bakelite form was 19 per cent at 13 megacycles.

To measure the dielectric loss in the coil form at higher frequencies, a group of 1.1 microhenry coils were wound, one on a smooth bakelite form, one on grooved bakelite with a groove about 0.04 centimeter deep and one without a supporting form, as described previously. The difference between the smooth bakelite and the air core samples was smaller than the experimental error. The loss in Q resulting from the groove in the bakelite is plotted in Figure 15, and amounts to 13 per cent at 20 megacycles.

To investigate the dielectric loss in the enamel wire insulation, two similar coils were wound, on smooth bakelite forms, one using enameled wire, the other bare copper wire carefully cleaned to remove corrosion. The per cent loss in Q resulting from the enamel is plotted in Figure 16 and amounts to about 6 per cent at 20 megacycles. This result is quite reasonable, since the dielectric path in a spaced winding is largely in air. Below 5 megacycles the dielectric loss in the wire insulation is difficult to detect. After a time bare copper wire corrodes and the dielectric loss in the enamel insulation may become smaller than the loss resulting from corrosion.

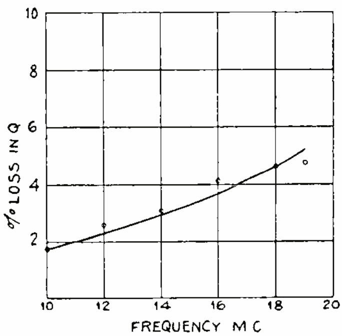

Fig. 16. Percent loss in Q resulting from enamel wire insulation. For this figure, $L = 1.1$ microhenrys, $D = 2.5$ cm., $b = 3.4$ cm., wire size, No. 18 B&S.

These data indicate the unimportance of the dielectric losses in determining coil performance. The choice of coil form material, within reason, and the use of enameled wire, have little effect on the merit of the coil. A loss of less than 10 or 15 per cent can be expected for small diameter coils, if a deeply grooved form is used. A shallower groove may be used with little sacrifice in rigidity and an improvement of a few per cent in the figure of merit. An ungrooved bakelite form causes very little loss in Q.

Summary

The merit of coils for frequencies above 4 megacycles is determined largely by the winding design. A suitable procedure for the optimum design of coils for these frequencies is:

1. Coil diameter and length of winding: Make as large as is consistent with the shield being used. The shield diameter should be twice the coil diameter and the ends of the coil should not come within one diameter of the ends of the shield.

2. A bakelite coil form with a shallow groove for the wire, and enameled wire may be used with little loss in the Q. The groove should not be any deeper than is necessary to give the requisite rigidity. The use of special coil form constructions and special materials does not appear to be justified.

3. Number of turns: Calculate from

$$N = \sqrt{\frac{L(102S + 45)}{D}} \tag{14}$$

4. Wire size: Calculate from

$$d_o = \frac{b}{\sqrt{2}\,N} \tag{17}$$

ACKNOWLEDGMENT

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APPENDIX I—NOTATION AND UNITS

b = length of winding, centimeters
C = capacitance, micromicrofarads
C_o = self-capacitance of coil, micromicrofarads
d = diameter of wire, centimeters
d_o = optimum wire diameter, centimeters
D = diameter of coil, centimeters
D_o = optimum coil diameter, centimeters
f = frequency, cycles per second
F = skin effect factor, see (1)
G = proximity effect factor, see (1)
$j = \sqrt{-1}$
K = function of S, see (1)
L = inductance, microhenrys
N = total number of turns of wire on coil
Q = figure of merit of tuned circuit
Q_m = figure of merit for optimum wire size
R = effective resistance of tuned circuit or of coil, ohms
R' = effective resistance of coil, including effect of self-capacitance
R_c = effective resistance of condenser
R_d = d-c resistance of coil, ohms

R_L = effective resistance of coil
S = ratio of length to diameter of coil $= b/D$
X = reactance of coil ohms
Z = skin effect coefficient, see (4)
γ = empirical constant in (22c)
ω = angular velocity $= 2\pi f$
ρ = specific resistivity of conductor, in ohms/cm. cube
For copper, $\rho = 1.72 \times 10^{-6}$ ohms/cm. cube

APPENDIX II—DERIVATION OF EQUATIONS

Butterworth[3] has shown that the effective resistance of a coil, including the skin and proximity effects, may be expressed by the relation

$$R = R_d \left[1 + F + \tfrac{1}{4} \left(\frac{KNd}{D} \right)^2 G \right] \tag{1}$$

where, at high frequencies,

$$1 + F = \frac{\sqrt{2}\,Z + 1}{4} \tag{2}$$

$$G = \frac{\sqrt{2}\,Z - 1}{8} \tag{3}$$

and

$$Z = \pi d \sqrt{\frac{2f}{10^9 \rho}} \tag{4}$$

The quantities are expressed in practical cgs units. K is a factor depending upon the ratio of coil diameter to length.

For copper wire

$$Z = 0.1 d \sqrt{f} \tag{5}$$

or, for #20 B&S wire

$$Z = 0.008 \sqrt{f} \tag{6}$$

and, for frequencies greater than 4 megacycles,

$$Z > 16$$

Consequently, for the high frequency case considered in this paper, (2) and (3) may be written, without much error

$$1 + F = \frac{\sqrt{2}}{4} Z = \frac{\pi d}{2} \sqrt{\frac{f}{10^9 \rho}} \tag{7}$$

$$G = \frac{\sqrt{2}}{4} Z = \frac{\pi d}{4} \sqrt{\frac{f}{10^9 \rho}} \tag{8}$$

Now, if the turns of the coil are assumed to be circular, which is not exactly true.

$$R_d = \frac{\rho l}{A} = \frac{4\rho DN}{d^2} \tag{9}$$

The factor, K, can be computed for any value of $S = b/D$ from series expansions. A simple empirical equation for K has been found from the expansions. This relation is

$$K = \frac{4}{S} \tag{10}$$

For values of S between 0.25 and 1.00 this simple expression gives K to within 3 per cent of its correct value.

Equations (7), (8), (9) and (10) may be substituted into (1), giving an expression for the wire resistance of a coil at high frequencies, in terms of its physical dimensions.

$$R = 2\pi \sqrt{10^{-9}\rho f} \left[\frac{DN}{d} + \frac{2N^3 d}{S^2 D} \right] \tag{11}$$

If the coil is used at a frequency near its natural resonant frequency, a correction should be made if the apparent resistance of the coil and its self-capacitance is desired. The correction is given by

$$R' = \frac{\mathrm{R}}{(1 - \omega^2 L C_o)^2} \tag{12}$$

assuming that the self-capacitance is lumped across the terminals of the coil.

It was found convenient, also, to be able to express the inductance of a coil as a function of the coil dimensions. For this purpose, Rayleigh's equation[16]

$$L = DN^2 \left[2\pi \log_e \frac{4D}{b} - \frac{1}{2} + \frac{b^2\pi}{4D^2} \log_e \left(\frac{4D}{b} + \frac{1}{4} \right) \right] \tag{13}$$

was found suitable. This equation, in common with most inductance formulas, was originally derived assuming the coil to have the form of a cylindrical current sheet, and consequently does not take account of the size of the wire or the frequency. A relatively simple empirical expression was found for the bracketed term, giving,[17]

$$L = \frac{DN^2}{102S + 45} \tag{14}$$

This agrees with Rayleigh's equation within 2 per cent for values of S between 0.25 and 1.00. Rayleigh's equation gives values of inductance within better than 3 per cent of their true values.

If the value of N which is obtained from (14) is substituted in (11), the result is

$$R = 2\pi\sqrt{10^{-9}\rho f}\left[\frac{L^{1/2}D^{1/2}(102S+45)^{1/2}}{d} + \frac{2dL^{3/2}(102S+45)^{3/2}}{S^2D^{5/2}}\right] \quad (15)$$

To find the optimum wire size for fixed values of L, S and D, (15) can be differentiated and the result equated to zero; that is, take

$$\frac{\partial R}{\partial d} = 0$$

If this operation is performed, the result obtained for the optimum wire size to give minimum resistance, is

$$d_o = b\sqrt{\frac{D}{2L\,(102S+45)}} \quad (16)$$

If the value of L from (14) is substituted in this equation, the particularly simple relation

$$d_o = \frac{b}{\sqrt{2}\,N} \quad (17)$$

is obtained. Differentiating (11) would yield this result directly. These results can be substituted in (15) for d to give the resistance of a coil, when the wire size is always optimum,

$$R_m = 4\pi\sqrt{2\times 10^{-9}\rho f}\left[\frac{L\,(102b+45D)}{Db}\right] \quad (18)$$

or

$$R_m = 4\pi\sqrt{2\times 10^{-9}\rho f}\,\frac{N^2}{S} \quad (19)$$

These equations may be applied directly to the prediction of Q, as well as of R. Substituting the value of R from (11) into

$$Q = \frac{X}{R}$$

gives, for any wire size

$$Q = \frac{f^{1/2}\,LdS^2D}{10^{3/2}\,\rho^{1/2}\,(S^2D^2N + 2N^3d^2)} \quad (20)$$

The use of (15) yields

$$Q = \frac{f^{1/2}\,L^{1/2}\,dS^2D^{5/2}}{10^{3/2}\,\rho^{1/2}\left[S^2D^3(102S+45)^{1/2} + 2d^2L\,(102S+45)^{3/2}\right]} \quad (21)$$

In the same way, the value of the figure of merit when the optimum wire size is employed, Q_m, is, from (18),

$$Q_m = \frac{f^{1/2}\, Db}{20^{3/2}\, \rho^{1/2}\, (102b + 45D)} \tag{22}$$

or, from (19)

$$Q_m = \frac{f^{1/2} L\, S}{20^{3/2}\, \rho^{1/2}\, N^2} \tag{23}$$

(15) may also be differentiated with respect to D, holding L, S and d constant, giving, for the optimum diameter

$$D_o = \sqrt[3]{\frac{10d^2 L (102S + 45)}{S^2}} \tag{24}$$

Note that the equations for the optimum wire size (7) and (8) are not consistent with (24) because a different set of variables is assumed in each case.

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[18] TABLE I.—EXPERIMENTAL VERIFICATION OF (16) AND (17).

Coil Description			*Optimum Wire Diam.*		*Percentage Difference*
Induc. *μh.*	*Diam.* *cm.*	*Length* *cm.*	*From Eq.* *cm.*	*From Exp't* *cm.*	*from Experimental Figure*
3.6	2.5	1.55	0.090	0.102	12 per cent
3.6	3.8	1.45	0.108	0.115	6 " "
3.6	2.5	1.55	0.090	0.115	22 " "
1.1	2.5	0.89	0.105	0.100	5 " "
1.1	2.5	0.56	0.079	0.093	15 " "
1.0	1.3	0.69	0.055	0.064*	13 " "
1.0	3.8	0.79	0.142	0.134*	6 " "

* From data in Barden and Grimes' papers.

CATHODE-RAY ENGINE-PRESSURE MEASURING EQUIPMENT

By

H. J. Schrader

Victor Division, RCA Manufacturing Company, Inc.

DURING the past year the use of electronic devices in the field of testing other than electrical equipment has greatly increased. At least a portion of this progress has been due to the introduction of dependable cathode-ray oscillographs. This instrument and its many advantages over string oscillographs has long been known,

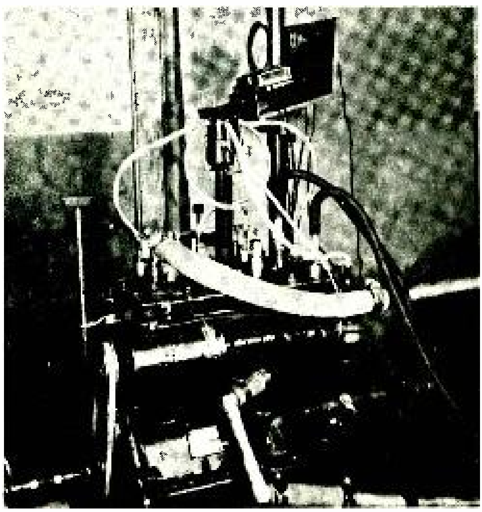

Fig. 1—Pressure-sensitive element attached to head of 4-cylinder test engine.

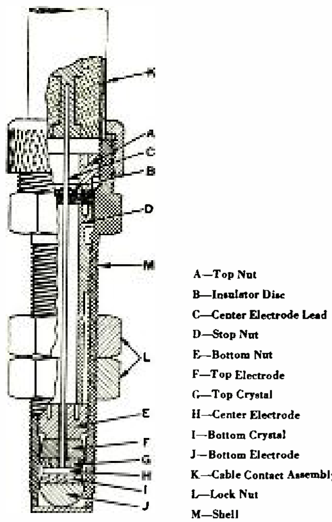

Fig. 2

but it is only during the past two or three years that cathode-ray tubes with high brilliancy and long life have been available. Since these tubes and oscillographs have been commercially introduced, a large number have been used in the study of vibrations, sound phenomena, welding, timing of electrical and mechanical circuits and many other phenomena.

Recently equipment using the cathode-ray oscillograph has been introduced to the automotive engineer for the study of engine-pressure diagrams. This equipment. because of its ability to picture each individual explosion in the cylinder, offers many advantages and opens up an entirely new field of investigation of internal-combustion engines. With this equipment it is possible to study roughness of operation at light engine loads, operation during acceleration, and operation of carburetion and ignition during each successive cycle at any load and speed condition. The same equipment may be used on either spark or compression-ignition engines without changes.

The equipment consists of four units; the pressure-sensitive unit, an amplifier, a synchronizer or timing circuit and the oscillograph. Figure 1 illustrates the attachment of the pressure-sensitive element to the head of a 4-cylinder, spark-ignition engine. As may be seen, special cylinder-head adapters for insertion of the pressure units must

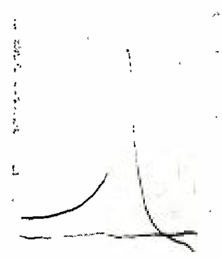

Fig. 3 — Engine-pressure diagram from 4-cycle engine operating at 1200 r.p.m.

be installed in the head before the equipment in its present form may be used. However, this is not difficult in the case of test engines.

The pressure-sensitive unit utilizes the piezo-electric property of quartz crystals for the conversion from mechanical pressure to electrical energy. This unit shown in cross section in Figure 2 consists of two quartz crystals mounted in a small cylindrical casing between two grounded electrodes. A third electrode insulated from ground by the crystals themselves is inserted between the two crystals and is the "high" electrical connection to the unit. The unit is assembled by means of the backing-screw plug and lock-screw plug. Gasses are excluded from the unit by the steel diaphragm, this end of the plug being in the explosion chamber, and the unit sealed against moisture and dirt on the other end by the mica washer.

When pressure is exerted on the diaphragm it is transferred by means of the steel hemisphere to the crystals which are thus stressed and an electrical charge appears on their surfaces. The crystals have been so cut and assembled that a similar polarity charge will appear on the two grounded surfaces and the opposite polarity charge on the two center surfaces. The two crystals are thus operating electrically in parallel.

If the insulation resistance of the center electrode were infinite this charge would be maintained indefinitely. However, some electrical leakage is always present in this unit and the associated amplifier and this leakage, after being reduced to a minimum, is the determining factor in the low-speed performance of the equipment. Because of the fact that the unit is quartz and mica insulated its leakage resistance is several thousand megohms. The time constant of the circuit is still further improved by placing in parallel with the crystals a padding capacitor. The effect of this capacitor, of course, is to decrease the voltage appearing between the terminals by the ratio of the crystal capacity to the padding capacity.

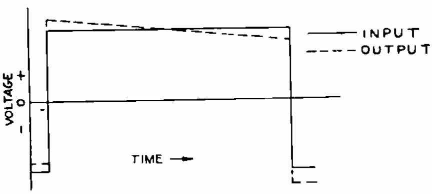

Fig. 4—Effect of coupling capacitor on response to square waves.

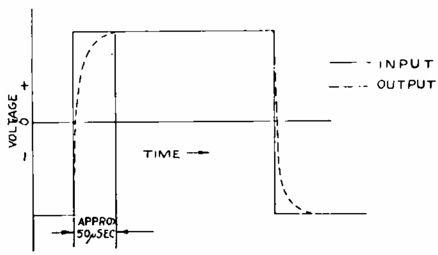

Fig. 5—Effect of circuit capacity on response to square waves.

The unit operating with a parallel capacitor of 10000 $\mu\mu f$ produces a voltage of approximately 2.5 millivolts per hundred pounds per square inch pressure. The sensitivity of the unit is unaffected by temperature up to 350 degrees centigrade. Above this temperature the sensitivity drops rapidly until at about 573 degrees it is zero. At this temperature the quartz changes from what is known as Alpha to Beta quartz. However, as soon as cooled the quartz returns to its previous form and the sensitivity of the unit returns to its previous value. When used in conjunction with the input circuit as employed in the special amplifier, next to be described, the time constant of the crystal circuit is such as to cause less than 5 per cent leakage in 1/10 second. This provides very good operation at speeds down to 1200 r.p.m. and only slight distortion of the pressure diagrams down to 800 r.p.m. with 4-cycle engines. With 2-cycle engines the speed at which it may be operated is, of course, reduced by two.

The amplifier to be used with the pressure-sensitive element has characteristics quite different from those of normal high-quality amplifiers. The wave form to be amplified contains not only a wide range of frequencies, but also it is practically of uni-polarity. Figure 3 is a photograph of an engine-pressure diagram obtained from a 4-cycle engine operating at 1200 r.p.m. As may be seen, one-half of the pressure wave has practically zero-pressure change. To properly amplify this wave form it has been found necessary to design the amplifiers to amplify square wave forms without distortion. A 4-cycle engine operating at 1200 r.p.m. produces one power wave for each two revolutions, the pressure wave occurring during one revolution and the exhaust and intake, when very little pressure variation exists, during the other revolution. Thus, at 1200 r.p.m. all time constants must be such as to cause negligible voltage change during 1/10 second or electrically to amplify a 10-cycle square wave form.

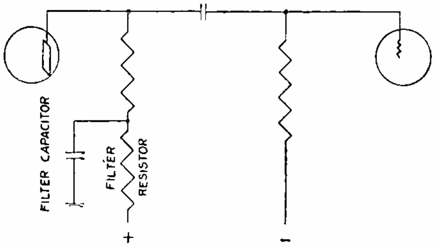

Fig. 6—Circuit for low-frequency compensation.

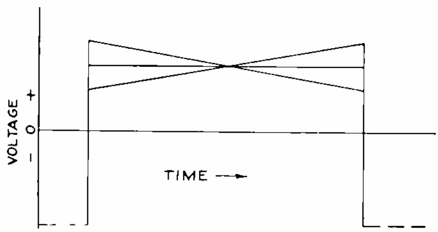

Fig. 7—Response of Circuit with low-frequency compensation.

The fidelity or amplitude-frequency characteristic of the usual resistance capacitance-coupled amplifier shows a drooping characteristic at both the high and low frequencies. The high-frequency droop is caused by distributed capacity in the circuit and the grid and plate capacities of the amplifier tubes. The low-frequency droop is caused by the relative impedance of the coupling capacitors and the grid resistors. The fidelity curve may be flattened considerably if certain means are employed to compensate for these losses.

Experience has proven that one of the best tests of amplifiers of this type is made by checking their ability to pass square wave forms without distortion. Figure 4 illustrates the effect of the coupling capacitors on this type of wave and Figure 5 illustrates the effect of the shunting capacities in the circuit.

Low-frequency distortion can, of course, be eliminated by the use of very large coupling capacitors and high-resistance grid resistors. However, the required capacitors become quite large mechanically if the low-frequency cutoff of the amplifier is to be located at less than one cycle. The grid resistor, of course, is limited in value by the grid currents of the amplifier tube. Low-frequency compensation is accomplished as shown in Figure 6 and the effect of this compensation is shown in Figure 7. Too large a compensating capacitor causes the curve to drop on the right and too small a compensating capacitor causes it to rise. By the correct choice of the coupling capacitor and the compensating capacitor and resistor, the zero-change portion of the square wave will be parallel with the zero axis.

High-frequency distortion is compensated for by the use of the circuit as shown in Figure 8. The effect of the plate choke is illustrated in Figure 9. The proper choice of this compensation choke is affected by the distributed capacities in the circuit and the value of the plate and grid resistors. By the proper choice of these components the square corners of the wave may be reproduced on a cathode-ray oscillograph with little visible distortion.

Both of these types of compensation are employed in the amplifiers of this equipment to insure the undistorted amplification of the pres-

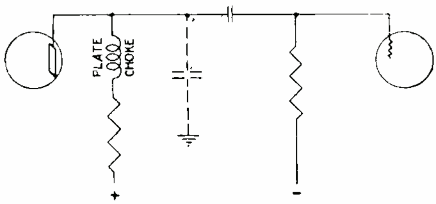

Fig. 8—Circuit for high-frequency compensation.

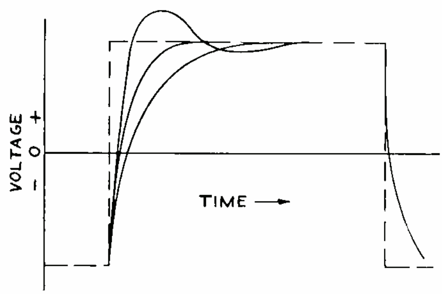

Fig. 9—Response of circuit with high-frequency compensation.

sure wave. Figure 10 is the circuit diagram of the complete amplifier and includes an electronic voltage-regulation circuit. This, besides eliminating changes in gain due to a-c line-voltage variation, also decreases the hum contents of the rectified d-c voltages. The frequency characteristic of this amplifier as tested in the normal manner with sine-wave input is shown in Figure 11. As may be seen this characteristic extends well below 1-cycle and above 17000 cycles.

As mentioned during the description of the pressure element, resistance across the crystals is a serious consideration. One of the most serious sources of this leakage resistance is the grid resistance of the first tube in the amplifier. The grid resistance of a vacuum tube is usually considered as being infinite. However, this is far from correct when we consider using grid leaks of the order of 100 megohms. Due to the heating of the grid by the filament and due to secondary emission of the grid caused by bombardment, the grid resistance may drop so low as 10 or 20 megohms. Lowering the heater voltage, the screen-grid voltage and the plate voltages have all been found to increase the grid resistance. A 6D6 tube is used as the input tube of this amplifier and by operating this tube with 4 volts filament supply, 22½ volts on the screen grid and 135 volts plate supply the grid resistance has been found to be of the order of 1000 megohms. Thus, using a 100-megohm grid resistor the total output resistance is approximately 90 megohms.

The voltage gain of the amplifier, about 100,000 maximum, is controlled by varying the bias voltage applied to the No. 1 and No. 3 grids of the second amplifier tube—a 6L7. This arrangement provides a variable gain of approximately 100 to 1 without producing appreciable distortion. The third stage of amplification employs a 6C6 which pro-

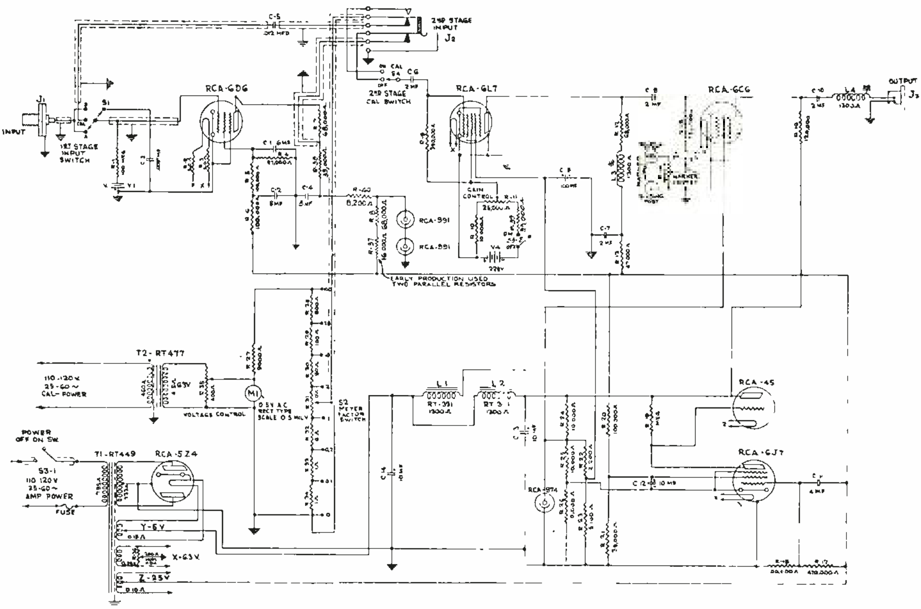

Fig. 10—Schematic circuit diagram of the amplifier with electronic voltage regulator.

vides ample output voltage for direct deflection of a 3-inch cathode-ray tube.

Provisions are made for calibration of the amplifier by including a 60-cycle variable voltage source which may be switched to the input of either the first or second stage of the amplifier. Thus, when a cathode-ray oscillograph is connected across the output of the amplifier it is always possible to calibrate the equipment in terms of volts-per-inch deflection. The pressure unit being factory calibrated in terms of volts per-pound per-square-inch of pressure, it is thus possible to calibrate the oscillograph vertical-deflection in terms of inches deflection per pound-per-square-inch of pressure.

Two different cathode-ray oscillographs are available for use with this equipment, one of the three-inch tube type and the other using a nine-inch tube. Both types of oscillographs are complete units, being a-c operated and include a timing axis, and horizontal and vertical amplifiers. When used for engine-pressure testing the choice between the two is entirely one of size of picture.

The cathode-ray oscillograph which uses a 3-inch cathode-ray tube has a vertical and a horizontal amplifier, both with flat frequency characteristics from 4 to 90,000 cycles and a timing axis with a frequency range of 4 to 18,000 cycles. When used with the special amplifier the vertical amplifier is not used, the output of this amplifier being directly connected to the vertical deflecting plates of the cathode-ray

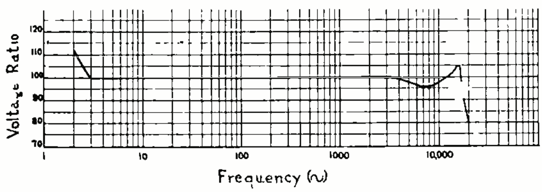

Fig. 11—Frequency characteristic under normal test.

tube. The timing-axis oscillator is employed to produce the horizontal deflection of the tube and pressure-time diagrams of engine cylinder thus pictured. The frequency of this oscillator is synchronized with engine rotation by means of the synchronizer unit to be described later.

The cathode-ray oscillograph employes the 914 (9-inch) cathode-ray tube. Vertical and horizontal amplifiers are provided with a flat frequency response range of from a few cycles to 300,000 cycles. The low-frequency response of these amplifiers is better indicated by the fact that they will faithfully reproduce square wave forms at 10 cycles per second. The timing-axis oscillator has a frequency range of from 5 to 50,000 cycles per second. The vertical amplifier of this oscillograph is always used as considerable voltage is otherwise required to produce full deflection of the cathode-ray tube. However, due to the compensation employed no distortion of the engine-pressure wave is produced. The input sensitivity of the oscillograph may be varied by means of frequency-compensated gain controls from maximum of 0.3 volt per inch to a minimum of 100 volts per inch. The complete circuit diagram of the oscillograph is shown in Figure 12.

The synchronizer unit as a part of the equipment has two purposes; synchronization of the timing-axis oscillator of the oscillograph with the rotation of the engine under test and furnishing an angle of rotation indicator. The unit consists of two simple magnetic alternators,

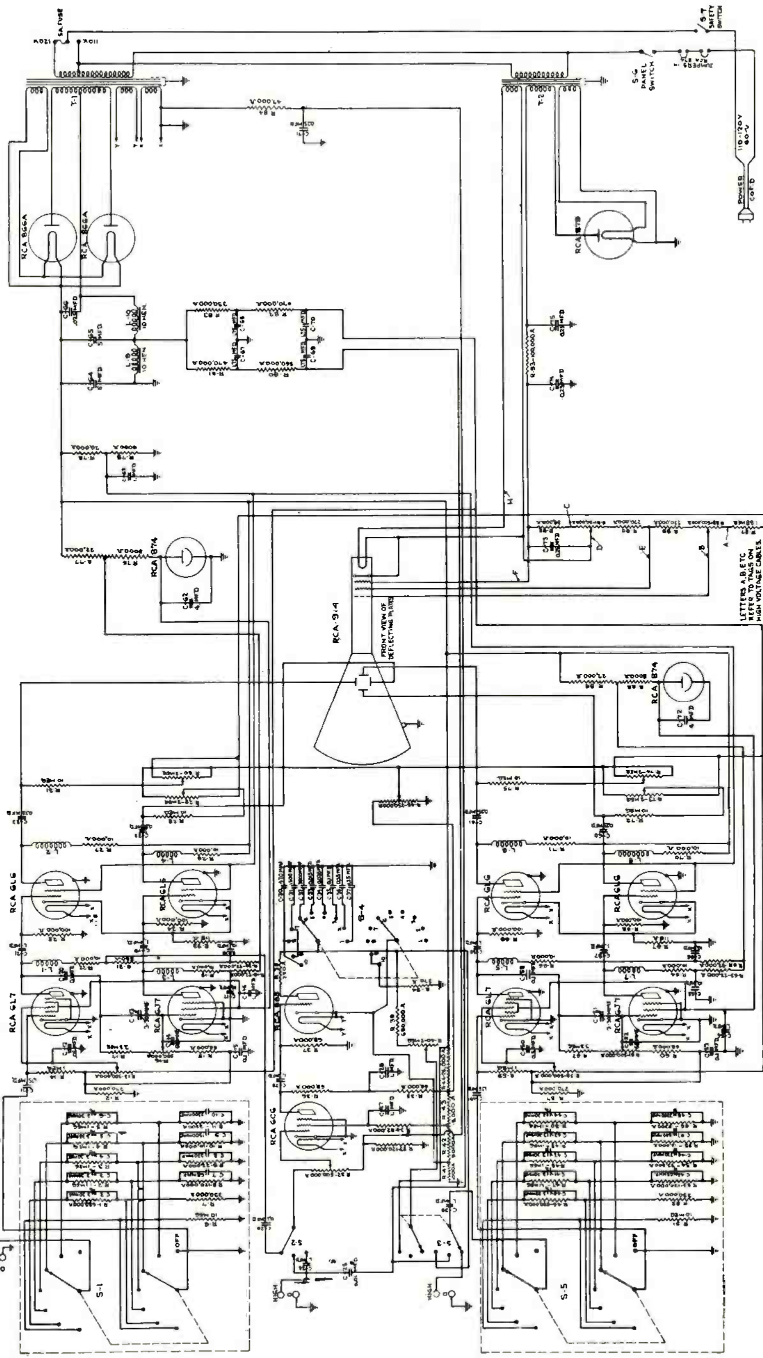

Fig. 12—Schematic circuit diagram of the oscillograph.

one producing 6 cycles per revolution, and the other two cycles. (Figure 13). The six-pole alternator is used for synchronization and the two-pole alternator for the marker voltage.

The synchronizing alternator consists of a rotor similar in appearance to a six-toothed gear which revolves between two pole pieces of a permanent magnet. A pickup coil, consisting of a large number of turns of wire, is wound on the magnet and when connected to the external synchronizing binding posts of the oscillograph effects the synchronization of the timing-axis oscillator. The position of the pole pieces with reference to the angle of rotation is variable so that the pressure picture may be "framed" on the oscillograph screen.

The marker voltage alternator is similar in construction to the synchronizing alternator. However, the rotor is cylindrical in form and

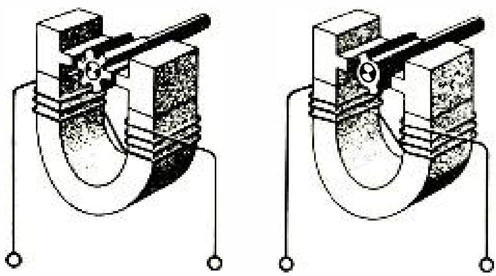

Fig. 13—The synchronizer unit.

has two narrow diagonally opposite slots cut in its circumference. This type rotor construction was found to give a more peaked wave form and thus a more accurate angle measurement. The permanent magnet assembly is rotatable similarly to that of the synchronizing alternator, but the angle of rotation is indicated by a dial on the end of the unit which is calibrated in degrees. When the unit is connected to the engine under test, the connection is tightened with the engine at top dead center, the dial on zero, and with the slots in the rotor and the stator pole pieces lined up. Thus, the center of the marker voltage wave will occur at engine top dead center. The pickup coil of the marker alternator is connected into the cathode circuit of the second amplifier tube of the special amplifier. Thus, this marker voltage is superimposed on the pressure generated voltage. The dial may thus be rotated and angles of rotation with reference to top dead center indicated on the pressure diagram.

It is usually desirable while making engine tests to make a permanent record of the pressure diagram under various running conditions. This may be readily accomplished by photography. A camera with an f:3.5 lens and using supersensitive panchromatic film is capable of photographing the picture on the screen of either the 3- or 9-inch tubes with a 1/10 second exposure. Figures 3 and 14 to 20 are unretouched photographs of pressure diagrams taken with a camera having an f:3.8 lens. Figure 3 represents the pressure diagram of an engine running 1200 r.p.m. with an open throttle and was taken with an exposure of 1/10 second. Figure 14 is the same engine accelerating with light load. A one-second exposure was used so that a number of individual pressure waves were obtained. The variation of peak pressure

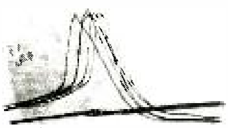

Fig. 14

Fig. 15

Fig. 16

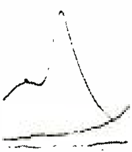

Fig. 17

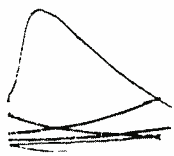

Fig. 18

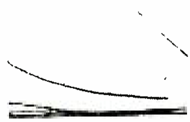

Fig. 19

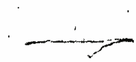

Fig. 20

Various engine-pressure diagrams.

and form of the pressure wave from cycle to cycle may be readily seen. Figure 15 was also exposed 1 second and shows the pressure variations with the engine accelerating with full load. Detonation as indicated by the high frequency superimposed on the pressure wave to the left of the peak pressure may be seen. This may be seen more clearly in Figure 16 which represents a steady running condition similar to Figure 3, but with the spark slightly more advanced.

Figure 17 is a 180-degree engine diagram with the spark retarded. The pressure has been built up to a peak before ignition occurs and actually drops before the gas starts to burn. The break at the top of the first peak is top dead center and ignition was set to occur at this same point.

Figure 18 is similar to Figure 3 but each horizontal line represents 90-degree crankshaft rotation. The shape of the pressure wave may now be readily studied. If the spark is advanced until detonation occurs, this diagram becomes that of Figure 19. Here the detonation pressure wave may be readily observed. If a small coupling capacitor is introduced between the amplifier and the oscillograph, the picture on the screen approaches a rate-of-change diagram. Figure 20 illustrates this effect.

The complete equipment has great versatility, being operative over a speed range of a 4-cycle engine of from 1200 r.p.m. to higher than present-day engines are capable of operating. The equipment may be used at speeds down to 600 r.p.m. if a small amount of distortion is not objectionable. Pressure curves as low as 100 pounds per square inch may be readily reproduced and use of the gain control allows pressure curves with peaks of 5000 pounds per square inch to be seen without producing overload of any of the components. By the addition of a second parallel amplifier input condenser this upper limit may be increased as desired. Temperatures up to 350 degrees C. have negligible effect on the sensitivity of the pressure element so water or air cooling is unnecessary even when operating with air-cooled engines which often operate at comparatively high temperatures. It is often desirable to picture the rate of change of pressure diagrams. A partial-derivative curve may be readily produced by coupling the amplifier to the oscillograph through a small capacitor.

TECHNICAL EDUCATIONAL REQUIREMENTS OF THE MODERN RADIO INDUSTRY

Part II: The Field of Receiver Servicing

By

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IN THIS second of a series of papers on "Technical Educational Requirements of the Modern Radio Industry," the first of which appeared in the April 1937 issue of the RCA Review, the writer wishes to further illustrate the need for higher educational standards for the receiver service technician.

This need exists because current types of radio receivers and associated apparatus have become very complex devices. With the past as a guide, it is logical to assume that the designs of next year and the years to follow will be more advanced and complicated. The technician's training must be of such a caliber that it will make him competent to handle the more advanced types of equipment.

It is possible to train a practical technician to perform satisfactorily a given intricate operation (at which he can become quite expert) on one model receiver. But, if the receiver replacing this model the following year incorporates circuits of a different basic design, the training he received on the current receiver will not enable him to decide upon a method of procedure on the newer model unless he has sufficient background to understand all the factors involved in its operating theory.

To make such a technician as expert on the new product as he was on the old one would require that he again be given a training course on the exact procedure to be followed on the newer product. On the other hand, the technician whose training is complete, even to the point of enabling him to analyze the circuit in a quantitative way, would be able to decide upon the possibility of applying some of those methods of approach, test, or procedure used in the past. Not only would he be able to appreciate the *need* for new methods of test, but he could also *devise* new methods of test or adjustment procedure. He would also be able to determine in advance the effect that a change in the value of circuit components would have on the operating characteristics of that circuit. This would make it possible for him, on finding the operating characteristics changed, to decide which altered circuit components were responsible for such change. Such an individual would, on receipt of data on a newly designed circuit, be able to analyze its operating characteristics, decide upon a practical method of test, and estimate the effect of change or failure of all circuit components. Thus he would be able to locate and remedy faults in the shortest possible time.

In the past, due in some cases to economic conditions, men with only the minimum of background and training were able to enter the receiver and associated service fields as technicians. Some of these men have developed with the industry, improving their knowledge of

Fig. 1—A standard radio-service-station test and work bench.

the science by a persistent program of self-study and, in some instances, education in schools of higher learning. These have been suitably rewarded for their efforts. However, a large number, due often to conditions beyond their control, are no better equipped today than they were when they entered the field. Among this group, the mortality in the past five years has been great indeed, and the large number who still remain find their daily work one of bewilderment and doubt, for they find themselves constantly confronted with problems that are plainly beyond their ability to understand, let alone solve.

To illustrate further the extent of technical knowledge required, let us consider the circuit diagram of one of the larger all-wave receiver and phonograph combinations of 1937. Figure 2 shows the schematic circuit diagram of a 15-tube superheterodyne receiver and phono combination. This unit incorporates such innovations as variable-permeability iron-core i-f transformers, automatic volume control, audio-range or volume expansion, variable-band-width selectivity accomplished by coupling a tertiary winding to the first i-f transformer, and automatic tone compensation at different volume levels.

Due to the high degree of fidelity of this unit, a small change in the quality of reproduction (which would go unheeded in an ordinary receiver) will cause the discriminating owner to call a technician to remedy the change. The cause may be anything from a tube whose characteristics have changed slightly from normal to improperly adjusted calibrating or alignment controls. To determine this cause would require a major service operation.

It is not uncommon today for a technician to be told by a client that "the quality of reproduction is not as good as it originally was." Yet, in many instances, the technician's reaction is, "It sounds good to me." Though the quality may compare satisfactorily with the average receiver, there has been a loss in fidelity as compared to the original factory specifications.

The technician must therefore realize that he cannot depend upon his physical reactions when dealing with the modern receiver. He must make actual measurements in order to determine the quality of reproduction. He also must appreciate, after having located and replaced parts which have changed value or failed, that he must check the receiver's alignment, selectivity and the fidelity of the audio channel before he can consider the service operation completed.

To perform all these operations it is necessary that he have available:

1st—A means of checking the characteristics of tubes with reasonable accuracy and determining small changes in their G_m.

2nd—A means of checking the d-c resistance of parts with values ranging from a fraction of an ohm up to more than 10 megohms.

3rd—A means of checking the capacitance and leakage resistance of capacitors at working voltages.

4th—A means of checking operating voltages and currents, both a.c. and d.c.

5th—A cathode ray oscilloscope with linear time base which is sufficiently flexible to be used in d-c, r-f and a-f circuits.

6th—A frequency and amplitude-modulated r-f all-wave signal generator.

7th—A beat-frequency a-f oscillator.

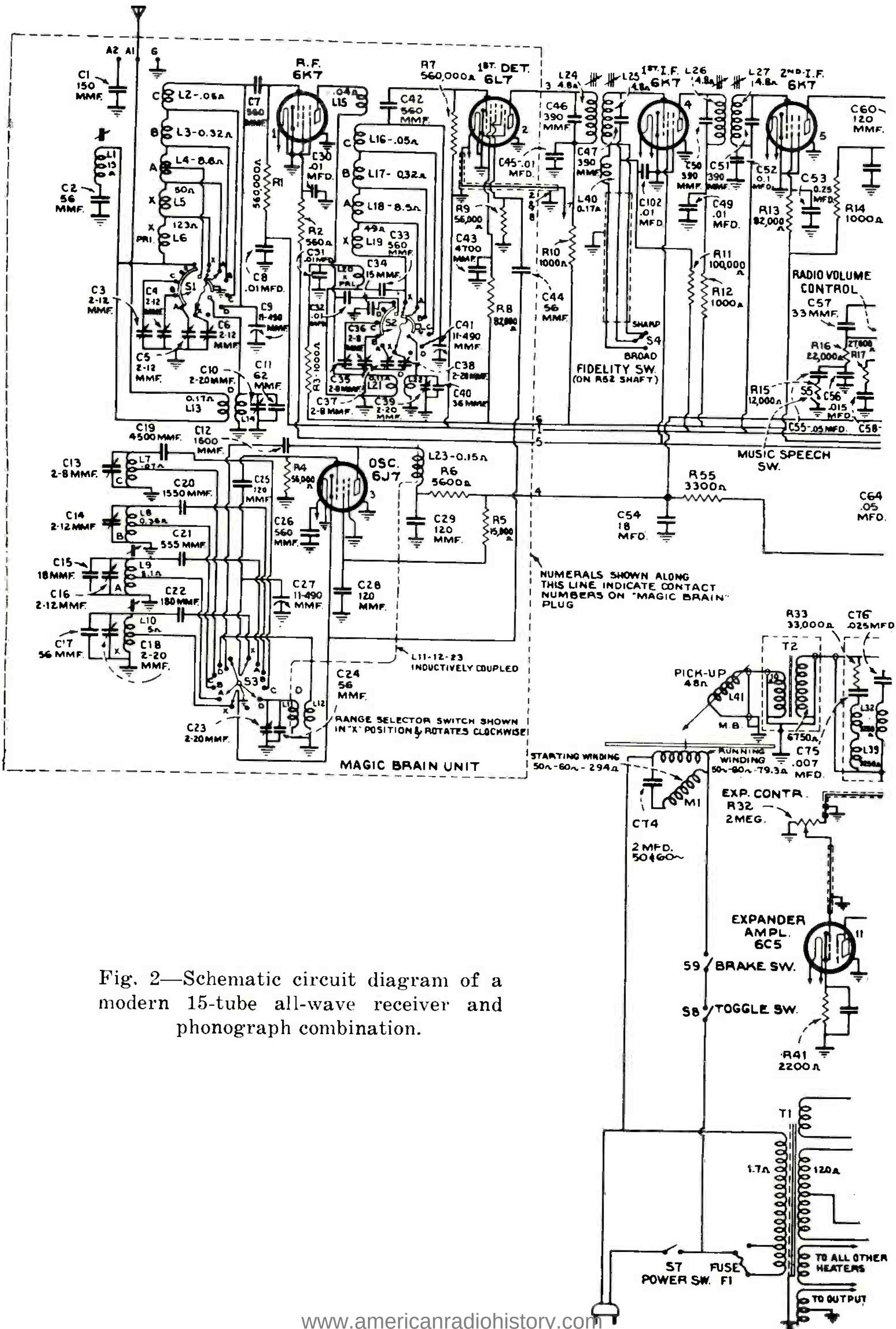
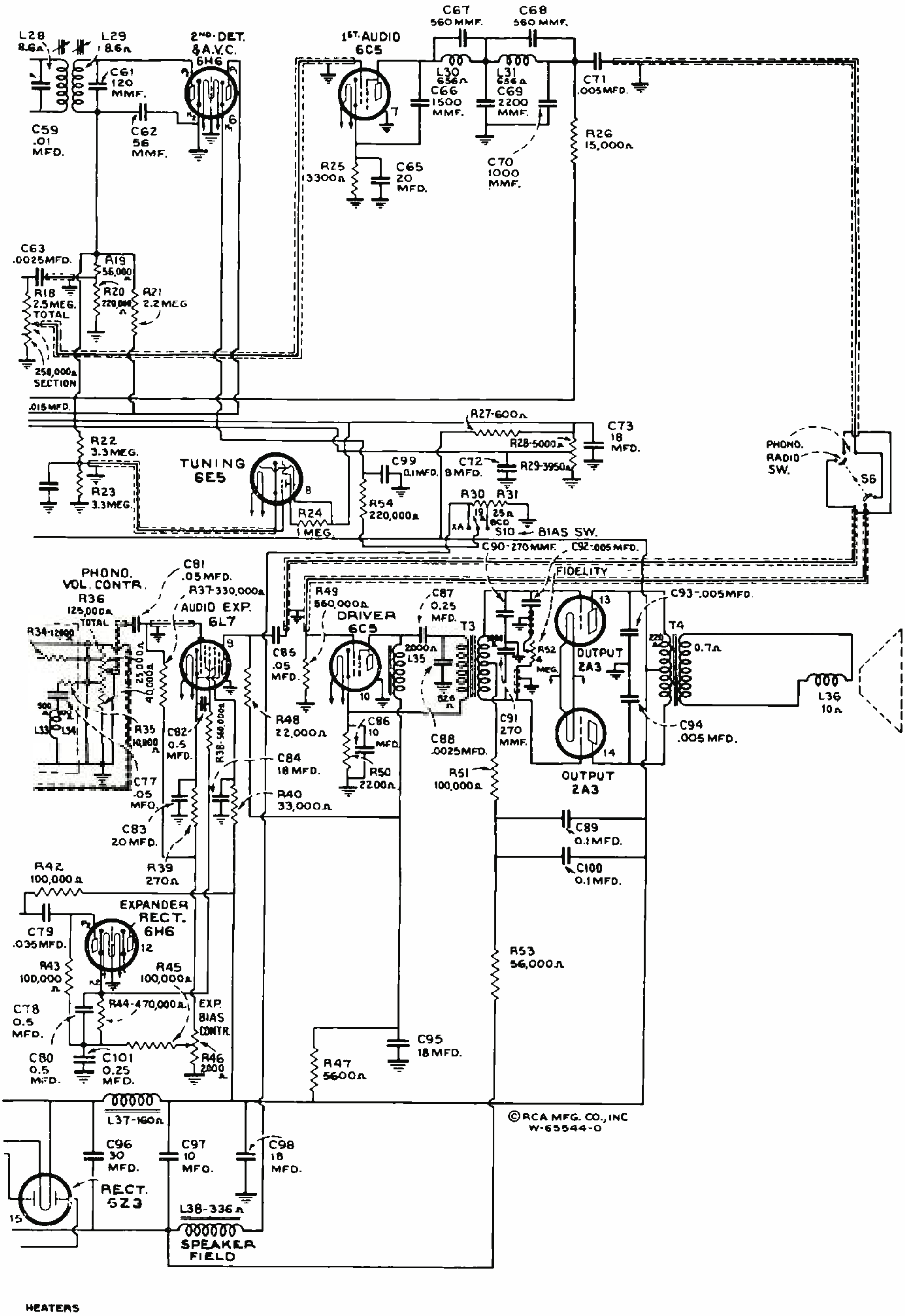


Fig. 2—Schematic circuit diagram of a modern 15-tube all-wave receiver and phonograph combination.



C67
560 MMF.
C68
560 MMF.
L28
8.6Ω
L29
8.6Ω
2ND. DET.
& A.V.C.
6H6
1ST. AUDIO
6C5
C61
120
MMF.
C59
.01
MFD.
C62
56
MMF.
L30
656Ω
C66
1500
MMF.
L31
656Ω
C69
2200
MMF.
C71
.005 MFD.
R26
15,000Ω
R25
3300Ω
C65
20
MFD.
C70
1000
MMF.
C63
.0025 MFD.
R19
56,000Ω
R20
220,000Ω
R21
2.2 MEG
R18
2.5 MEG.
TOTAL
250,000Ω
SECTION
.015 MFD.
R27-600Ω
C73
18
MFD.
R28-5000Ω
R22
3.3 MEG.
TUNING
6E5
C99
0.1 MFD.
C72
8 MFD.
R29-3950Ω
PHONO.
RADIO
SW.
S6
R23
3.3 MEG.
R24
1 MEG.
R54
220,000Ω
R30
R31
25Ω
S10
BIAS SW.
C90-270 MMF.
C92-.005 MFD.
FIDELITY
PHONO.
VOL. CONTR.
R36
125,000Ω
TOTAL
C81
.05 MFD.
R37-330,000Ω
AUDIO EXP.
6L7
R49
560,000Ω
C87
0.25
MFD.
DRIVER
6C5
T3
C93-.005 MFD.
T4
R34-12000
25,000Ω
40,000Ω
C85
.05
MFD.
L35
2000Ω
R52
4
MEG.
OUTPUT
2A3
0.7Ω
L36
10Ω
L33
L34
R35
14,000
Ω
C82
0.5
MFD.
R38-560,000Ω
R48
22,000Ω
C86
10
MFD.
C88
.0025 MFD.
C91
270
MMF.
C94
.005 MFD.
C84
18 MFD.
R50
2200Ω
R51
100,000Ω
OUTPUT
2A3
C77
.05
MFD.
R40
33,000Ω
C83
20 MFD.
C89
0.1 MFD.
R39
270Ω
C100
0.1 MFD.
R42
100,000Ω
EXPANDER
RECT.
6H6
C79
.035 MFD.
R43
100,000
Ω
R45
100,000Ω
R53
56,000Ω
C78
0.5
MFD.
R44-470,000Ω
EXP.
BIAS
CONTR.
C95
18 MFD.
C80
0.5
MFD.
C101
0.25
MFD.
R46
2000
Ω
R47
5600Ω
L37-160Ω
©RCA MFG. CO., INC
W-65544-0
C96
30
MFD.
C97
10
MFD.
C98
18
MFD.
RECT.
5Z3
L38-336Ω
SPEAKER
FIELD
HEATERS

8th—Volumes of receiver circuit diagrams of receivers made by leading manufacturers and the necessary supply of special tools for the alignment and adjustment of various circuits and parts.

The 5th, 6th and 7th items are essential if he is to determine the operating characteristics of the r-f and i-f circuits and the characteristics of the audio amplifying circuits, even after all parts have been checked for their values. Since in the majority of instances the original specifications for selectivity, sensitivity, fidelity, etc. are not available to him, he must have a broad enough technical background to be able with the equipment and information at hand to determine how best to adjust circuit components in order to leave the unit with operating characteristics at least the equivalent of what they were when they left the factory. Should the receiver be several years old, he should be able to adjust its circuits so as to enable it to take advantage of changed and improved broadcasting and transmission characteristics. Only by so doing can he retain the confidence and continued favor of his customers and prevent possible damage to the reputation of the manufacturer of the product and himself.

Let us consider the matter of the adjustment of i-f transformers and main tuning condenser circuits. Remembering that the majority of service technicians operate as individuals, it is unlikely that he has encountered the same receiver previously. He therefore does not know whether the i-f transformers are overcoupled or not, nor to what band width they will or should adjust when peaked. In order that he arrive at a decision as to where to leave the adjustments to give the receiver optimum operating characteristics, he must not only know his theory of transformers and filter circuits, but must also have determined the possibilities of the a-f system and the local receiving conditions under which the receiver is to operate. He will then be able to decide whether the loss of sensitivity, which usually accompanies the broadening of the selectivity curve in loosely coupled windings will reduce the receivable signals in that locality below an acceptable number.

On the other hand, he must know enough about the theory of operation of his test equipment to be able to determine whether the selectivity curve as seen on his cathode ray tube screen actually represents the over-all selectivity of his i-f amplifier or, whether due to some fault in design or change in the test equipment circuit components, the selectivity, as shown, is actually greater or less, by an appreciable amount, than it ought to be. He must also be able to decide whether this change is beyond acceptable limits and what practical effect it will have on the operation of receivers aligned under these conditions.

In those receivers in which the fidelity control varies coupling between primaries and secondaries of i-f transformers, he must be able to determine whether it is best to make his alignment adjustments at minimum or maximum coupling or at some intermediate point in order that the over-all characteristics and quality be acceptable at all positions of the fidelity control. When making the above adjustments, he must be able to determine to what extent the automatic volume control circuit affects the shape and width of the selectivity curve obtained, in order that he be able to decide whether or not the automatic volume control action need be interrupted during the alignment procedure.

In the volume-expansion circuit of the receiver in Figure 2 the failure or normal wear and change of tubes and circuit components will have an appreciable effect upon the amount of expansion taking place and upon the quality of the resulting signal at the speaker. The technician must know his tube, amplifier, and acoustic theory well enough to be able to determine what constitutes too much, or not enough, expansion. Also, he must be able to determine by measurement how much expansion there is, and how much distortion, if any, is introduced in the process, where it occurs and what circuit changes could be responsible for it.

From the preceding illustrations, it should be evident that the technical requirements of the modern radio service technician (and there are other equally important requirements which it is not the object of this paper to discuss) demand that he have a thorough grounding in mathematics through trigonometry and complex numbers, electricity and magnetism, a-c and d-c circuit theory, filters, vacuum tubes and their circuit applications at audio and radio frequencies, r-f wave propagation, acoustics, as well as a practical knowledge of measurement methods and procedure.

Equipped with a background as outlined, the technician is in a position to keep abreast of advancing progress in the industry. This he can do by reading engineering papers on new developments applying to his work, and in this manner keep his knowledge of new circuits and devices considerably in advance of the actual adoption of those circuits in current receivers. Thus, by the time he is called upon to render service on such a circuit, he should be familiar with its operational theory and equipped to render immediate and efficient service on it. This will result in good-will toward both the serviceman and the manufacturer of the product serviced. This good-will invariably results in better business and increased profits for any industry and those engaged in it.

the advent within the past five or six years of phonographs driven by synchronous motors. In an art such as sound recording and reproduction, in which so many factors must be correct to produce a satisfactory overall result, it is not usual to find great refinements introduced in the effort to render one factor perfect so long as there are serious faults of other kinds. For this reason driving systems that today we might regard as short of satisfactory were in general considered good enough, until some of the recent improvements made the shortcomings in the matter of speed more evident. Examples of such advances are electric recording on wax,[2,3] the "orthophonic" phonograph,[2] electric reproduction from disk,[4] improvements in amplifiers and loud speakers,[5,6,7] the development of record materials having lower surface noise,[8] the working out of satisfactory recording on film, and the extension of the frequency range[9] through recent advances in technic. Nevertheless, there were evidently engineers and inventors who, considerably before these improvements came into being, were conscious of shortcomings in the driving systems, and introduced the idea of filtering out irregularities in speed. A French patent No. 10,377, issued to the Fabrica Italiana, Pellicole Parlate, dated June 19, 1909, shows a record turntable driven through a spring, the spring being wound with tape in order to damp out oscillations.

John Constable in a 1918 patent[10] showed a cylindrical record driven through springs, and gives an excellent description of the function of the springs in taking up the vibrations imparted by the driving system. It is possible that these inventors were stimulated to employ filtering systems because they did not have as good governors and gears as could be produced, for their devices did not come into general use until a much later date, and then in greatly modified form. Constant speed in recording or reproduction means not only uniformity of rotation of the record, but avoidance of other relative movements of the record and pick-up device. Thus, the effect of vibration of the pick-up, or record, or both, is the same as if there were changes in rotational speed of corresponding frequency and amplitude. E. H. Amet in a 1917 patent[11] relating to sound motion pictures, shows a very well thought-out system for preventing the transmission of vibrations from his picture projector to his phonograph.

Application of Filters to Turntables

With the need of lower-speed turntables for sound motion pictures, the difficulties of securing constant speed were greatly multiplied. The difficulties were further increased by the necessity of synchronous operation, which meant geared drive.

Although it is possible to produce gears having a very high order of precision, it is practically impossible to drive any mechanism through gears without the introduction of some vibration of the tooth frequency. As soon as attempts are made to build geared machines in any reasonable numbers and with reasonable manufacturing tolerances, it has almost invariably turned out that complete dependence could not be placed upon the perfect functioning of gears. The errors may be small, but they are sufficient to impair sound quality. It is comparatively easy to eliminate the tooth vibrations by driving the turntable, which has considerable moment of inertia, through a flexible element. Such an arrangement is called a mechanical "filter."

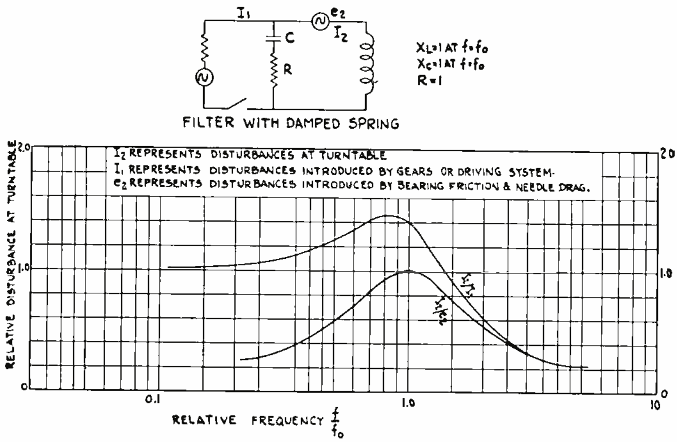


Fig. 2—Characteristics of commercial type of filter.

Figure 2 shows the characteristics of a simple filter comprising a single mass and a single elastic element, for example, a turntable driven through a spring. The ordinates show the ratio of amplitude of swing or speed variation of the turntable to that of the driving gear at the other end of the spring. For disturbances of a frequency at which this ratio is greater than unity, the filter does more harm than good. It is evident from Figure 2, that damping must be provided, since otherwise a bad resonance occurs and persistent oscillations are encountered. Damping may be applied to the spring, or to the turntable. Both systems have been employed, but for the most part the damping has been applied to the spring.

It will be noted that the filter is of no use except when the frequency of the disturbance is above a certain rate. The successful employment of a filter therefore depends upon avoiding, so far as possible, disturbances of slow periodicity, and then designing the filter so that it will be effective at the lowest frequency that must be expected; which means making the spring flexible enough and the turntable heavy enough. This is often spoken of as giving the filter a "low cut-off frequency." Too often the design has been guessed rather than calculated. H. C. Harrison in U. S. Patent No. 1,847,181 gives a very complete formulation of the requirements.

The lowest disturbing frequency is generally the rotation rate of the slowest gear of the train, and with motor-driven machines this can be the turntable speed. When the turntable speed itself is as low as 33⅓ rpm. (this low speed having been adopted to provide long playing records for sound picture work) filtering presents serious problems.

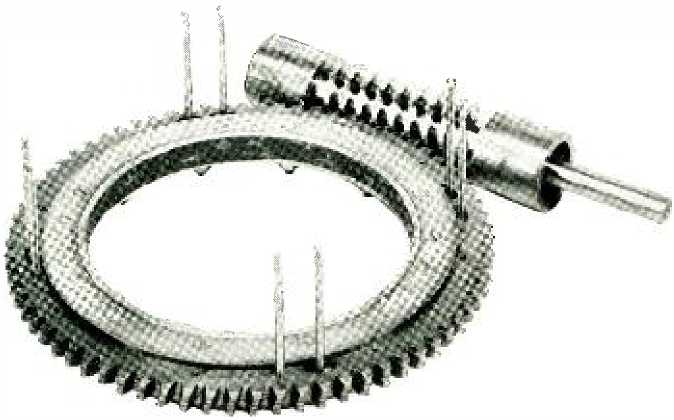


Fig. 3—Laminated worm-gear of Elmer and Blattner.

An interesting arrangement for reducing the magnitude of the disturbances to be filtered out, and at the same time increasing the frequency of their recurrence, was described by Elmer and Blattner in 1929.[12] It is assumed in considering the principle of this driving system, that practically all the disturbance of fundamental or turntable frequency will be due to imperfections in indexing the gear on the turntable spindle. Such indexing imperfections are primarily traceable to the master gear in the miller upon which the gear is cut, and there is no way of averaging out these imperfections in the cutting operation. The imperfections in the gear as cut, however, can be averaged by the scheme described by Elmer and Blattner. The gear consists of four laminae which are bolted together and milled and hobbed in the usual manner. The four laminae are then separated and each rotated 90 degrees with reference to the adjacent ones. They are mounted in such a way that slight play between layers can occur without excessive friction. This compound gear is driven from a single worm. Figure 3 shows the compound gear with worm. It will be appreciated that if a certain imperfection in the indexing tends to cause one layer to rotate above average speed, the same effect will not be produced upon the layer that is reversed with respect to the first layer until the gear as a whole has rotated 180 degrees. Thus, although the speed of the individual layers is no more nearly constant than that of a single gear, the average speed of the four layers will be much more nearly constant, and such irregularities as occur in the average speed will repeat themselves four times per revolution.

The system for averaging the speeds of the four laminae is analogous to the action of a whiffletree which provides an equalization of the motions of several horses, as illustrated in Figure 4. On each lamina are mounted two vertical pins which project through slots in

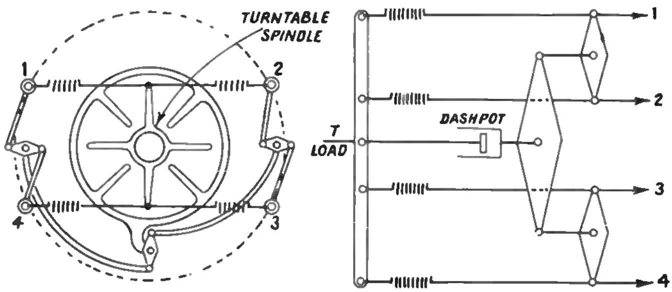


Fig. 4—Diagrams illustrating averaging linkage described by Elmer and Blattner, and straight-line analogue. In actual structure, each gear lamina carried two pins, and equalizing links were in duplicate.

the laminae above, thus providing eight points of attachment for the linkages which equalize the movement. Figure 4 shows the equalizing linkages in schematic form. The turntable is driven through springs, while an oil dash-pot having radial vanes provides the damping. The dash-pot is driven from the linkage. Thus, filtering is employed in addition to the speed-averaging device just described, which reduces the sources of disturbance that must be filtered out.

Bedford Planetary Drive

A radically different method of increasing the frequency of repetition of such disturbances as may be due to faulty gears has been employed by A. V. Bedford[13] and is illustrated in Figure 5. On the turntable spindle is a gear E, and below this is a stationary gear I, of slightly smaller diameter. Two pinions F and U, which are coupled together and rotated by a planetary arrangement, engage gears E and I. Were these gears of the same diameter, no rotation of E would result; but since E differs in diameter from I, each revolution of the pinion-carriage G results in a slight rotation of the gear E. The diameters are so chosen that the pinions must make about seven revolutions to produce one revolution of the turntable spindle. In the course of this one revolution, all parts of gear E have been engaged six times and all parts of gear I seven times. In order further to reduce the possibilities of imparting low-frequency disturbances to the turntable, its spindle rotates not in a stationary bearing but in a bearing that rotates with the planetary pinion carriage G.

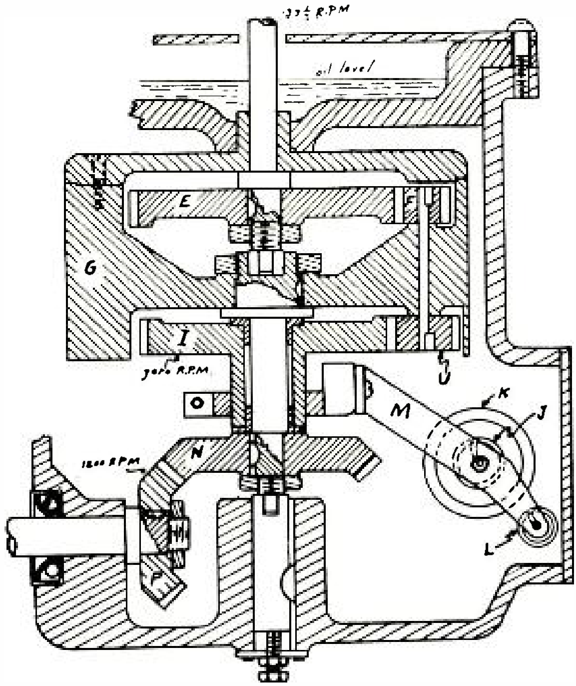


Fig. 5—Planetary turntable drive of A. V. Bedford.

Any torque imparted to the gear E reacts upon the gear I. Were the latter not restrained it would rotate instead of gear E. Instead of providing a rigid anchorage for the stationary gear I, it is restrained by means of a spring L, and damping is provided by the dash-pot K. If the turntable is designed with a reasonable moment of inertia, irregularities in the speed of gear E with reference to gear I, such as may result from gear imperfections, tend to produce movements of gear I rather than speed fluctuations of the turntable gear E. Since the moment of inertia associated with gear I is quite small, I takes up the irregularities, thus providing the filtering upon which reliance is placed to eliminate the comparatively high-frequency disturbances that the planetary drive may introduce.

Brute-Force Methods

It may seem paradoxical that such elaborate arrangements and complicated constructions as are here described should be necessary in order to provide constant motion when, according to the laws formulated by Isaac Newton, all that is necessary to cause an object to move with uniform velocity is to *let it alone.* Fortunately for disk record work, the power requirements are extremely low, and only such considerations as quick starting, saving of space or weight, or the necessity of exact synchronism, make it necessary to resort to any other expedients than the closest practicable approach to "letting it alone."

The employment of very large moment of inertia is often described as a "brute-force" method, but this approach to the problem is good

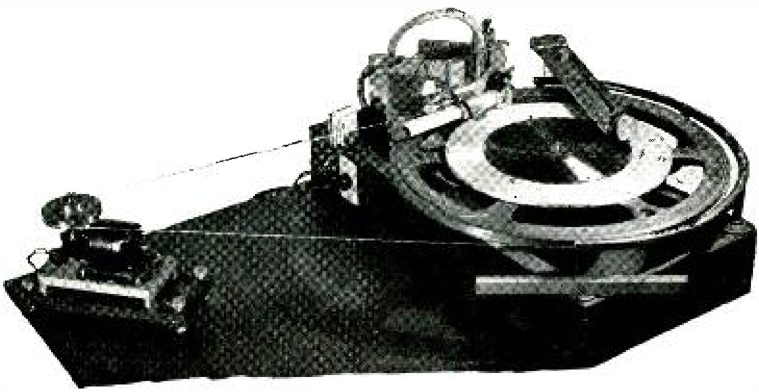


Fig. 6—Turntable used for reference standard.

only when extreme care is exercised at the same time to apply a minimum of forces to the rotating mass. Figure 6 shows the turntable that comes the nearest to constant speed of any with which the author has had acquaintance. The flywheel is 27 inches in diameter, with a heavy rim. The spindle runs in a carefully made sleeve bearing, and the weight is carried by a steel ball at the bottom. Careful bearing design minimizes the disturbing forces from this source, and the supply of the power through a long thread minimizes disturbing forces from the driving system.

Measurement of Speed Fluctuations

The testing of turntables for speed constancy has involved even more serious difficulties than the provision of a turntable having a minimum of speed fluctuations. Stroboscopic methods become increasingly unsatisfactory as the frequency of the disturbance increases. A better system consists in causing the apparatus under test to generate an electric current, the frequency of which depends upon the speed of running. This alternating current is applied to a circuit that is slightly off tune, and the variations in frequency then appear as variations in voltage which can be measured and recorded.[14] For generating the tone it has not been found altogether satisfactory to cut a record and play it back. Better results have been obtained by means of a carefully constructed magnetic tone-wheel. The problems of measurement of speed fluctuations in turntables is discussed by A. R. Morgan and the writer in a paper published in the *Journal of the Acoustical Society of America* for April, 1936. Using the magnetic tone-wheel on the turn-

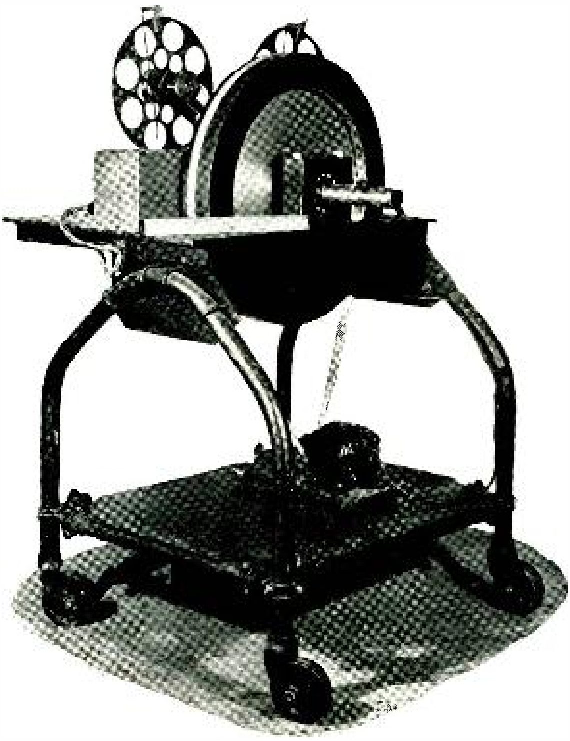


Fig. 7—Example of "brute-force" constant-speed film phonograph.

table shown in Figure 6, we have found it possible to get registrations of as low as 0.1 per cent between the lowest and the highest speeds registered in a period of 6 seconds. The measuring system shows full sensitivity to fluctuations up to 100 per second.

Sound-Film Systems

Photographic recording of sound on film began a good many years ago. The problem of obtaining constant speed does not differ in fundamental principles from that of providing constant speed for disk records, but does present new problems in that the entire record does not move as a unit, but is unwound from a supply reel, pulled past the point where a record is made or reproduced, and again wound upon a reel. The motion at the translation point must be protected from irregular pulls due to the reels. An additional difficulty is generally brought in by the requirement that a specified number of sprocket holes must pass per second, regardless of the absolute length of film represented by this number of sprocket holes. In other words, the linear speed for a shrunken film must be less than for a non-shrunken film. Apparent lack of concern for speed constancy upon the part of earlier experimenters in photographic sound recording is

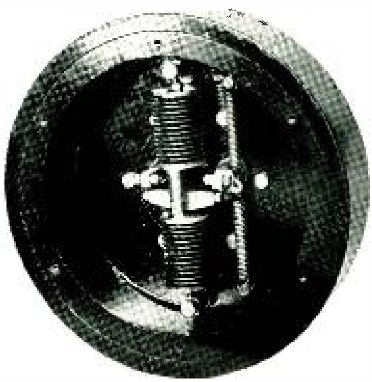


Fig. 8—Sylphon damping arrangement
L. A. Elmer.

probably due to the fact that they had too many worries about other things. That the speed should be constant, however, has always been accepted as an axiom. Machines employing the "brute-force" method of moving a film at constant speed have been of extreme value as laboratory tools. Figure 7 shows such a machine, known in the RCA Photophone laboratories as the "Grindstone." Such extremes of size and mass are hardly necessary. In general, rotational speeds in film machines can be considerably higher than in disk turntables (for example, 180 to 360 rpm.), and at these speeds flywheels of moderate size are sufficient.

Filters on Sprocket Shaft

In most of the earlier commercial reproducing machines, and in many of the recording machines, the film was moved past the optical system by means of a sprocket; and, practically without exception, the sprocket drive has been filtered. In other words, a flywheel was mounted upon the sprocket shaft, and power was supplied through an elastic connection from a driving gear. As has already been explained, a filter may do more harm than good if it is not adequately damped. The damping has in some cases been applied to the flywheel itself in the form of a viscous brake, and in other designs the elastic connection between the driving gear and the flywheel was damped. Figure 8 shows a widely used damping arrangement employed in projector sound heads.[15] The two sylphons, or bellows, which are mounted upon the flywheel are completely filled with oil and connected through an adjustable orifice. The mechanism is so arranged that whenever there is relative movement of the flywheel and driving gear, one bellows is stretched and the other is compressed, and oil must flow between them. A device of this form provides very effective damping, but must be

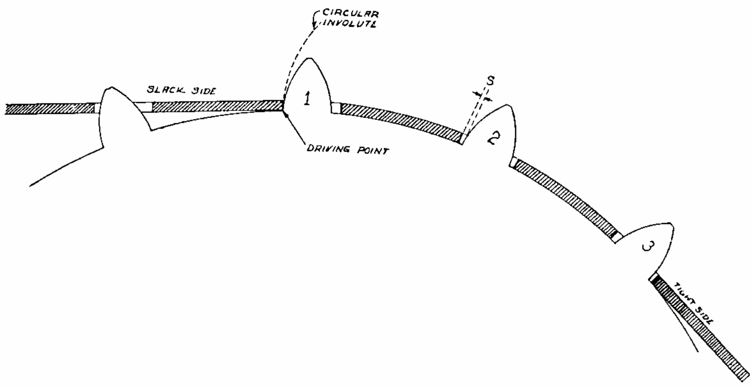


Fig. 9—Common relation between sprocket teeth and perforations.

designed with great care in order that changes in the torque transmitted may not result in appreciable unbalance. All filtered systems with horizontal axes are extremely sensitive to balance.

Inherent Faults in Sprocket Drive

Although the painful effects of slow speed fluctuations, particularly in the reproduction of music, were early observed, and efforts made to eradicate them, general recognition of the fact that rapid fluctuations of small amplitude can also do serious damage to quality has been extremely tardy. This is perhaps due to the fact that such rapid fluctuations produce radically different effects, which were not recognized as being due to speed changes. Moreover, other common faults in recording and reproduction cause impairment of quality so similar to that due to rapid speed variations as to mask the results in comparative tests. The art of sound reproduction may be metaphorically described as "strewn with the wrecks of efforts" to obtain constant speed by means of gears and by means of sprockets. The shortcomings of sprocket drives are made much more serious because the pitch of the film perforations varies with the shrinkage of the film. No matter how the sprocket is designed, there is only one chance in many that the film will really fit the sprocket.[24]

Figure 9 shows the relation between teeth and sprocket holes when the pitch of the perforation is slightly greater than that of the teeth. It is seen that the tooth on the left is doing the pulling and will continue to do so until the film is stripped off, at which time the film must slip back on the sprocket by a distance *S;* whereupon tooth No. 2 does the pulling. Thus, the film moves at sprocket-tooth speed only for short intervals, interspersed with moments of slipping back. If the sprocket-hole pitch is less than the tooth pitch, there must be a forward adjustment as each new tooth engages, instead of a slipping

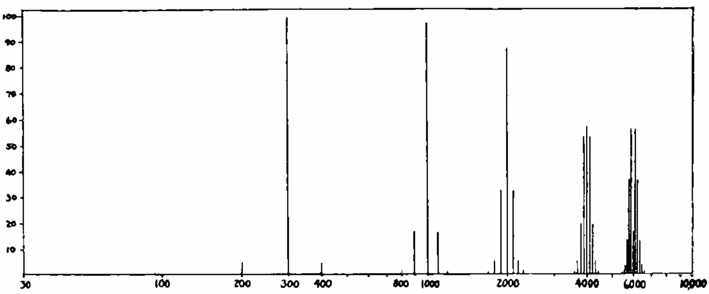


Fig. 10—Effect of film speed modulation upon sine wave recordings of 300, 1000, 2000, 4000, and 6000 cps. (Assumed: speed modulation of 1 mil amplitude at 100 cps.)

back. In either case, there is a decided irregularity in film movement, the effect of which is to cause less satisfactory reproduction of high-frequency tones.

Figure 10 shows what happens to tones of several different frequencies as a result of speed fluctuations at sprocket-hole frequency.[16,17] If a sine wave is recorded and is reproduced by a machine that introduces a variation of 0.001 inch in amplitude approximately 100 times per second, there is little effect upon a 300-cycle tone, although small amounts of 200- and 400-cycle tones are generated. If the recorded frequency is 1000 cps. the magnitude of the 900- and 1100-cycle tones is about 15 per cent of the original. At 2000 cycles the first sideband tones are fully 30 per cent of the original, while additional tones of 1800 and 2200 become appreciable. If the recorded tone is 4000 cps., the sidebands are practically as large as the fundamental, while at 6000 cps. there are four sidebands that exceed the fundamental and two more that are nearly equal to the fundamental. It could hardly be expected that a recorded 6000-cycle note reproduced on such a machine would sound clean and clear, although some semblance of a high-frequency note still results.

Since satisfactory reproduction cannot be expected if the film is propelled past the optical system directly by a sprocket, the best machines have employed smooth drums on which the films were supported, dependence being placed upon friction to insure the film's moving with the drum without slipping. Soft-tired rollers pressing the film against the drum have proved satisfactory and make little trouble if properly designed. Various methods of driving the drum have been employed, and among these methods it is regrettable that there have been many attempts to drive the drum by gearing it to the same driv-

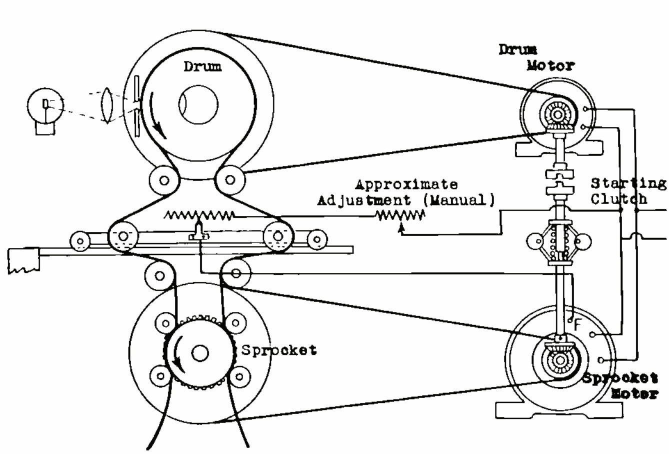


Fig. 11—Electrical drum speed control of C. A. Hoxie.

ing system as the sprocket, without provision for permitting the drum to operate at any other than a fixed speed with reference to the sprocket. Such machines may have the appearance of operating as if the drum were performing a useful function, but as a matter of fact, the film must be slowly slipping with reference to the drum surface throughout the operation, and all benefit from the constant-speed drum is thereby sacrificed. The drum must be permitted to select its own speed, depending upon the film shrinkage.

The simplest way to let the drum select its own speed is to drive it by means of the film, which acts as a belt. In applying this principle, dependence has almost always been placed upon a flywheel to provide for uniform rotation of the drum. It would be hard to imagine a simpler system that would eliminate the disturbance due to sprocket teeth. The sprocket tooth disturbance present at the sprocket pulling the film is filtered out because of a certain amount of flexibility in the loop of film between the sprocket and the drum cooperating with the effective mass at the drum due to the flywheel.

The trouble with the simple drum and flywheel system is that it constitutes an undamped filter such as was discussed in connection with Figure 2. Oscillations are easily set up which persist for a long time. A small amount of damping is, to be sure, provided by the viscosity of the bearing lubricant; and if through careful construction, or good design, or good luck (or, perhaps, by tedious and painful reconstruction), no disturbance of frequency near the natural frequency of the oscillatory system is introduced, a machine of this type may give creditable performance. A number of other methods of driving the drum at the correct speed have been employed, and some of these are of much interest in view of their ingenuity.

Figure 11 shows the arrangement of an early laboratory reproducing machine of Charles A. Hoxie, of the General Electric Co. The drum is driven by a separate d-c. motor, the speed of which is controlled through a fine-adjustment rheostat, the position of which is changed by alterations in the lengths of the loops between the sprocket and the drum.[18] This system is not subject to oscillation, since the motor assumes its new speed almost instantly when the rheostat is changed. It is in systems in which the *acceleration* of the drum (rather than its speed) varies in proportion to film loop length, that oscillations are likely to occur.

C. L. Heisler, also of the General Electric Co., designed a recorder that was widely used commercially, using a mechanical method of introducing the fine control of the drum, instead of an electrical method.[19] Figure 12 shows some of the features of the Heisler machine. The principle is in no wise different from the employment of two oppositely tapered cones, with a belt or friction wheel that can be moved axially to the position at which the speed ratio is correct. Instead, however, of simple conical pulleys, Mr. Heisler employed doubly curved surfaces of revolution, *A* and *C* of Figure 12, the elements of which were circular arcs. This made it possible to rotate the axis of the intermediate friction wheel *B*, instead of shifting the wheel along its axis. The intermediate friction wheel was mounted much like a gyroscope, except that its axis had only one degree of freedom. On the drum side, the running surface is on the interior of a hollow member *C*. It will be observed that as the axis of the intermediate wheel *B* is rotated in a clockwise direction, its rim is brought into contact with a larger diameter portion of the driving cone *A;* and, on the other side, it runs against a part of the driven member *C* where the diameter of the latter is less. This provides a continuous adjustment of speed.

The upper drawing of Figure 12 shows the loop of film between the sprocket of the drum, engaging a movable idler at the bottom. Changes in the position of the axis of the friction wheel are caused by the lengthening and shortening of this loop of film.

Geared Compensator of C. L. Heisler

Figure 13 shows another arrangement by Heisler.[20] A ring-gear, *A*, is mounted on the shaft whose speed is to be controlled, and a slightly

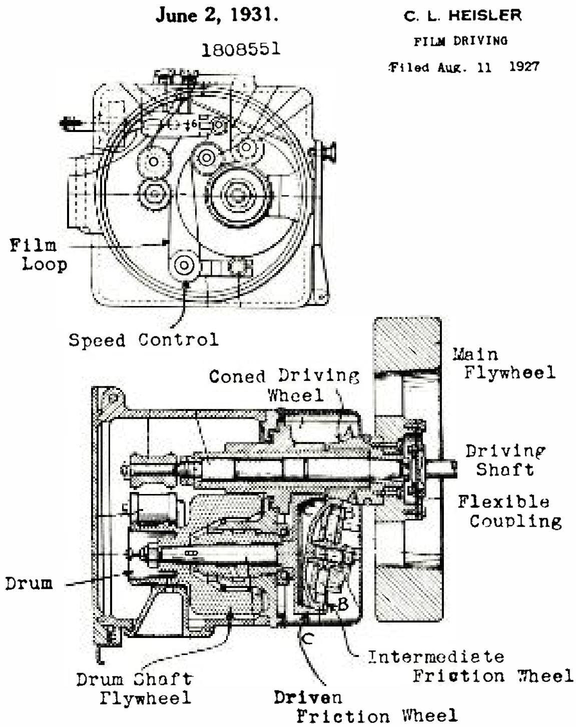
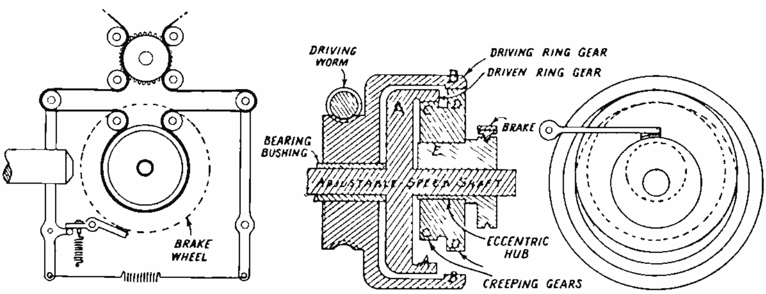


Fig. 12—Mechanical drum speed control or compensator of C. L. Heisler.

larger ring-gear, *B*, is driven at fixed speed from the motor. Two gears *C* and *D*, mounted on a single hub, which the writer has called "creeping gears," mesh with the two ring gears. The creeping gears run on an eccentric hub *E*. When this hub is permitted to rotate with the rest of the assembly, there is no relative motion of the four gears, and the driving and driven ring-gears rotate at the same speed. Application of a brake, however, to the eccentric hub causes the pair of creeping gears to run around inside the internal gears, and in doing so they cause one to shift slowly with respect to the other.

The smaller the difference between the diameters of the two internal gears, the smaller is the creeping produced by applying the brake to the eccentric hub, and the arrangement may be so designed that the entire difference in speed of the driven shaft produced by applying a brake is only 2 or 3 per cent of the mean speed.

The drawing on the left in Figure 13 shows the manner in which the loops of film between the sprocket and the drum control the application of the brake. This drawing shows the arrangement as if the geared compensator were applied to alter the speed of the drum, but in most of the machines that Heisler designed it was not the speed of the drum that was changed, but the speed of the rest of the machine, comprising all the sprockets. This is entirely permissible in a projector, which does not have to be synchronized with other machines. Since the geared compensator is a means of controlling average speed,



SCHEMATIC REPRESENTATION OF GEARED COMPENSATOR OF C.L.HEISLER

Fig. 13—Geared compensator of C. L. Heisler.

but does so by causing small fluctuations, it is logical to drive the drum directly through gears and a filter from the motor, and to employ the compensator to insure that the sprockets will have the correct average speed to maintain proper film loops. In such a machine it is evident that the film speed in linear feet per minute is the fixed quantity, while the number of sprocket holes per minute will vary, depending upon film shrinkage.

In an alternative construction, the creeping of one gear with respect to another was brought about by an arrangement of worms that rotated with the assembly, but did not turn upon their axes unless a brake was applied to a suitably arranged brake-wheel. In practice Heisler found that the action of the brake was such as to cause the brake-wheel to assume an intermediate speed rather than to alternate between full speed and stop. This was due to the facts that the brake-wheel possessed considerable inertia and the entire mechanism operated under conditions of abundant lubrication so that the brake produced a semi-viscous drag. Since the total speed change produced by stopping the brake-wheel amounted to only 2 or 3 per cent of the average, it is evident that even a rough approximation to constant brake-wheel slip would result in a very nearly constant speed of the driven member. Moreover, filtering may be employed between the driven member of the compensator and a drum. This method of operation permits the application of the geared compensator to machines such as recorders, which have to be strictly synchronized.

RUBBER-TIRED COMPENSATOR OF C. A. HOXIE

Hoxie built a sound head in which he provided continuous fine adjustment of the drum speed by altering the pressure between a

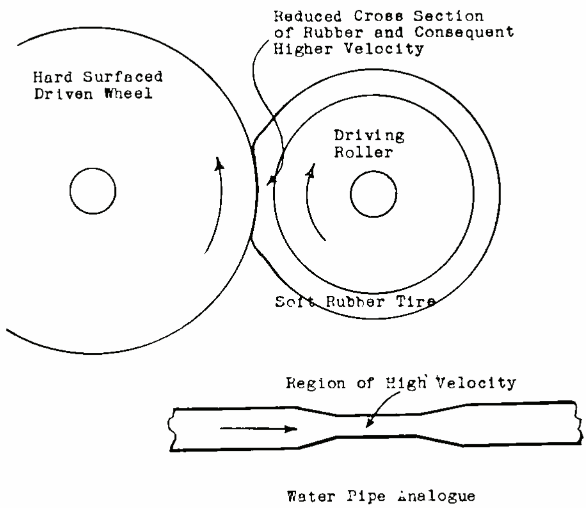


Fig. 14—Speed control by pressure on rubber tire, used by C. A. Hoxie.

rubber-tired roller, driven at constant speed, and a smooth-rimmed metal wheel.[21] I am not sure how Mr. Hoxie expected the device to work when he made the first model, but it may be said that a number of us expected exactly the opposite from the actual result. The working radius of the driving wheel becomes less as the tire is compressed. We therefore inferred that the driven wheel would run slower as the pressure was increased, assuming, of course, that the driven wheel rotated easily and that the pressure was never reduced to a point where appreciable slipping occurred. What actually happened was that the greater the pressure, the faster the driven wheel ran.

The proposition may be illustrated in terms of a truck. We shall assume that the pavement is smooth and level, and that the truck has thick, soft tires of solid rubber, and that the diameter of the rear wheels is such that when they are rotating at 200 rpm. the truck will run 20 miles per hour when empty. The truck is now loaded sufficiently to compress the tires materially. With the same wheel rpm., will the loaded truck run slower or faster than it ran when empty? The answer is that the more you load it the faster it will go. I know of no actual test with a truck, but believe that the analogy is a fair one, and the explanation is quite simple.

Figure 14 illustrates the relation of the rollers and also shows a diagram of a Venturi tube. For such changes of shape as a piece of soft rubber can undergo it acts much like a liquid, in that it permits large shearing deformations with little resistance. It is evident that at the point of maximum compression, the tire has smaller cross-section than at other points, but it is also obvious that the total volume of rubber passing this point in a given interval is the same as that passing other points. Therefore the velocity of the rubber must be greater where the tire is compressed, and the greater the compression, the higher will be the velocity of the rubber, just as the velocity of water at the constricted portion of the Venturi tube is greater than that in other parts of the pipe. Directly under the truck wheels the surface velocity of the rubber with reference to the truck may considerably exceed the velocity of other parts of the tire surface, and this effect much more than offsets the slight reduction in radius of action. If the tire were compressed to ½ its normal cross-section, the velocity of the rubber at this point would need to be doubled, on the average; and since this increase in average velocity will be mostly at the surface and none at the rim, very substantial increases in speed may occur.

Pinch-Roller Compensator of A. V. Bedford

A method of compensation that has surpassed even the rubber-tire compensator in its ability to provide subject-matter for discussion and argument, and likewise resembles the rubber-tire arrangement in its extreme simplicity, is the pinch-roller compensator, first proposed by Bedford for projectors[22] and later applied to printers.[23]

In Figure 15, the upper drawing shows the general mechanical arrangement, while the lower drawing shows the principal elements in simplified form. Roller *4* is driven at fixed speed, while roller *5*, which serves to hold the film in contact with roller *4*, may assume whatever speed the film imparts to it. If we could stretch or compress each film to such a degree that 64 sprocket holes corresponded to exactly one foot, thereby mechanically compensating for shrinkage, it would obviously be possible even in a synchronous machine to propel film by means of a smooth roller having a fixed peripheral speed. Fundamentally it is not necessary to stretch or compress the film throughout its entire cross-section, for the film is driven solely by contact between one surface and roller *4*. It is therefore sufficient to stretch or compress the material at this surface, and it is well known that the simple act of bending a flexible strip serves to compress one surface and stretch the other. It will be readily seen from the drawing that if a sufficient amount of film accumulates in the loop between the sprocket and the driving roller, the film will bend around the

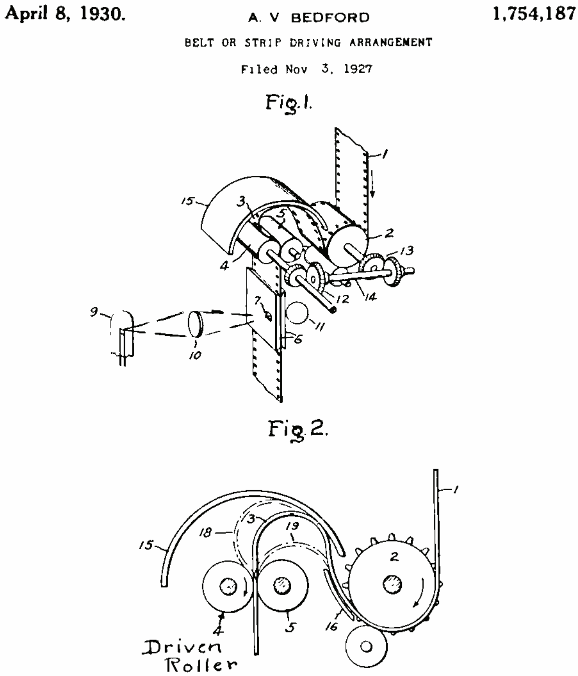


Fig. 15—Film speed control of A. V. Bedford.

roller in such a way that the latter works on the concave side of the bent film. On the other hand, with a short loop, as indicated at *19*, the driving roller will operate on the convex side of the bent film. When the driving is applied to the convex side, the mean film speed is less than that of the driven surface. In other words, when the loop is short the film does not pass so rapidly through the driving point between rollers *4* and *5*. Whatever the degree of shrinkage of the film, an equilibrium is soon reached with the loop length such that the rate of passage between the rollers is exactly equal to the rate at which it is fed through by the sprocket. Engineers dealing with problems of belt drive are well acquainted with the necessity of making a distinction between the pulley surface speed and mean belt speed, or the speed of the middle of the belt, and the Bedford drive depends upon exactly the same principle.

(To be continued.)

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GRAPHICS OF NON-LINEAR CIRCUITS

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PART II

(Continued from July issue)

VI. APPLICATION TO TRIODE.

The above constructions can be applied to a triode, (or multi-grid tube if all but one grid are at constant potentials), with the only restriction of course, that the connected circuits can pass direct current. In the case of a triode we note that the impressed voltage e equals E_B, the generated plate supply voltage, and is therefore constant. The instantaneous resistance of the tube, however, is a function of the grid voltage. If the latter is a known function of time, then the approximate value of the instantaneous resistance is known during any time interval $\triangle t$, and the graphical construction is therefore possible. To illustrate the application, let the triode characteris-

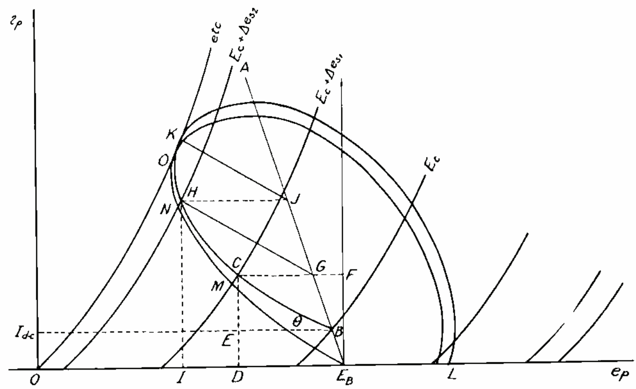

Fig. 9.

tics be those shown in Figure 9. Suppose an inductance L, and resistance R are in series with the plate and the "B" voltage E_B. There are really two transients: (1) when the plate supply switch is first closed (establishment of initial d-c component), and (2) when signal voltage $e_s = f(t)$ is first impressed in the grid circuit in addition to the normal bias voltage, e_c. Ordinarily we are not interested in the transient set up in (1), hence we determine the initial d-c component by merely drawing the load line for the resistance R of the load, in the well-known manner. The load line for condition (2) is the one we now wish to determine. We proceed as follows:

From the plate voltage E_B we draw the load line for R as shown in the figure, line AE_B. Its intersection B with the bias voltage curve. E_c, determines the initial d-c component.

The instantaneous signal voltage e_s is now determined for each time interval $\triangle t$, and the corresponding curve of the plate family used. As shown in Figure 9, e_s is assumed to be sinusoidal, and to start from its zero value. Only these curves of the plate family have been plotted, as shown, and in the figure large time intervals $\triangle t$ have been taken in order to make the figure clearer. If the grid voltage in the first time interval $\triangle t$ changes from E_c to $(E_c + \triangle e_{s1})$, the current rises along BC, where

$$\theta = cot^{-1}(L/\triangle t + R) \tag{28}$$

and CE represents $\triangle i_1$, and CD the total current i at that instant. The point C is projected over to the E_B ordinate as F, and the latter is really the point on the overall load line, all points of which will be along this E_B ordinate. However, in the case of the triode we prefer to call such points as C, H, K, etc., the overall load line, in which case our meaning is a plot of functional relation of plate current versus grid

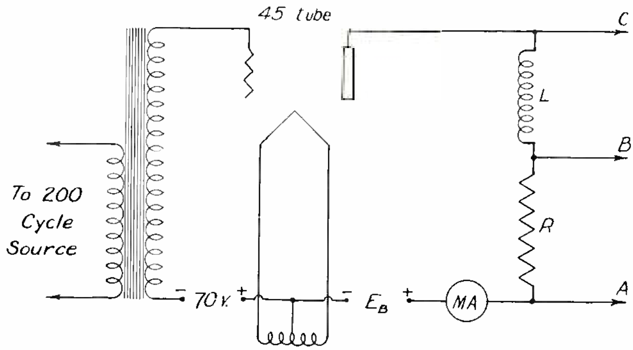

Fig. 10.

voltage, rather than actual energy-supplying voltage, E_B, impressed in the circuit. From the bottom of F, or point E_B, a line is drawn at the angle whose cotangent is R; in other words E_BA. This line intersects the current abscissa through C and F in G. Through G we draw GH parallel to BC and HI is the next instantaneous value of current.

In the case of a triode, the procedure just outlined is unnecessary because the impressed voltage E_B is constant, whereas if E were variable with time every step would be necessary. Here we merely need to project the points over to the load line for R, and then draw the finite operator curve through the projected point over to the next value of the grid parameter. Thus H is projected over to J, and JK drawn parallel to CB. Then K is the next point of the load line, and the rest are found in the same manner.

It will be noted from Figure 9 that if cut-off of the plate current occurs during the negative portion of the grid swing, at point L, the curve will continue to the left along the e_p-axis to the point E_B, and rise from there for every cycle of grid swing thereafter. Hence, if the steady-state solution is desired, and it is foreseen that cut-off will occur, the construction can be started at E_B, and the initial loop *BCHK* ignored. This is a fortunate reduction in the amount of work necessary in a very non-linear case: that of operation beyond cut-off.

In analyzing the theory of this application, we may regard the triode as a resistance which varies both with current and time in a determinable manner, while the impressed voltage remains constant at the value E_B. On the other hand, we may regard the construction as the discrete projections on the e_p-i_p plane of the intersection of a shifting finite operator surface with the tube surface.

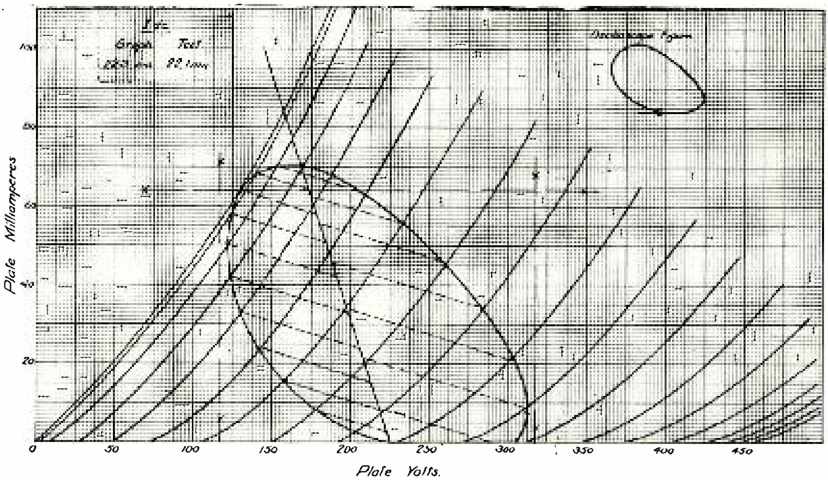

Fig. 11—Inductive load line. Type 45 Triode. $E_B = 225$ volts.

VII. Experimental Verification

Experimental set-ups were made to check some of the circuits shown. One such is shown in Figure 10. A Type 45 tube was measured for its plate family characteristics. Thus calibrated, it was connected up to an inductance L, of one henry, and a resistance R as shown. The total resistance of the load circuit was adjusted to 800 ohms. The grid was connected to a beat frequency oscillator set to 200 cycles, and the grid signal voltage, of sinusoidal wave shape, adjusted so that its peak value was 70 volts, and therefore equal to the bias. In this way as large a grid swing as possible was obtained without the grid being driven positive and thus contributing distortion products to the plate-current wave-shape. The plate voltage E_B was adjusted to various values as shown on the accompanying graphical diagrams, and wires A and B were connected to the vertical deflection plates of a cathode ray oscilloscope, and A and C to the horizontal deflection plates. In this way a figure was obtained on the screen which would correspond to the graphical construction for the operating conditions.

The results are shown in Figures 11, 12, and 13. The small figure on each graph represents a copy of the oscilloscope trace. It will be noted that qualitatively the two correspond, although the trace shows a transient which the author feels originated in the oscilloscope. Quantitative checks were not possible, and investigation showed that the small figure of thick line on the oscilloscope could hardly serve for accurate comparison. Moreover, it was found that the deflection plates were not quite at right angles to one another. Hence a crest voltmeter was used to measure minimum and maximum plate voltages. These

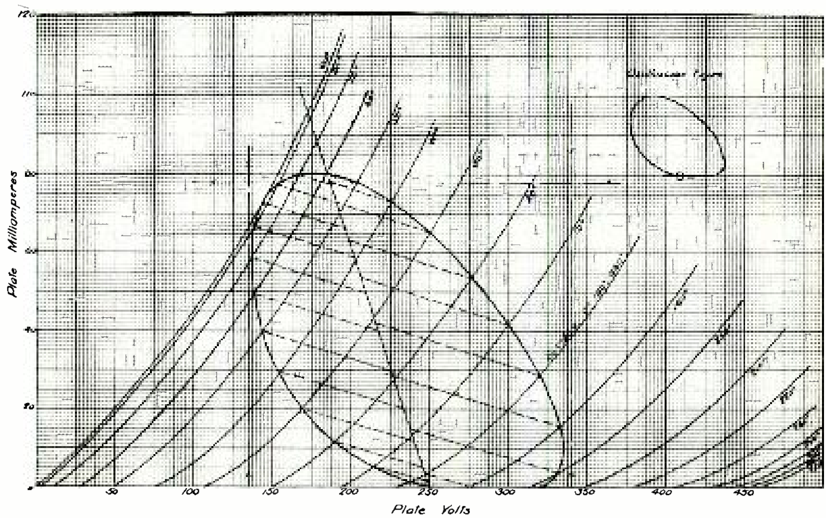

Fig. 12—Inductive load line. Type 45 Triode. $E_B = 250$ volts.
I dc. Graph 26.8 ma. Test 26.5 ma.

checked fairly well with the graphical construction, and the values are shown on the diagram by crosses.

A brief summary of the graphical method will be given. The procedure is as in Article IV, where $C = 0$ or $R_2 = \infty$, with the difference that a triode is used here. The grid voltage wave is

$$e_g = 70 \sin 400 \pi t \qquad (29)$$

This wave is broken up into 10° intervals, so that $\triangle t = 1/7200$ sec. Then $L/\triangle t = 7200$, and $(R + L/\triangle t) = 8000$. Taking account of the scales, the rise is ten divisions for every thirty-two divisions horizontally for $(R + L/\triangle t)$, and 50 divisions up for 16 divisions horizontally for R $(= 800$ ohms). In these examples, the current cuts off during a portion of the cycle, hence steady-state overall load lines can immediately be drawn by starting the construction at $i_p = 0$ and $e_p = E_B$, as shown.

In comparing the results, it was found that the maximum deviation between the construction and the peak voltmeter readings was approximately four per cent for plate voltage. A study of the diagrams will show, however, that the readings are probably in error, particularly the peak current readings. The load line must start on the plate voltage axis at E_B, and be tangent to the zero grid voltage curve. If it is attempted to draw a load line through these two points and the peak current and voltages experimentally determined, it will be found that the load line does not at all resemble the figure of the oscilloscope, which at least qualitatively checks the graphical construction.

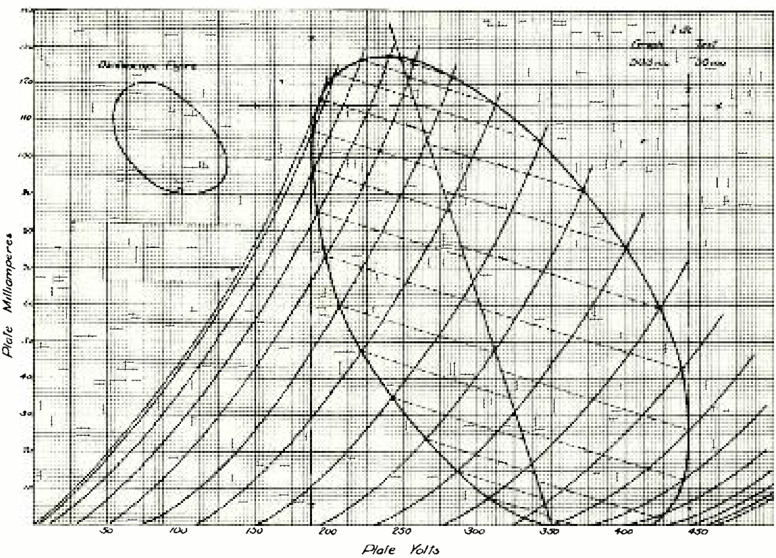

Fig. 13—Inductive load line. Type 45 Triode. $E_B = 350$ volts.

Another check for the latter is as follows: If the load line be replotted as a plate current-time wave, and analyzed for its d-c component, it will be found that it checks the d-c plate milliammeter very closely. It is felt that this check, together wih the fairly close maximum and minimum plate voltage peak voltmeter readings, constitute sufficient proof of the correctness of the method and the construction. Of course some error is to be expected in the latter since finite, rather than infinitesimal time intervals are used.

It was felt that a check with a harmonic analyzer would not be conclusive, since the shape of the load line depends not only upon the amplitudes of the various harmonics, but their phase relative to the fundamental, a quantity not measured by the analyzer.

The second example is that of a dynatron circuit. The construction of Article IV is to be used. The circuit and value of parameters employed is shown in Figure 14. The inductance had an ohmic resistance of 500 ohms, so that the total resistance of that branch was 800 ohms. The screen voltage was maintained constant at 130 volts, and the plate supply voltage at each of three values: 45 volts, 22.5 volts, and 13.5 volts.

In utilizing the construction of Article IV, we note two simplifying features: (1) no appreciable resistance in the capacitive branch, and (2) the applied voltage is constant at one or other of the three values mentioned above. Hence a form of construction as shown in Figure 15 can be employed. The dynatron characteristic is represented by r, and the plate supply voltage by E_o. Assume that this can be applied in such manner that no oscillations start, so that a steady current BM flows as determined in the well-known manner by R. This

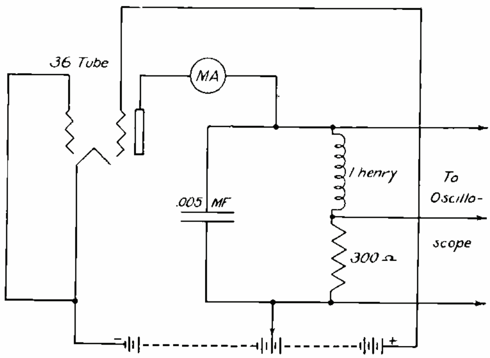

Fig. 14.

is our initial condition. Then BA represents the voltage drop in R due to BM. There are two equal voltages present in the parallel branches of the circuit, each of value $AB = (BM)R$. By the methods outlined in this paper, they may be replaced by two equivalent voltages appearing in series with E_o, namely $(AB)\dfrac{R}{Z_L}Z_t$ and $(AB)\dfrac{Z_t}{Z_L}$. The former represents the series voltage drop equivalent to that actually across R, and the latter the series voltage drop equivalent to the actual voltage across C due to its stored charge. If these two equivalent voltage drops be added algebraically to E_o, we obtain $E_o - \left[(AB)\left(\dfrac{Z_t}{Z_C} + \dfrac{RZ_t}{Z_L}\right)\right] = E_o - AB$. Thus, if we attempt to proceed with a graphical construction from our initial condition by proceeding as in Article IV, and draw the finite operator curve for Z_t from $(E_o - AB)$ to r, we find that Z_t starts from point B on r, and thus our construction does not move from its initial position.

B. This means that the circuit is in equilibrium, and must receive a shock impulse in order to start oscillating. Suppose this is accomplished by imparting an additional charge on C and thus changing the voltage across the latter by an amount E_C. The equivalent voltage in series with E_o is $E_C \dfrac{Z_t}{Z_C}$, and is represented (Figure 15) by BC. We can now proceed. From C, CD is drawn to represent Z_t. Then DG represents $(i_C + \triangle i_L)$, and GA represents the voltage E across the parallel resonant tank circuit. From A, AF is drawn to represent Z_L. Then FG is $\triangle i_L$, and FD is i_C, and it is noted that due to the negative

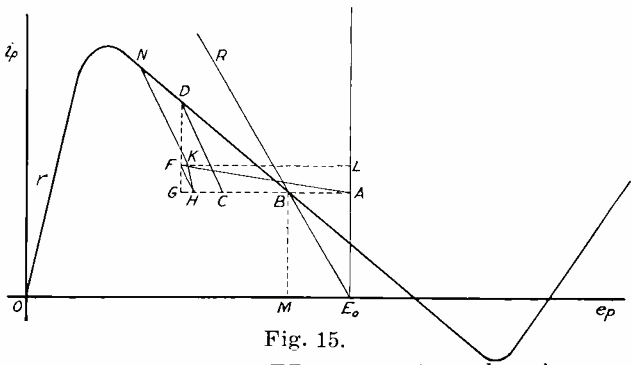

Fig. 15.

resistance characteristic of r, FD represents a charging current into the condenser C. Two equivalent series voltages must now be added algebraically to C, namely $(\triangle i_L)\left(\dfrac{RZ_t}{Z_L}\right)$ and $(i_C Z_t)$. This can be done by drawing FH parallel to DC, and then HK to represent the finite operator curve $\left(\dfrac{RZ_t}{Z_L}\right)$. Then K is the next point from which Z_t can be drawn to r to get the next values of i_C and $\triangle i_L$.

The construction can be further simplified in two ways:

(1) From F a finite operator curve of value $Z_t\left(1 - \dfrac{R}{Z_L}\right)$ can be drawn to GA, and the intersection projected up to FL to obtain point K. This replaces two finite operator curves FH and HK with one. Thus for each point three finite operator curves, Z_t, Z_L and $Z_t\left(1 - \dfrac{R}{Z_L}\right)$ are required.

(2) In the examples to be presented, Z_L comes out of such high value that its operator curve is nearly parallel to the voltage axis. Hence $\triangle i_L$ (or FG) and FK were computed directly upon a slide rule to obtain greater accuracy. Thus, after CD is drawn, voltage GA can

be read off from the graph. Then this voltage can be divided by Z_L and the quotient $\triangle i_L$ laid off graphically as GF. Furthermore, $\triangle i_L$ is multiplied by $Z_t\left(1 - \dfrac{R}{Z_L}\right)$, and the product laid off graphically as FK. Then from K a line KN, parallel to CD and thus representing Z_t, is again drawn to r, and the process repeated. The construction thus becomes one graphical and two slide-rule manipulations for each point, and represents a welcome saving in construction lines in the solution. We thus use the graphical construction where

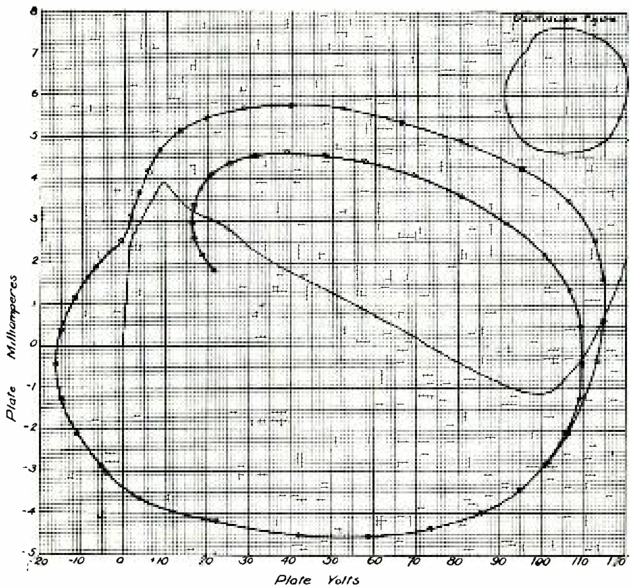

Fig. 16—Dynatron load line. Type 36 tube. Plate voltage = 45 volts.

it is indispensable: to find $(i_C + \triangle i_L)$, or the intersection of the finite operator curve Z_t with the irregular curve representing the load line or characteristic for r.

The characteristic to be plotted here is the current in the inductive branch versus the tank voltage E. Point F in Figure 15 represents one such point.

In the example cited in Figure 14, the finite operator values were calculated for a $\triangle t$ equal to 1/72,000 second. Then $Z_L = \dfrac{1}{1/72000} + 800 = 72800$; $Z_C = \dfrac{1/72000}{5 \times 10^{-9}} = 2777$; $Z_t = \dfrac{72800 \times 2777}{72800 + 2777} = 2675$; and $Z_t\left(1 - \dfrac{R}{Z_L}\right) = 2675\left(\dfrac{800}{72800} - 1\right) = 2650$. These values are then divided by 1,000 to obtain the proper cotangents when the current scale is in milliamperes.

The characteristic for a plate voltage of 45 volts is shown in Figure 16. It will be noted that there is a close similarity between the graphical construction and the figure obtained on the oscilloscope. In the former, the initial charge assumed was insufficient to give a closed loop in one cycle, but it was obtained in the second excursion around the dynatron characteristic. The number of points in the closed loop indicate the frequency in that they are 1/72000 of a second apart, and total to one period. Thus there are 35 points, and the corresponding

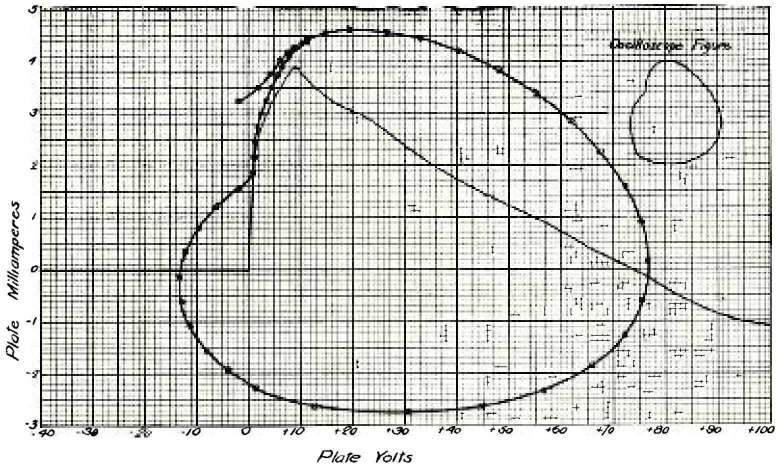

Fig. 17—Dynatron load line. Type 36 tube. Plate voltage = 22.5 volts.

frequency is $\dfrac{1}{35(1/72000)} = 2060$ c.p.s. The experimentally determined value was 2100 c.p.s., and is in good agreement.

From the points of the characteristic curve (load line for inductive branch) a time wave was drawn, and analyzed for its d-c component. The value obtained was 0.79 milliampere. The experimental value read on the d-c milliammeter varied from 0.75 to 0.8 milliampere during the test. In reference to this it is to be noted that the dynatron characteristic varied over any appreciable period of time, and was due apparently to decrease in secondary emission with time. Hence it was felt that only moderate agreement could be expected between the graphical and experimental values. The graphical figure was checked quantitatively against the oscilloscope figure after the latter was calibrated, and although very good agreement was obtained, it was decided that the precision of measurement was too low to warrant this check as a means of confirmation of the graphical method.

In Figure 17 is shown the characteristic for a plate supply voltage of 22.5 volts. There is a close similarity between the graphical and oscilloscope figures, particularly the cut-in at the left-hand side. The frequency (graphical) was 1750 c.p.s. (corresponding to 41 points), and the experimental value was the same. The d-c component was 1.42 ma. (graphical) as compared to 1.3 ma. average (experimental).

In Figure 18 is shown the characteristic for a plate supply voltage of 13.5 volts. The oscillations are comparatively feeble, and hence it was found advisable to enlarge the scale of the graph. The frequency was 1567 c.p.s. (graphical) corresponding to 46 points, as compared

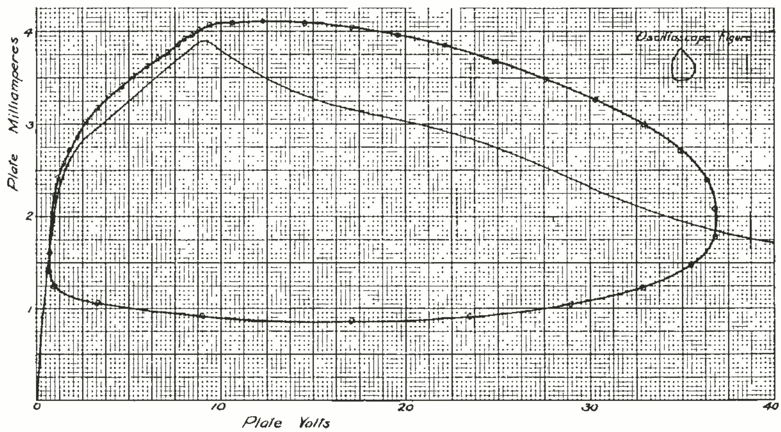

Fig. 18—Dynatron load line. Type 36 tube. Plate voltage = 13.5 volts.

to 1580 c.p.s. (experimental). The d-c component was 2.79 ma. (graphical) as compared to 3.04 to 3.28 ma. (experimental). Although the agreement is poor, it is to be expected considering the weak oscillations. No attempt was made to use a crest voltmeter in these runs as it was found that the capacitance of the leads and meter had an appreciable effect upon the characteristic. In the main it was felt that good experimental verification of the graphical method had been obtained, particularly for the larger amplitudes of oscillation.

Interesting and very simple graphical constructions and results can be obtained from the method just outlined if some of the circuit parameters are allowed to approach zero or infinity. Thus, if $L = 0$, we have a resistance R by-passed by a condenser C, and if the non-linear resistance r is that of a triode, which varies in known manner with time (signal voltage on grid known) then a transrectification diagram can be made if the triode be adjusted to act as a grid bias detector. However, lack of space here precludes any further discussion of these matters, and it may be merely mentioned that some of the results mentioned by R. Usui* can be verified by the construction given above.

* See Reference (1).

Conclusions

By breaking up a derivative ratio into two parts, and using one part as a finite operator curve, a graphical method of construction has been developed of wide scope and comparatively simple manipulation. In contrast to the usual method of isoclines,* each lineal element (starting with the initial point) helps to determine the next one, so that no visual judgment is required in choosing these to blend into a smooth curve.

Questions can be raised as to the smallness of time intervals required, the possibility of cumulative error in proceeding from point to point, and the effects of discontinuous variations in the non-linear elements. With regard to the latter, it is to be noted that the time interval can be decreased in the neighborhood of a discontinuity, and increased again in the more uniform portions of the characteristic.

In conclusion, the author wishes to express his gratitude to Mr. H. C. Wai, a graduate of RCA Institutes, Inc., for his invaluable aid in checking experimentally some of the constructions. The author also wishes to express his appreciation to Mr. C. E. Kilgour, Chief Research Engineer of the Crosley Radio Corporation, for his interest and sympathetic criticism of the methods, and to Professor Ernst Weber, Research Professor of Electrical Engineering at the Brooklyn Polytechnic Institute, for his checking of the theory and citing of references on the subject.

* See, for instance, Reference 11.

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REVIEW OF MICROPHONES

By

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IT IS a difficult task to classify microphones on a logical basis. Colloquial usage is broad and not precise. When the terms "dynamic" and "electro-dynamic" microphone are used they are generally regarded as referring to microphones having a coil moving in a magnetic field. Microphones constructed and operated in this fashion are usually termed "dynamic", although occasionally what is technically known as an "inductor" microphone is loosely characterized as "dynamic". There are various other types of microphones which do not have a designation based on scientific classification. Thus there are types which are known as velocity ribbon, pressure ribbon, sound-pressure, or uni-directional microphones or the like.

Technically, all of the preceding microphones are "electro-dynamic" devices. Each of them is a system having electrical, mechanical, and acoustical elements. The electrical and mechanical elements are either coils, straight conductors, or ribbons. The acoustic-exciter element is always air. In technical literature microphones may be described as "mass-controlled", or "resistance-controlled" electro-dynamic microphones. Colloquially the terms degenerate to "ribbons", "dynamics" or the like.

A close examination of the many different microphones on the market moreover reveals that it is not a simple matter to put microphones into a small number of groups. Thus one might be tempted to devise three classes—mass-, resistance-, and stiffness-controlled microphones—yet there are transducers which are controlled by one element within a certain frequency range and by another element within another region. A somewhat similar argument holds if one attempts to classify microphones according to whether their output voltage corresponds to variations in pressure, density, temperature, or particle-velocity of the sound wave. Perhaps the simplest classification would consist in merely stating the type of force available to actuate a microphone—whether pressure or pressure-gradient—and referring all other types to a special group.

THE PRESSURE-GRADIENT MICROPHONE

The quality of a transducer used to translate sound energy into electrical impulses is determined by four factors: frequency response, directional characteristic, freedom from noise, and absence of distortion. One way of stating the requirement for uniform frequency response in a microphone is by saying that the ratio of generated e.m.f. to the pressure or velocity in the sound wave shall be independent of the frequency. Another way of describing a uniformly sensitive electrodynamic microphone would be to say that the ratio of the velocity of the moving element to the pressure or velocity in the sound wave should be independent of the frequency. Thus in a high-quality velocity microphone we would expect the velocity of the ribbon to be proportional to the particle velocity; and in a high-quality moving-coil microphone, that the diaphragm with its rigidly attached coil would

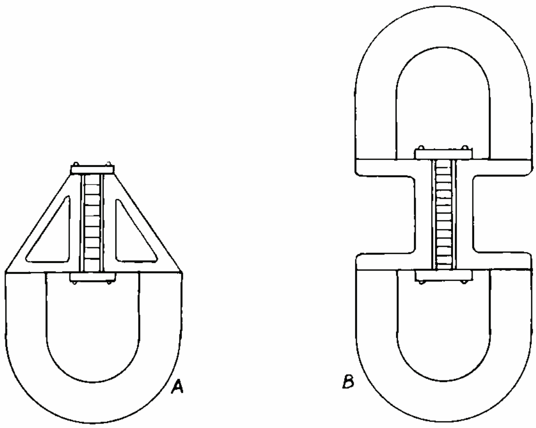

Fig. 1—Pressure-gradient microphones.

have at all frequencies the same velocity per unit pressure in the actuating sound wave. There are, of course, synonymous expressions to describe this quality, such as saying that the ratio of power output to sound intensity must be independent of the frequency, or that the mechanical impedance of the diaphragm be the same for all frequencies; in all cases, however, we expect to obtain a straight line when response is plotted against frequency.

Figure 1 shows the construction of two types of velocity ribbon microphones. Type B is 6 db more sensitive than Type A, as can at once be determined by considering the equation for the generated voltage in the ribbon

$$e = Blv/10^8 \text{ volts}$$

where B represents the flux density; l, the length of the ribbon, and v the velocity of the ribbon. The ribbon is driven from its equilibrium position by the difference in pressure which exists at the two sides of the ribbon. This movement induces an e.m.f. in the ribbon which later is suitably amplified. The aluminum-alloy ribbon, about 0.0005 cm. thick, 0.4 to 0.6 cm. wide, and from 4 to 5 cm. long, is corrugated to make it sufficiently stiff so that it can vibrate in a symmetrical fashion. The magnets are powerful permanent magnets, although in the earliest models, and still in laboratory standards, the magnetic flux is achieved by the use of electro-magnets.

The magnitude of the pressure difference is dependent on the path-length between the front and the back of the ribbon, and in the

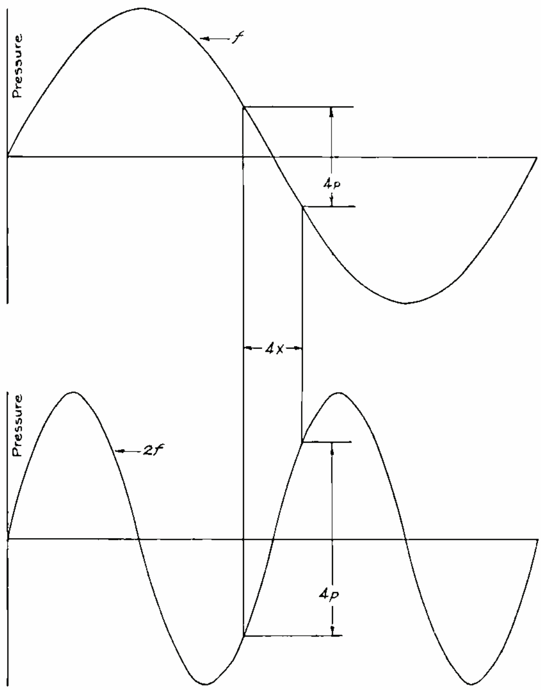

Fig. 2—Pressure-gradient curve.

main is a function of the size of the pole-pieces. It increases with an increase in this baffle-length, reaching a maximum when this length corresponds to half a wavelength. Hence, if the velocity of the ribbon is to be independent of the frequency, the driving force must increase with frequency in the manner in which the mechanical impedance of the mass-controlled element increases with the frequency of the actuating sound wave. It can be shown that the pressure-gradient is proportional to the frequency (see Figure 2), and also that the most important terms constituting the total mechanical impedance are functions of frequency. Therefore, the generated e.m.f. is free from frequency discrimination, except when the microphone is very close to the source of sound, a condition which shall be considered later. Because the ribbon is so light, its mechanical impedance closely approaches the acoustic impedance of the sound wave to a frequency of about 5,000 cycles. Hence for frequencies below this value, when the "fluid resistance" becomes negligible in proportion to the mass-reactance of the ribbon, the ribbon actually follows the motion of the air particles.

In the RCA velocity ribbon microphone the ribbon is connected to a step-up transformer which is a part of the microphone proper. Connections between ribbon and transformer primary winding are made by two soldered leads at the ribbon clamps to insure a secure

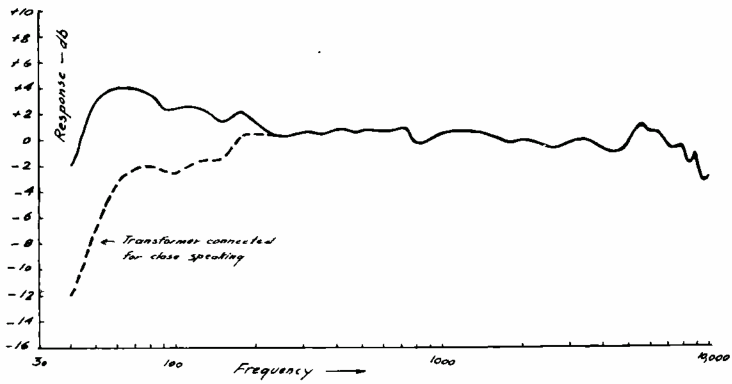

Fig. 3—Calibration curve of velocity microphone.

contact. It may be noted that the slits between the ribbon and the pole-pieces have no effect on the frequency response of the microphone; if these slits are made larger than a few mils, the net effect will be a reduction in sensitivity only. The resonant frequency of the ribbon is made lower than the lowest frequency to be reproduced, and is generally of the order of from 15 to 25 cycles.

If the microphone is near a source, the actuating sound waves will be spherical, and the response for the low frequencies will be increased. This is because the particle velocity, of which this microphone gives a measure, is a funtion of l/r^2, where r is the distance between source and microphone; while the pressure, of which a pressure-operated microphone gives a measure, is a function only of l/r. When r becomes large, however, or when the frequency is high, the particle velocity is closely proportional to l/r. It is important, therefore, to have the source of sound at least 3 feet distant from the microphone if no increased low frequency is desired. Since speakers and singers before a microphone often have a tendency to step up close to the instrument, a microphone has been built in which the secondary windings of the microphone transformer can be connected in such a fashion that this aforementioned increase in low-frequency response is compensated for electrically (see Figure 3).

Because the microphone employs an open-field structure so that the acoustic air pressures have ready access to both sides of the ribbon, an internal wind screen is frequently provided to make the microphone less susceptible to air currents. Sometimes, also, a silken hood is put over the microphone when it is used on location in motion-picture production.

The directional characteristic of the velocity microphone resembles the figure eight, and is the same for practically all frequencies. For this reason the microphone does not produce frequency distortion due to its directional quality.

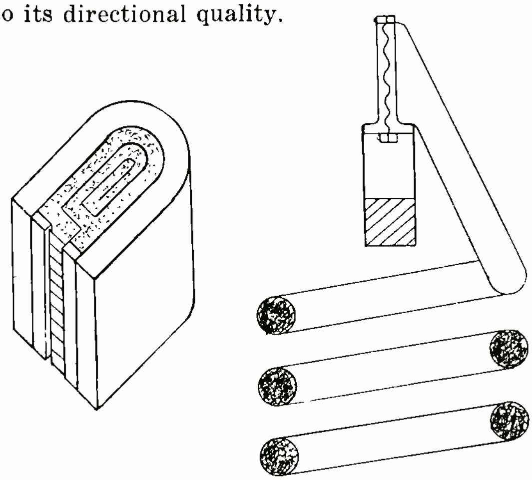

Fig. 4—Pressure-ribbon microphones.

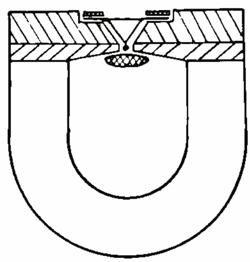

Fig. 5—Inductor microphone.

PRESSURE-OPERATED MICROPHONE

The pressure-ribbon microphone differs from the velocity-ribbon microphone in that one side of the ribbon is closed in by an acoustic resistance consisting of tufts of felt placed in a spiral pipe which is sealed against the pole-pieces. The ribbon, therefore, is actuated by the pressure of the incident sound wave. Since the sound pressure is inactive on the side of the ribbon facing the acoustic labyrinth, the microphone is non-directional and the ribbon can be driven from its equilibrium position even when the sound comes from a direction in plane of the ribbon. Hence the directional characteristic is that of a sphere, at least so long as the dimension of the microphone is comparable to the wavelength of the incident sound.

In order to obtain a straight frequency-response curve, the velocity of the ribbon must be independent of frequency. Hence the acoustic impedance given by p/v must be independent of frequency. This impedance consists of the mass reactance of the ribbon, the air-load on the ribbon, the compliance of the suspension, and the resistance of the acoustic labyrinth. Since not all of these terms are independent of frequency, it becomes necessary to adopt a construction in which the terms which are independent of frequency are predominant. This, of course, is achieved by making the microphone resistance-controlled, and requires a sufficiently long pipe with the tufts of felt so placed at intervals that the acoustic reactance term becomes negligible and the pipe presents in the main an acoustic resistance. Figure 4 shows two types of pressure-ribbon microphones.

THE INDUCTOR MICROPHONE

The inductor microphone consists of a diaphragm having attached to it a straight conductor which can vibrate in a magnetic field. The

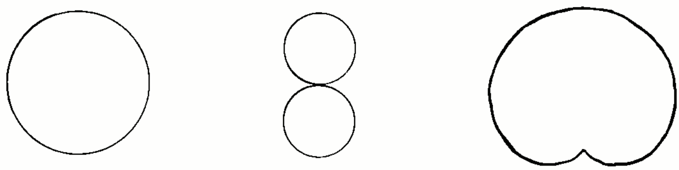

Fig. 6

(a) Directional response of pressure-operated microphone. (b) Directional response of pressure-gradient microphone. (c) Directional response of uni-directional microphone.

diaphragm, which may be of metal or impregnated paper, is in form of a "V" trough (see Figure 5). In order that the diaphragm have at all frequencies the same value of velocity per unit pressure in the actuating sound wave, the controlling element must again be a resistance. This resistance must be small, however, in order to obtain enough sensitivity, because the velocity is inversely proportional to the resistance. This means that compensating measures must be taken to offset the stiffness reactance at the high frequencies, stiffness and mass having to be small for sufficient output. Hence acoustic resonance circuits are incorporated in the construction of this microphone. Thus a small pipe connected to the atmosphere is placed in the microphone to "boost" the low frequencies, while a bolt of silk is placed directly behind the pole-pieces to provide compensation for the mass reactance of the diaphragm. Because of pressure doubling, giving rise to an increased response at the higher frequencies, from one to three layers of silk are placed in front of the diaphragm to counteract this effect. A suitable transformer, placed within the microphone housing, is used to raise the value of the electric impedance of the conductor to a value appropriate for the input circuit of a vacuum tube amplifier.

A velocity ribbon microphone may be combined with a pressure ribbon microphone to give a transducer having a directional characteristic as shown by Figure 6(c). Since it is important that the two ribbons are as close together as possible, the microphone is so constructed that half of the ribbon is actuated by the pressure, and the other half, by the difference in pressure between the two sides of the ribbon. Because it is difficult to construct a pressure operated microphone the frequency response of which does not rise toward the high end of the audio spectrum, this microphone shows a slight increase in response for the higher registers. Theoretically it is possible, of course, to construct such a "uni-directional" microphone in a variety of fash-

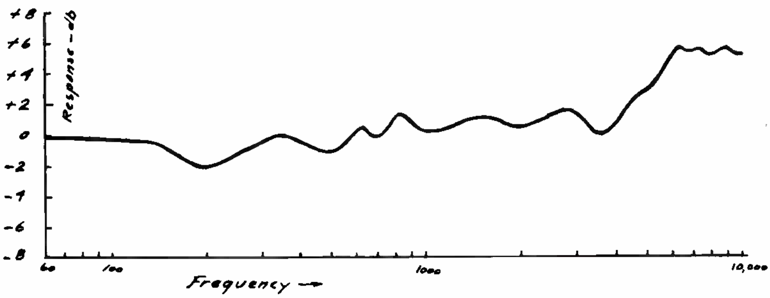

Fig. 7—Frequency response of uni-directional microphone.

ions, and even an "inductor" or "moving-coil" microphone might be combined with a velocity ribbon microphone to give a transducer of cardioid directional characteristic. Again a slight increase in low frequency response may be expected if the source of sound is close to the transmitter, although this increase is less pronounced than in the ordinary velocity ribbon microphone.

It is important of course that a proper phase relationship exist between the two microphones. When a plane wave strikes the uni-directional microphone from the front, the two ribbons must vibrate exactly in phase, while when the sound comes from the back, the two ribbons must vibrate with a phase difference of 180 degrees.

The advantages of a uni-directional microphone lie mainly in its uniform response, in the large solid angle of reception over which the microphone receives sound without appreciable attenuation, and the fact that the microphone can be so oriented that undesirable noises coming from a certain direction can be prevented from actuating the microphone. Figure 7 shows the frequency-response curve of a uni-directional microphone.

THE REQUIREMENTS AND PERFORMANCE OF A NEW ULTRA-HIGH-FREQUENCY POWER TUBE

By

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RADIO communication has been a very rapidly developing and growing art and today has advanced to a stage where it is a well established industry. However, it still has frontiers where more knowledge is desirable and where the equipment is insufficient to meet the latest requirements. One of the most intriguing of these frontiers is that of the high-frequency limit and the power available for its exploration.

Let us consider that at the present time the useful frequency limit has been pushed up to 60 megacycles (60,000,000 cycles per second), or a wavelength of five meters. It is now becoming apparent that in the very near future the range from 60 to 300 megacycles, or from 5 meters down to 1 meter, will be considered a familiar and useful territory. The available frequency range will then be enlarged to five times its present extent.

Some of the problems introduced by the attempt to raise the high-frequency limitations of radio equipment can best be appreciated by talking in terms of wavelengths. The wavelength determines directly in the same units the approximate maximum physical size which the equipment to produce that frequency may attain. This results from the fact that the greatest speed with which energy may be sent along electrical circuits is the same as that of electromagnetic energy in space, the velocity of light. Usually, however, we do not approach the ideal maximum of the velocity of light due to the fact that lumped capacity or lumped inductance lowers the speed in electrical circuits. The circuits must then be much smaller in extent than the wavelength, and the tubes themselves must be even smaller in physical size.

In the design of a tube the time required for an electron to travel even a small distance introduces another very serious problem. For instance, consider the case of an electron being accelerated away from a cathode by a potential of 100 volts applied to a grid. If the time of transit of the electron from the cathode to the grid is not to exceed one-sixth of the time period of a one-meter oscillation, the spacing must be less than 3/64 of an inch. For best operation of a tube, the transit time should be kept much smaller than a sixth of a cycle, a requirement which necessitates close spacing of the tube elements.

These limitations in physical size must be weighed directly against the demands for increased power. For the control of power, the tube and circuits must have a size somewhat in proportion to the power controlled because of the inherent losses that must accompany the flow of current and the need for sufficient insulation to withstand the voltages present. The size may be kept down by increasing the intensity of the electrical and thermal phenomena present. But very real physical and practical limitations are soon met which prevent this method of attack from providing substantial gains.

However, demands for greater power at higher frequencies are continually being made. In the past such demands have always been justified in view of the improved reliability and service that resulted. In the ultra-high-frequency field the exploratory work has been done with small power and limited equipment, exactly as was the case with the lower frequencies, but now, as before, the problem is to attain greater controlled power.

Up to the present time no practical method of amplifying radio-frequency signals is known other than the use of the conventional type of vacuum tube in which an input signal is applied to the grid circuit and controlled power is taken from the plate circuit. Other types of radio-frequency primary generators are known, the most important of which are the magnetron and the Barkhausen-Kurz type of oscillator. Both of these are subject to severe limitations in frequency stability and control. These facts make them much less desirable than conventional tubes used as amplifiers and carried to as high frequencies as they are capable of operating.

The requirements of a vacuum tube for amplifying large power at ultra-high frequencies with the conventional type of power amplifier are as follows:

1 — The tube must have elements and lead wires which will permit a large fraction of the total resulting circuit to be outside the tube.

2 — All electron paths must be short in order that the time of transit from the cathode to plate will be small.

3 — The insulation must not introduce losses or permit breakdown under the applied voltages.

4 — The leads must be capable of carrying the r-f currents because they are part of the electrical circuits.

5 — The cathode must be rugged to withstand both ion and electron bombardment.

6 — The power-controlling and power-dissipating elements must be capable of handling a large wattage per unit area.

7 — The tube must be adaptable to r-f circuits.

All present conventional tube types meet these requirements in some degree. However, the degree for each tube type is apt to depend

Fig. 1—Water-cooled high-frequency triode, RCA 888 750 watts Class C output. Overall length is 7 inches from top of grid post to near end of flexible filament leads.

on the requirements of the art at the time of development of the tube. It is not very far back to the time when 8 and 15 megacycles lay in pioneer territory and transmitters were temperamental and tricky devices to handle. (Some still are.) Most earlier types of vacuum tubes proved satisfactory for the needs of the art at that time. However, when such tubes are forced to operate at frequencies far beyond their original design, it often happens that phenomena not anticipated cause them to fail prematurely. Today new tube designs must stand complete performance tests up to at least 30 and 60 megacycles.

A water-cooled triode developed to meet the special requirements in the region above 60 megacycles is shown in Figure 1. Figure 2 is a sectional cut of the tube and shows the rugged compact structure. No

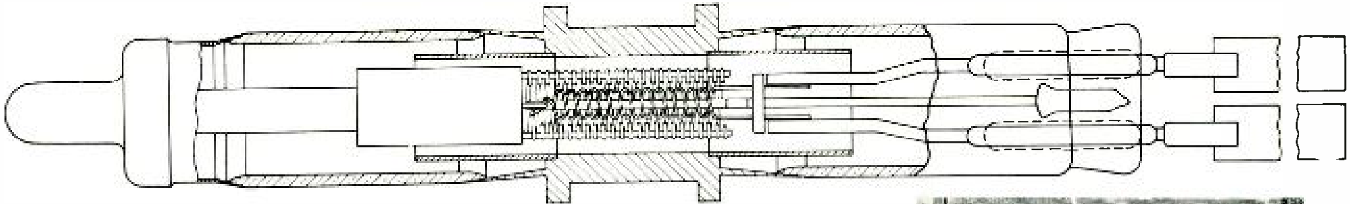

Fig. 2—Sectional view of water-cooled high-frequency triode, Type 888, without the integral water jacket and grid shield.

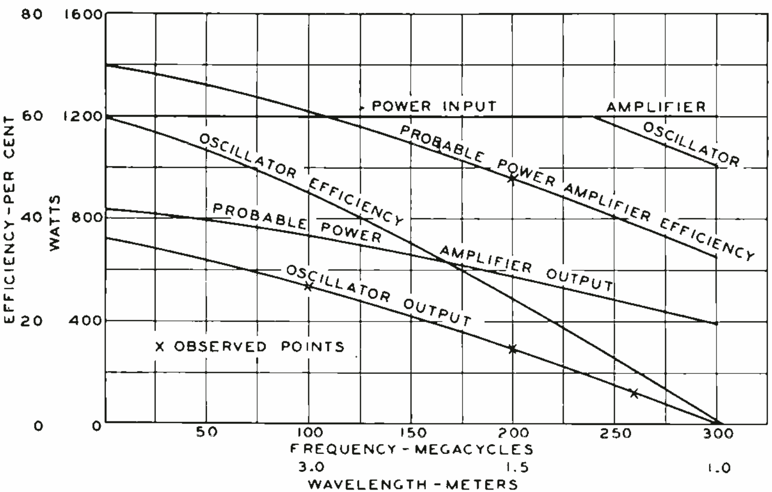

Fig. 3—Efficiency and output curves of the 888 as a function of frequency.

Fig. 4—90 megacycles (3.3 meters) push-pull oscillator employing two 888's and delivering 1100 watts to the lamp load.

internal insulation is used; the filament and grid structures are self-supporting from the glass envelope. The cylindrical plate at the center is water-cooled and has a dissipation rating of 1000 watts. The cathode is a pure tungsten filament in the form of a double helix. Such a rugged cathode is quite desirable because of the increased magnitude of r-f charging currents at ultra-high frequencies, the high voltages involved, and the electron bombardment that occurs. Some electrons fail to follow the quick changes of grid potential at such very high frequencies, are halted, and driven back to bombard the cathode.

In view of the requirements which have been established it should be noted that the filament and grid leads are as short as possible consistent with the high plate voltage, and a safe temperature of the glass seals which are necessary to bring the leads to the external r-f circuit. The leads have been made large and the envelope small to permit low inductance circuits to be formed. The interelement spacings have been reduced to as small values as are practical for the power involved in order to reduce the time of transit of electrons from filament to plate. The lack of internal insulation eliminates the major insulator problem. The large size of grid and filament leads used to obtain low inductance will also carry high r-f charging currents. The pure tungsten filament, tantalum grid, and water-cooled plate are the most rugged elements readily available for use in vacuum tubes. All dimensions of the tube are small and the external circuit may be placed compactly about the tube to form good r-f circuits.

Tentative ratings of this tube, RCA-888, for Class C Telegraphy are as follows:

Filament Voltage	11 volts
Filament Current	24 amperes
Amplification Factor	30
Max. d-c Plate Voltage	3000 volts
Max. d-c Grid Voltage	—500 volts
Max. d-c Plate Current	0.400 ampere
Max. d-c Grid Current	0.100 ampere
Max. Plate Input	1200 watts
Max. Plate Dissipation	1000 watts

A similar tube, Type RCA-887, has also been developed which has an amplification factor of 10.

In Figure 3 are given curves of the plate-circuit efficiency and useful power output as a function of frequency for operation as a self-excited oscillator and for the expected efficiency and tube output as a neutralized power amplifier.

The efficiency falls off with increasing frequency due to the fact that electrons take a finite amount of time to go from filament to plate.

At higher frequencies the time of transit becomes a larger and larger part of the cycle. In an oscillator it is usually impossible to adjust the phase of the swinging plate voltage so that the electrons arrive close to the time when the plate voltage swings down to low values. Thus, the electrons are drawn to the plate under the action of high plate voltages and the plate losses are correspondingly high. In the case of the neutralized power amplifier due to the isolation of grid and plate circuits the phase of the plate voltage can be adjusted freely to have its low potential swing correspond approximately to the arrival of electrons. Thus, the total energy of the electron given up at the plate is less, and the efficiency does not fall off as rapidly.

Figure 4 shows two of these tubes in a standard push-pull oscillator circuit operating at 90 megacycles (3.3 meters). The useful output of

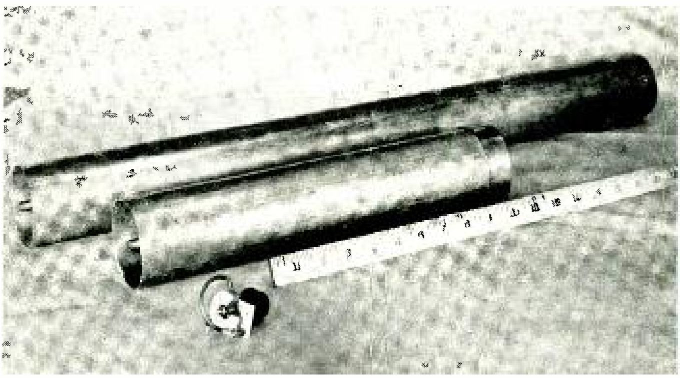

Fig. 5—1.5 meter resonant circuits: lumped capacity-lumped inductance circuit, quarter-wave closed-end line circuit, and half-wave open-end line circuit.

1100 watts is being dissipated in the bank of 8 standard 125-volt 300-watt lamps. Due to the high voltages developed across the long lead wires within each lamp it is not possible to operate the lamps at more than about half rating. The lamps are connected across a turn and a quarter of the two-turn plate coil, which is about a 5-inch length of 5/16 inch diameter copper tubing.

Probably one of the most important considerations in designing tubes and circuits to handle large powers at ultra-high frequencies is that of proper construction of the circuits to attain large dimensions. Figure 5 shows three electrical circuits each resonant to 200 megacycles or 1.5 meters. The midget type of variable condenser needs only about 2½ inches of wire length and about a fourth of its maximum capacity setting to form the resonant circuit. However, the transmission-line type of resonant circuits with distributed inductance and distributed capacity can be a full quarter of a wavelength long for the closed-end type, and a full half of a wavelength long for the open-end type, as shown in Figure 5. The capacity per unit length of these line circuits can be varied without changing their overall length because the distributed inductance per unit length changes proportionally. For instance, the diameter of the small center tubing in the line circuits shown in Figure 5 could be increased until the radial clearance to the outer tubing is no more than the separation between the plates of the midget variable condenser shown in the picture, and the line circuits would still be a quarter of a wavelength and half of a wavelength long, respectively.

At the ultra-high frequencies the tube and its external circuit should be considered as integral parts of the electrical circuit. The interelectrode tube capacities and the inductance of the tube leads are a large portion of the total capacity and inductance of the electrical circuit. In order to attain the maximum overall size of the circuit, the tube elements should constitute a portion of a transmission line circuit which may be extended outside the tube. This ideal is difficult to achieve because of the close spacing required between the cathode, grid, and plate, and the larger spacing required between the leads to the electrodes to allow for insulation clearances, glass seals, neutralization circuits, and any other circuit irregularity. The desirable condition of uniformly distributed inductance and capacity along the tube leads and active elements has been approached in the design of the RCA-888. Though the ideal is not attained the arrangement does make possible the use of larger transmission line circuits than with other tubes, and hence larger power at higher frequencies.

I wish to acknowledge the many contributions of those associated in the work of developing these tubes and especially the help of Mr. J. B. Fitzpatrick who aided greatly by solving many of the basic problems of manufacture.

HORN LOUD SPEAKERS

BY

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PART II. EFFICIENCY AND DISTORTION[1]

INTRODUCTION

LARGE scale reproduction of sound involving several acoustical watts output is becoming quite commonplace. Since high power amplifiers are costly, it is logical to reduce the amplifier output to a minimum by the use of high efficiency loudspeakers. At the present time horn loudspeakers seem to be the only satisfactory high efficiency system for large-scale reproduction. For applications requiring high quality reproduction of intense sound, some consideration should be given to the introduction of frequencies not present in the output due to nonlinearity of the operating characteristics of the elements which constitute the vibrating system of the loudspeaker. It is the purpose of this paper to consider some of the factors which influence the efficiency and distortion characteristics of a horn loudspeaker.

EFFICIENCY

The absolute efficiency of a loudspeaker is the ratio of the useful acoustical power radiated to the electrical power supplied to the load, the current wave of which exercises a controlling influence on the shape of the sound pressure. This definition of efficiency excludes all incidental power supplied for such purposes as transformer losses, field excitation, etc. Sometimes another definition[2] is used involving both the vacuum tube and the loudspeaker. However, at the present time, there are many types of power vacuum tubes, ranging from those designed to feed into a load greater than the internal impedance of the tube to those designed to feed into a load smaller than the internal impedance of the tube. Therefore, in order to simplify the following discussion, the absolute efficiency will be used rather than the definition involving the vacuum tube.

The absolute efficiency[3] of a loudspeaker may be expressed as,

$$EFF = \frac{r_{em}}{r_{em} + r_{ed}} \tag{1}$$

where, r_{em} = motional resistance, ohms,
r_{ed} = damped resistance of the voice coil, ohms,

[1] Part I, Impedance and Directional Characteristics, RCA REVIEW, April, 1937.

[2] Olson and Massa, Applied Acoustics, P. Blakiston's Son & Co., Philadelphia, p. 253.

[3] Kennelly and Pierce, *Proc. A.A.A.S.* Vol. 48. No. 6. 1912.

The motional resistance of the mechanical system of a dynamic loudspeaker is the real part of,

$$z_{em} = \frac{(Bl)^2}{z_m} \times 10^{-9} \text{ ohms} \tag{2}$$

where, B = flux density, gausses,
l = length of wire in the voice coil, centimeters,
z_m = mechanical impedance of the vibrating system at the voice coil, mechanical ohms.

This expression of efficiency assumes[4] that no energy is lost in the form of mechanical hysterisis, but that all of the motional resistance is due to useful acoustic radiation and that the force factor is real.

Equation 2 may also be used to determine the efficiency of a loudspeaker experimentally. As in all tests of this kind care must be taken to obtain conditions which will insure the validity of the measurements.

Horn loudspeakers are usually of the dynamic type incorporating features which result in high efficiency. Therefore Equation 2 may be used to predict the efficiency and performance quite accurately. It is the purpose of this section to consider the performance of horn loudspeaker systems using the motional resistance method as a basis for comparison. The principal parameters which govern the performance are the horn dimensions, the diaphragm area, mass and suspension stiffness, the voice coil, the flux density and the air chamber coupling the diaphragm to the horn.

The first consideration will be a simple system consisting of a dynamically-driven diaphragm coupled to a horn having a throat impedance which is an acoustic resistance of constant value. To further simplify the problem, the stiffness of the diaphragm suspension will be considered negligible and the capacitance of the air chamber will be assumed to be zero. The problem is to consider the efficiency of the loudspeaker as a function of the throat dimensions.

The expression for the efficiency of this system is given by,

$$EFF = \frac{42\,(Bl)^2\,A_D^2\,A_T}{r_{ed}\,[\,(42A_D^2)^2 + (\omega m A_T)^2]\,10^9 + 42\,(Bl)^2\,A_D^2\,A_T} \tag{3}$$

where, A_D = area of the diaphragm, square centimeters,
A_T = area of the throat, square centimeters,
m = mass of the diaphragm and voice coil, grams.

[4] In all of the systems considered in this section, the radiation from the free or open side of the diaphragm has been neglected. In certain cases at the higher frequencies this may be of the order of several per cent of the total output. For a consideration of the air-load on the back or open side of the diaphragm see H. F. Olson, *Jour. Acous. Soc. Amer.* Vol. 2, No. 4, p. 485.

To illustrate the relations between the various factors in Equation 3 the efficiency characteristics, for various loads upon a diaphragm driven by an aluminum coil of equal mass operating in a field of 22000 gausses, are shown in Figure 1. These data show that it is comparatively simple to obtain high efficiencies at the lower frequencies. However, at the higher frequencies the efficiency is limited by the mass of the diaphragm and voice coil. In this example the mass of the diaphragm has been chosen equal to the mass of the coil. It has been found that a diaphragm lighter than the voice coil is usually too fragile to be of value and these results represent the practical limit of mass reduction.

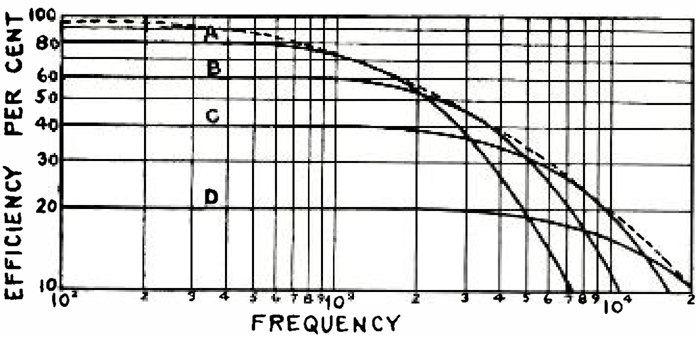

Fig. 1—Efficiency characteristic of a 1-gram aluminum voice coil, in a 22000-gauss field, driving a 1-gram diaphragm. The graphs are for the following values of mechanical resistances presented to the diaphragm: A, 16000; B, 42000; C, 96000; D, 254000 mechanical ohms. Dotted curve shows the maximum efficiency possible at each frequency.

The data of Figure 1 show that, to obtain the maximum efficiency[5, 6, 7, 8] over a wide frequency range with a system of this type, it is advantageous to use two or more units, each one designed for maximum efficiency over the frequency range which it supplies.

The equivalent electrical circuit and typical efficiency characteristic of a system of a single degree of freedom, consisting of a tuned cone coupled to a large throat horn, is shown in Figure 2. High efficiency is obtained over a relatively narrow range. Such a system may be used for speech reproduction. However, a "broader" frequency characteristic which may be obtained with two or more degrees of freedom results in more natural reproduction and better articulation.

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[5] Hanna, C. R., *A.I.E.E.* Vol. 47, No. 4, p. 253.
[6] Wente and Thuras, *A.I.E.E.* Vol. 53, No. 1, p. 17.
[7] Olson and Massa, *Jour. Acous. Soc. Amer.* Vol. 8, No. 1, p. 48.
[8] Massa, F. *Jour. Acous. Soc. Amer.* Vol. 8, No. 2, p. 126.

A system of two degrees of freedom, consisting of a dynamically driven tuned diaphragm coupled by means of an air chamber to a horn, is shown in Figure 3.

The acoustic impedance at p is given by,

$$z = \frac{z_1 z_2 + z_1 z_3 + z_2 z_3}{z_2 + z_3} \tag{4}$$

where $\quad z_1 = j\omega M + \dfrac{1}{j C_1}$

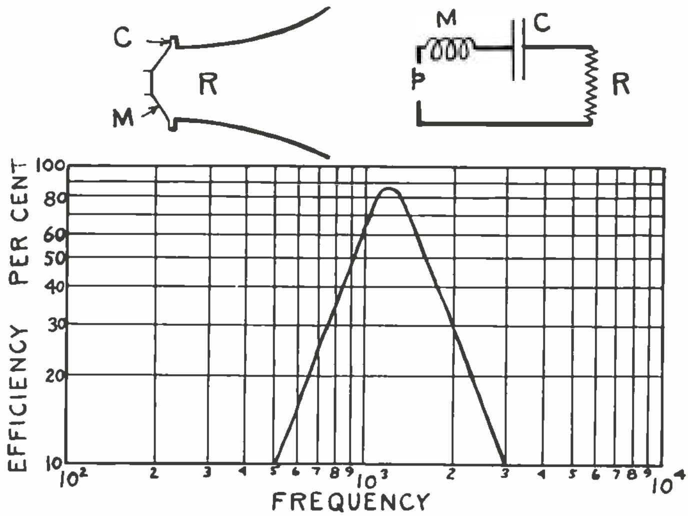

Fig. 2—Efficiency characteristic of a system of a single degree of freedom.

$$z_2 = \frac{1}{j\omega C_2}$$

$z_3 = R$
$M =$ inertance of the cone,
$C_1 =$ capacitance of the diaphragm suspension system,
$C_2 =$ capacitance of the air chamber,
$R =$ horn throat acoustic resistance.

The efficiency of this system may be predicted by means of Equations 1, 2, and 4. A typical efficiency characteristic of two degrees of freedom is shown in Figure 3. A loudspeaker having this type of efficiency characteristic has been found to be useful for speech reproduction at very high sound levels.

The results shown in Figure 1 were obtained by assuming the capacitance of the air chamber to be zero. In general, it is impractical to design a high efficiency loudspeaker to cover a wide frequency range without an air chamber, because the diaphragm is usually larger than the throat. However, the addition of an air chamber[9, 10, 11, 12] is actually an advantage because the efficiency may be increased over a wide frequency band by employing an appropriate design. The air chamber introduces a capacitance in parallel with the horn throat impedance which reduces the effective reactance of the vibrating system. The

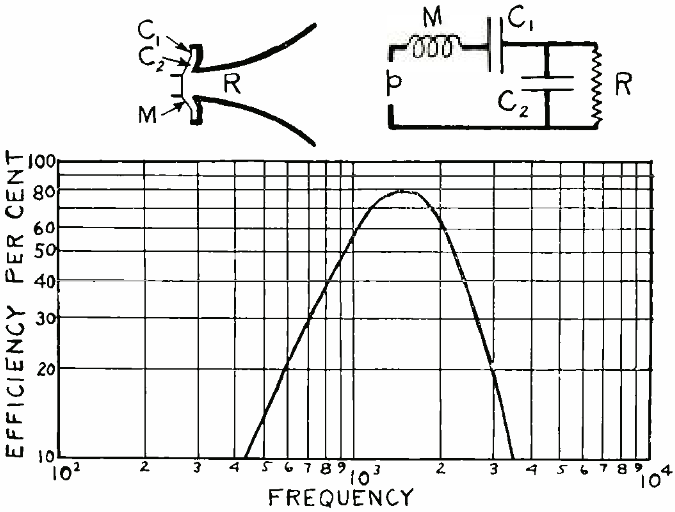

Fig. 3—Efficiency characteristics of a system of two degrees of freedom.

efficiency characteristic of a horn loud speaker with and without an air chamber is shown in Figure 4. These results show that the efficiency is substantially increased over a range of two octaves by the introduction of an air chamber. Due to the sharp high frequency cut-off the air chamber should be designed so that this occurs at the upper frequency limit of the reproducing system.

The results shown in Figures 3 and 4 indicate that the effective mass of the vibrating system may be reduced and the response increased by means of a multi-resonant system. This method provides a means for reducing the effective mass and for improving the efficiency.

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[9] Hanna and Slepian, *A.I.E.E.* Vol. 43, No. 3, p. 251.
[10] Wente and Thuras, *Bell System Tech. Jour.* Jan. 1928, p. 140.
[11] Olson, H. F. *Jour. Acous. Soc. Amer.* Vol. 2, p. 242.
[12] Wente and Thuras, *A.I.E.E.* Vol. 53, No. 1, p. 17.

The vibrating systems considered in this section have employed aluminum voice coils operating in an air gap of 22000 gausses and driving either fibre, paper, bakelite or aluminum-alloy diaphragms. These materials represent the practical limit for obtaining high efficiencies at the present time. When new materials are developed which will permit an increase in the air-gap flux density or a reduction in mass of the diaphragm or a reduction in the resistance-density product of the voice coil it will be possible to improve the efficiency.

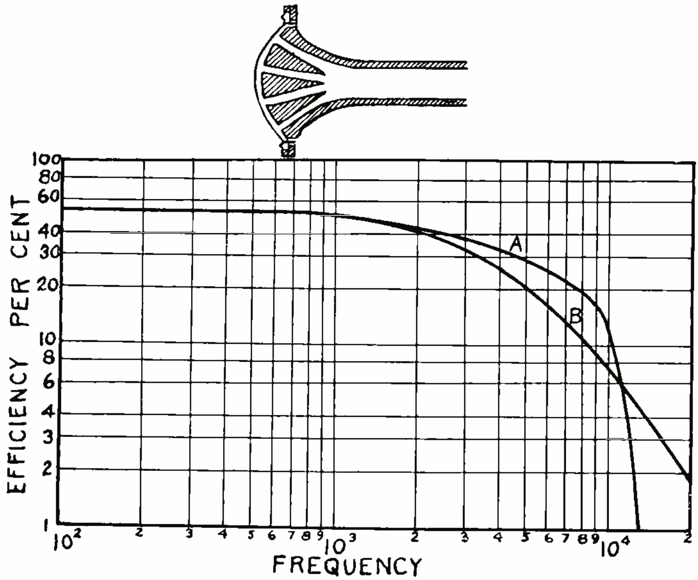

Fig. 4—A, Efficiency characteristic of a diaphragm coupled to a horn by means of an air chamber of the type shown. B, Efficiency characteristic in the absence of an air chamber. The air chamber shown consists of annular slits coupling the diaphragm to the horn throat. The same increase in efficiency can be obtained with other designs—for example, a series of cylindrical holes coupling the diaphragm to the horn.

The voice coil, in high-power loudspeakers may run at a very high temperature which results in a reduction of efficiency. The damped resistance is given by,

$$r_{ed} = (1 + kt)\, r_{edo} \tag{5}$$

where $k = .0042$

$t =$ temperature centigrade

$r_{edo} =$ resistance at 0° C.

The efficiency as a function of the temperature, for various initial efficiencies at 20° C, is shown in Figure 5. These results show that the effect of temperature in reducing the efficiency is most pronounced in loudspeakers of low efficiency.

Distortion and Overload

It is the purpose of this section to consider some types of amplitude distortion which may occur in horn loudspeakers together with factors which determine the loudspeaker rating.

In general, a sound wave of large amplitude cannot be propagated in air without a change in wave form and as a result the production of harmonics. If equal positive and negative changes of pressure are

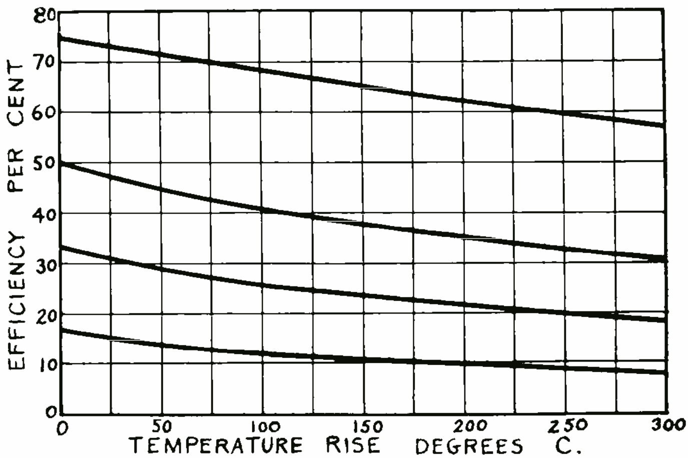

Fig. 5—The effect of temperature rise upon the efficiency of a loud speaker, with an aluminum voice coil, having the indicated initial efficiencies. Initial temperature 20° C.

impressed upon a mass of air the changes in volume of the mass will not be the same. The volume-change for an increase in pressure will be less than the volume-change for an equal decrease in pressure. Physically, the distortion may be said to be due to the non-linearity of air.

The magnitude of the harmonic frequencies may be obtained theoretically from the differential equation of wave propagation. Rocard[13] was the first to investigate the generation of harmonics in an exponential horn. The subject was later investigated theoretically and experimentally by Thuras, Jenkins and O'Neil[14] and theoretically by

[13] Rocard, *Comptes Rendus*, Vol. 196, p. 161, 1933.

[14] Thuras, Jenkins and O'Neil, *Jour. Acous. Soc. Amer.* Vol. 6, p. 173, 1935.

Goldstein and McLachlin[15]. For constant sound-power output the distortion is proportional to the square of the frequency. Further, the nearer the observation frequency is to the cut-off frequency the smaller the distortion. For this reason there is an advantage in the use of two or more units dividing the range into two or more parts.

The distortion due to non-linearity of the air is, at the present time, one of the most important as well as troublesome factors in the design of high efficiency loudspeakers for large outputs. In order to obtain high efficiency, particularly at the higher frequencies, it is necessary to couple the relatively heavy diaphragm to a throat having an area

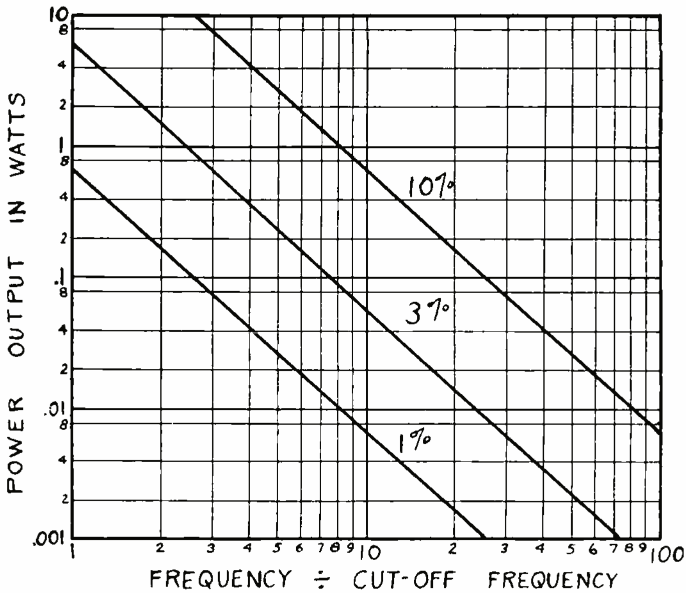

Fig. 6—The power output of infinite exponential horns, per square centimeter of throat area, for 1, 3 and 10 per cent distortion, as a function of the ratio of the frequency under consideration to the cut-off frequency.

small compared to the diaphragm. On the other hand, for certain allowable distortion the power output is directly proportional to the area of the throat. As a consequence, to deliver large outputs at high efficiencies requires a very large throat which may be suitably coupled to a correspondingly large diaphragm or a large number of lightly driven small throat units.

The power which can be transmitted per square centimeter of throat area of an infinite exponential horn, as a function of the ratio of the

[15] Goldstein and McLachlin, *Jour. Acous. Soc. Amer.* Vol. 6, p. 275, 1935.

frequency under consideration to the cut-off frequency, with the production of 1, 3 and 10 per cent distortion is shown in Figure 6. For sake of generality the curves shown in Figure 6 refer to an infinite horn. Furthermore, the increase in power which may be transmitted by a practical finite horn is only a few per cent greater than that shown in Figure 6.

In practical applications the distortion due to the nonlinearity of the air is most important in theatre reproduction where high-power, wide-range loudspeakers with small distortion are required. The permissible distortion is usually placed at 3 per cent. The average throat area of commercial high-frequency loudspeakers, available to-day, of the type shown in Figure 9 of Part I is approximately 10 sq. cm. The power which this loudspeaker can deliver with 3 per cent distortion is 10 times the value shown in Figure 6. For a cut-off of 200 cycles the power output is 10 watts at 500 cycles, 2.5 watts at 1000 cycles, .6 watts at 2000 cycles, etc. Taking into account the energy distribution of speech and music as a function of the frequency, this distortion characteristic has been found to be satisfactory at the present time.

In general, acoustic and electrical networks are assumed to be invariable; that is, the constants and connections of the network or system do not vary or change with time. A network which includes a circuit element that varies continuously or discontinuously with time is called a variable network. In some cases the variable elements are assumed to be a certain function of the time; that is, the variations are controlled by outside forces which do not appear in the equations or statement of the problem. In another type of variable-circuit element the variation is not an explicit time function, but a function of the current (and its derivatives) which is flowing through the circuit.

An example of the latter type of variable-circuit element in an acoustical system is the air-chamber capacitance in a horn loudspeaker. The excursions of the diaphragm changes the capacitance. The acoustic capacitance of the air chamber, Figure 7, is given by,

$$C = \frac{V}{\rho c^2} = \frac{A(d + x)}{\rho c^2} \tag{6}$$

where
ρ = density of air, grams per cubic centimeter,
c = velocity of sound, centimeters per second,
V = volume of the air chamber, cubic centimeters,
A = projected area of the air chamber upon the diaphragm, sq. cm.
d = distance between the diaphragm and front boundary of the air chamber in the absence of motion, centimeters,
x = displacement of the diaphragm, centimeters.

In general, the distortion which this variable element introduces is small because for constant-sound output the amplitude of the diaphragm is inversely proportional to the frequency. At the low frequencies where the amplitude of the diaphragm may be so large that the volume becomes alternately zero and two times the normal volume the reactance of the capacitance is very small compared to the resistance. At the high frequencies where the reactance of the capacitance is comparable to the resistance, the amplitude of the diaphragm for the same output is so small that the variation in capacitance may be neglected. However, the picture is changed somewhat when both a high and a low frequency are impressed upon the system. Under these conditions considerable change in capacitance occurs due to the large amplitude of the diaphragm for the impressed low-frequency. The

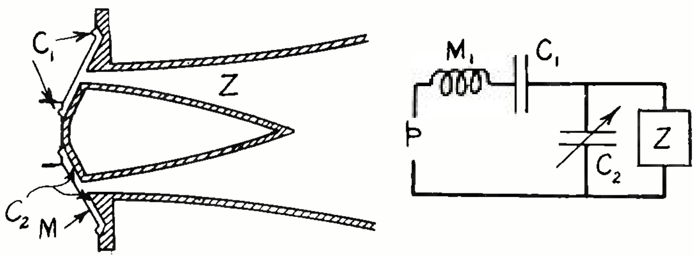

Fig. 7—A mechanism with an air chamber coupling the diaphragm to the horn. The variation in volume of the air chamber introduces a non-linear element in the form of the capacitance C_2. The equivalent electrical circuit indicates the effect of the non-linear element upon the system.

resultant change in capacitance introduces a variable element, for the impressed high-frequency, which may have variations in impedance as large as the impedance of the other elements of the system. When this condition obtains, particularly with close spacing between the diaphragm and front boundary of the air chamber, the distortion may be tremendous.

In the above discussion the air chamber is assumed to be a pure capacitance. This assumption is not correct at the higher frequencies where the dimensions of the air chamber are comparable to the wavelength. Regardless of the form of this impedance, it is nevertheless a function of the spacing between the diaphragm and the air chamber and is therefore a non-linear element.

The outside diaphragm suspension is another example of a variable-circuit element in an acoustic system. In certain types, or as a matter of fact, for unlimited amplitudes in all types of suspension systems, the stiffness is not a constant, but a function of the amplitude and in general increases for the larger amplitudes.

In the case of a horn loudspeaker the velocity of the diaphragm for constant sound output is independent of the frequency. Under the same conditions the amplitude is inversely proportional to the frequency. Consequently, the greatest distortion due to the suspension system will occur at the low-frequency end of the working range.

In many suspension designs, including paper, fiber and metal diaphragms, considerable distortion occurs at the lower frequencies. To test for this type of distortion a large throat and air chamber should

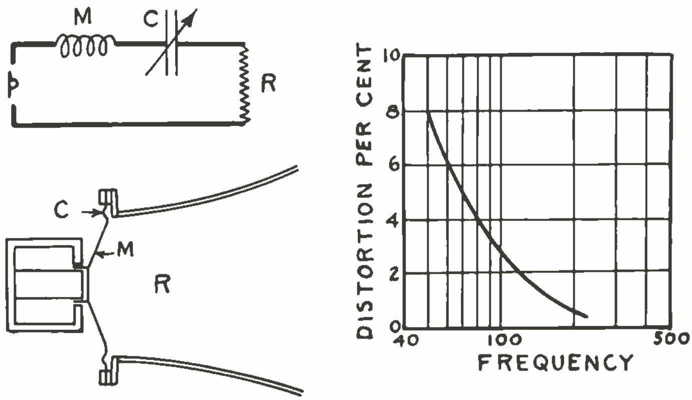

Fig. 8—Mechanism having a diaphragm with a non-linear suspension system. Equivalent electrical circuit of the vibrating system indicates the effect of the non-linear element. Graph shows a typical distortion characteristic obtained on an 8″ diaphragm feeding 5 watts to a large throat horn. A large throat horn is used to minimize distortion due to the air chamber and non-linearity of the air.

be used to reduce the error due to non-linearity of the air and variation in the capacitance of the air chamber. The system and method is shown in Figure 8. The amount of distortion introduced by the suspension system when it is coupled in the normal manner to a smaller throat can be computed from the results obtained on the larger throat. The results in Figure 8 show that considerable distortion may occur if an improperly designed suspension is used. Of course, these values will be relatively reduced when referred to the smaller throat.

Inhomogeneity of the flux density through which the voice coil moves is another source of distortion. The result is that the driving force does not correspond to the voltage developed by the generator in the electrical driving system. Furthermore, the motional impedance is a function of the amplitude. This type of distortion can be eliminated by making an air gap of a sufficient axial length so that the voice coil remains at all times in a uniform field. This type of distortion can also be eliminated by making the voice coil longer than the air gap so that the summation of the products of each turn and the flux density is a constant.

The distortions referred to above have been concerned with higher harmonics; that is, multiples of the fundamental. It has been analytically shown by Pederson[16] that subharmonics are possible in certain

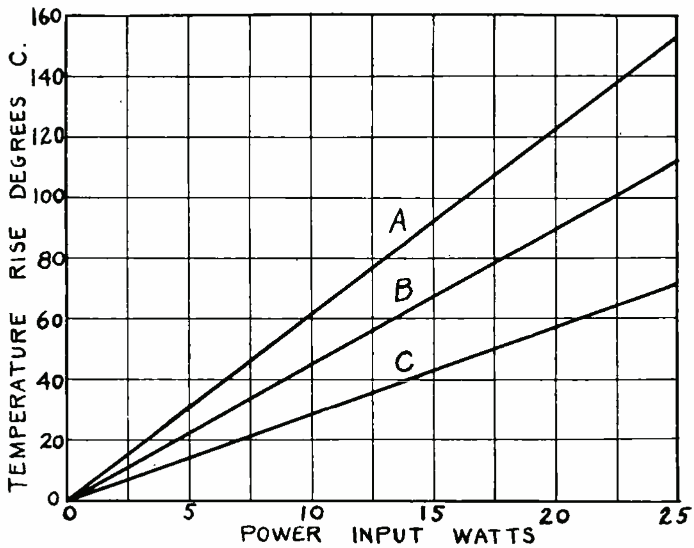

Fig. 9—The temperature rise as a function of the power delivered to a voice coil for air gap clearances, as follows: A, .021". B, .015". C, .009". Coil 1½" diameter and 0.25" length.

vibrating systems. The existence of subharmonics in direct radiator loudspeakers is quite well known. However, in horn loudspeakers the diaphragms are relatively small and quite rigid. Consequently the conditions for the production of subharmonics are not particularly favorable. Nevertheless in certain horn loudspeakers subharmonics can be produced when the diaphragms are driven with large inputs. It has been noticed that by impressing a steady tone upon a system which produces both subharmonics and higher harmonics that the subharmonics are more pronounced and objectionable to the ear than the higher harmonics. However, by actual measurement under these conditions the subharmonic was less than one per cent, while the higher

Pederson, P. O, *Jour. Acous. Soc. Amer.*, Vol. 6, p. 227, 1935.

harmonics were several per cent of the fundamental. The explanation appears to be that it is more difficult to mask a low tone with a high tone than the reverse procedure. Another feature of subharmonic phenomena is the relatively long time required for "build up". Conventional sound reproduction does not usually require the reproduction of a single isolated high frequency tone of long duration. Furthermore, as pointed out above, with relatively small rigid diaphragms and large resistive loads the production of subharmonics is quite small. Therefore, at the present time it seems that subharmonic distortions in horn loudspeakers are not as troublesome nor as important a problem as the other types discussed above.

The maximum allowable distortion may determine the power rating for the loudspeaker. However, in certain loudspeakers the maximum allowable temperature of the voice coil determines the power rating. This is particularly true of high frequency horn loudspeakers.

By making the efficiency a maximum, the dissipation in, and the resulting temperature of, the voice coil for a certain acoustic output will be a minimum. Practically all the heat energy developed in the voice coil is transmitted across the thin air film between the voice coil and the pole pieces and from the pole pieces to the field structure and thence into the surrounding air. In this heat circuit practically all the drop in temperature occurs in the thin air film. The temperature of the voice coil approaches the temperature of the pole pieces as the thickness of the air film is decreased. The temperature rise as a function of the power dissipated in the voice coil for various clearances between the voice coil and pole pieces is shown in Figure 9. These results are obtained for no motion of the voice coil. When motion occurs, the thermal impedance of the air film is reduced and the temperature of the voice coil is diminished.

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MICHAEL RETTINGER graduated from the University of California with degrees of B.A. and M.A. in Physics. It was there that he developed, with the guidance of Professor Dr. Vern Knudsen, the theories of sound-absorption and sound-transmission of porous, non-porous, flexible and non-flexible materials, which theories appeared in print in the *Journal of the Acoustical Society of America.* He joined the engineering staff of the RCA Manufacturing Company at Los Angeles, in October, 1936.

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