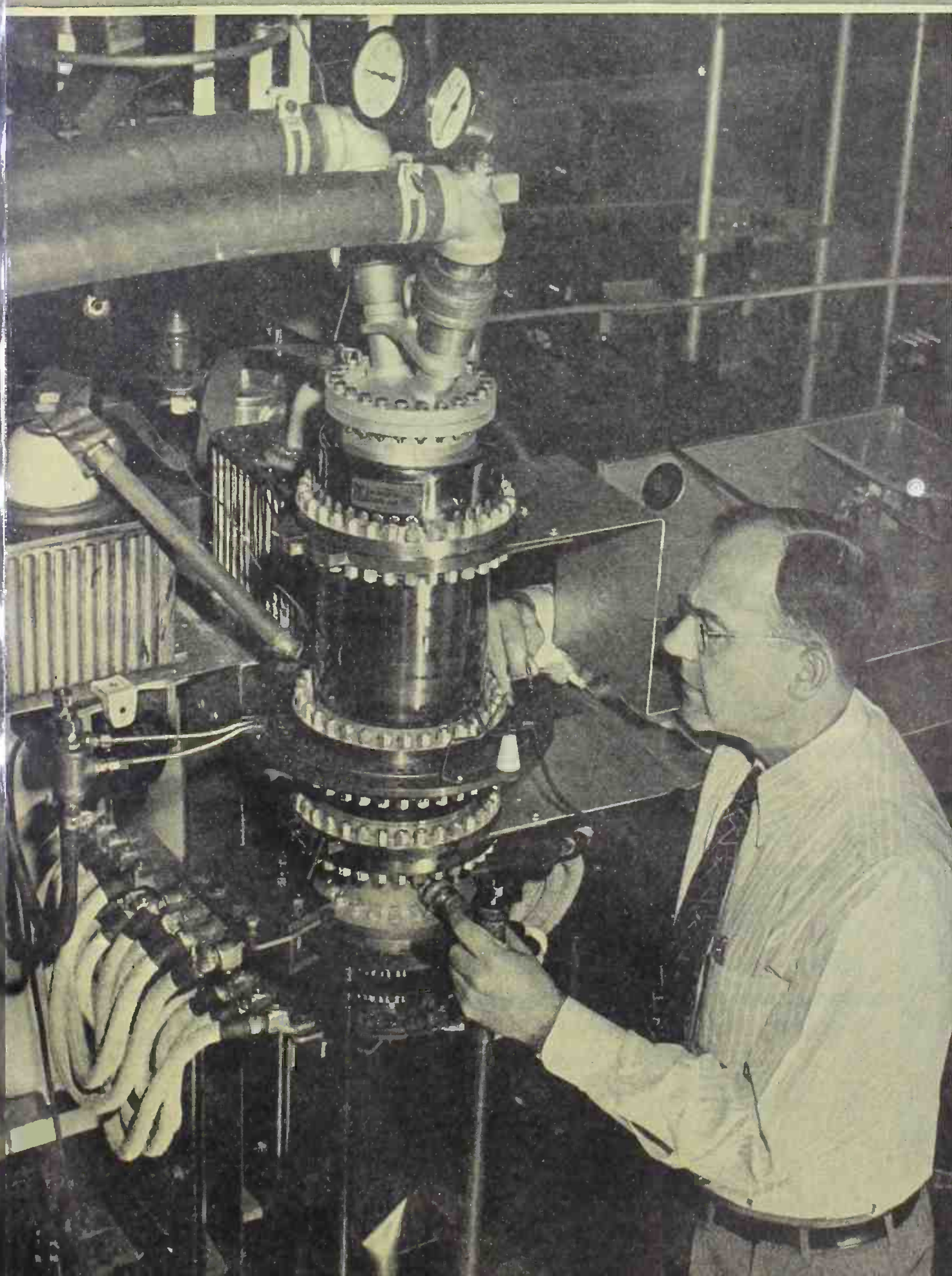
MAY, 1950

RADIO & TELEVISION NEWS

RADIO-ELECTRONIC Engineering



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MAY, 1950

DEPARTMENTS

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COVER PHOTO—Courtesy of RCA

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CUSTOM SOUND INSTALLATION

By ARTHUR SHAPIRO

Consulting Engineer, Chicago

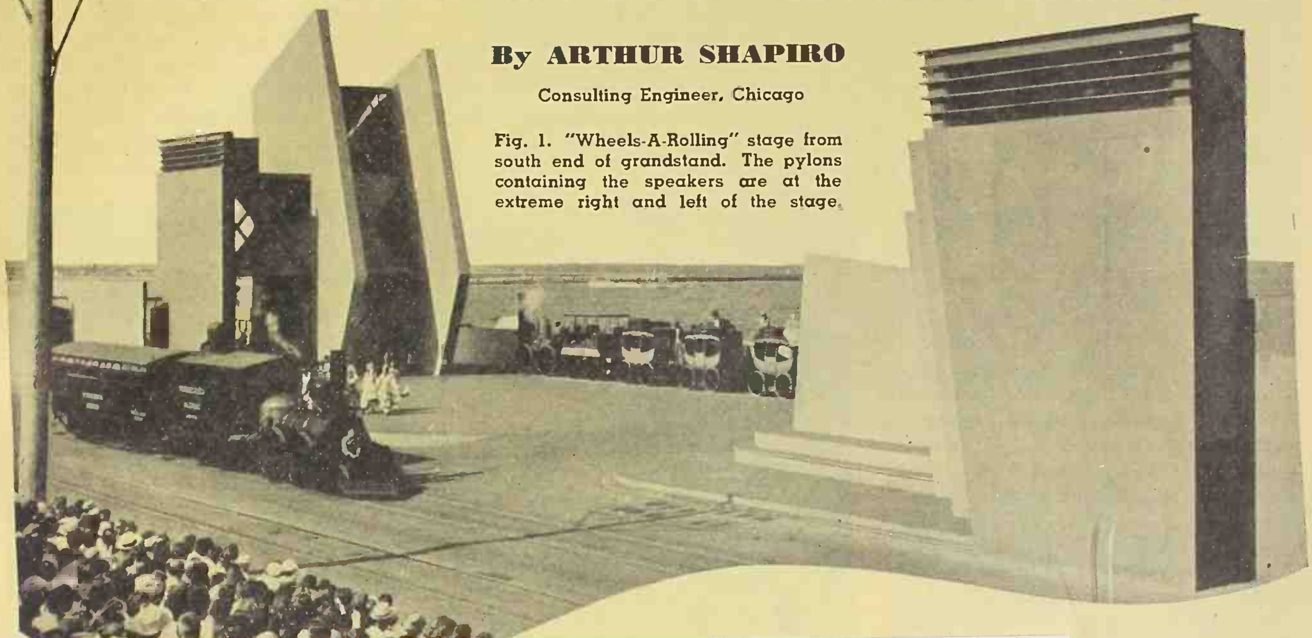

Fig. 1. "Wheels-A-Rolling" stage from south end of grandstand. The pylons containing the speakers are at the extreme right and left of the stage.

...in the pageant ... Railroad Fair.

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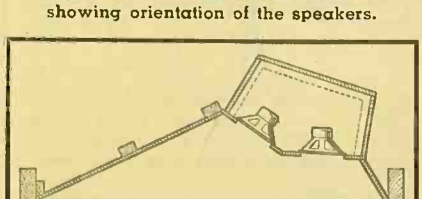
The audience of 6200 that witnessed this spectacle four times daily was housed in a grandstand that measured 318 feet from stage right to left and 80 feet deep starting at ground level at the stage and rising to 30 feet at the rear of the stands. The grandstand had a metal roof extending from end to end and covered approximately two-thirds of the seats. The stage proper was over two hundred feet wide and 100 feet deep with its wings extending

picked up on stage by means of concealed microphones which were plugged in under cover of stage action.

The pageant was scheduled for presentation four times daily for the one-hundred days from June 25 to October

Fig. 2. Cross-section of the baffles, showing orientation of the speakers.



...and trouble-free equipment was a definite requirement. Such items as weatherproof "Twist-Lock" microphone connectors, instant metering of circuits, duplicate wiring and stand-by units were all provided to obviate the possibility of breakdown of the equipment. Since the stage was located on the lake front, the weatherproofing of the equipment was a definite factor to be considered. The high humidity made it necessary to take special care that no space was imperfectly sealed in the equipment that would allow the formation of condensation. On open equipment, sufficient ventilation was required to prevent condensation even when the humidity hovered

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CUSTOM SOUND INSTALLATION

By ARTHUR SHAPIRO

Consulting Engineer, Chicago

Fig. 1. "Wheels-A-Rolling" stage from south end of grandstand. The pylons containing the speakers are at the extreme right and left of the stage.

Design of sound equipment used in the pageant "Wheels-A-Rolling" at Chicago's Railroad Fair.

IN designing the sound installation for the great pageant, "Wheels-A-Rolling," presented at the Chicago Railroad Fair of 1948-49, many unusual conditions had to be met. The tremendous scope of this show can be realized by the great numbers of people and equipment placed on the stage during the one hour and eight minutes of pageantry that told the complete story of transportation in America from early Indian times to the present day. To present this show required a cast of over one hundred actors, 40 engineers, 25 stagehands, stablehands, orchestra, singers and production staff—a total of 250 people who, together with the 94 horses, 35 locomotives and their trains, 32 old automobiles and 100 pieces of horse-drawn equipment, were all controlled from a central point by means of sound, intercommunication, buzzer, light and railroad signal equipment which had to be designed especially for this show.

The audience of 6200 that witnessed this spectacle four times daily was housed in a grandstand that measured 318 feet from stage right to left and 80 feet deep starting at ground level at the stage and rising to 30 feet at the rear of the stands. The grandstand had a metal roof extending from end to end and covered approximately two-thirds of the seats. The stage proper was over two hundred feet wide and 100 feet deep with its wings extending either end to make a total play area 500 feet long.

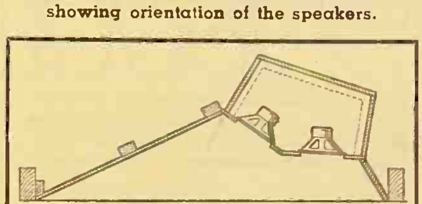
The pageant was historical in tenor with variety and comedy that made for a complete and enjoyable show. Music, especially arranged by Isaac VanGrove, was played throughout the show by an orchestra housed in a special band shell at the south end of the grandstand. The offstage singers were also in the orchestra shell as was the control keyboard for the *Deagan* Celesta-Chime, an electronically amplified chime which will be noted later.

Two narrators, located in a booth at the top of the grandstand, told the story of transportation and gave descriptions and comments of the events as they were enacted on stage. Actors' speaking parts and group singing were picked up on stage by means of concealed microphones which were plugged in under cover of stage action.

The pageant was scheduled for presentation four times daily for the one-hundred days from June 25 to October 2nd with performances at 2, 4, 7, and 9 P.M. From this schedule it can be seen that the equipment would be in almost constant service from early afternoon through ten at night. Before the actual show started, there was recorded music to fill the time while the audience found their places. This music was presented for one hour before the 2 and 7 o'clock performances and one-half hour before the other shows. In case of power failure on the grounds, there was a gasoline powered generator as a stand-by to operate the sound system only, so that the show could be presented during the day and to keep the sound operating to avoid panic in the event of power failure during the night shows.

Heavy-duty and trouble-free equipment was a definite requirement. Such items as weatherproof "Twist-Lock" microphone connectors, instant metering of circuits, duplicate wiring and stand-by units were all provided to obviate the possibility of breakdown of the equipment. Since the stage was located on the lake front, the weatherproofing of the equipment was a definite factor to be considered. The high humidity made it necessary to take special care that no space was imperfectly sealed in the equipment that would allow the formation of condensation. On open equipment, sufficient ventilation was required to prevent condensation even when the humidity hovered

Fig. 2. Cross-section of the baffles, showing orientation of the speakers.



around 98 per-cent. As an indication of the severe humidity, it was not unusual to pick up a tool that had been perfect the night before and find it covered with rust the next morning.

As can be seen from the accompanying drawing, Fig. 7, eleven microphone receptacles were installed, four in the orchestra, one in the room below the orchestra (used for the Celesta-Chime), four on stage and two in the narrators' booth. These receptacles were of the heavy-duty, three wire "Twist Lock" type mounted in weatherproof condulets with spring loaded lift covers. *Shure* Model 556 Super-Cardioid microphones, set at 250 ohms, were used in all locations except at stage right where a lapel microphone was used in front of the speakers by the actor portraying Abraham Lincoln. This microphone, a high-impedance crystal, required a transformer at the point where it was plugged in to feed the 250 ohm balanced input line.

All microphone lines in the stage area were run underground in steel conduit to a junction box set in the center of the stage apron and then through a two inch conduit under the railroad tracks at the front of the stage to the front of the grandstand where they joined the lines from the orchestra pit. The eleven microphone lines were then brought up to the operating position near the top of the grandstand to a 24 jack patch panel. This panel terminated all the microphone lines and normalled them through to the ten console inputs. The patch panel was installed to permit the instantaneous substitution of an unused console preamp in case of failure of one in use. Inserting a patch cord in the top, or microphone, row of the panel automatically disconnected both sides of the line and transferred the incoming signal to the plug which could then be inserted into the bottom, or console input, row. The eleventh microphone line terminated in a fixed "H" pad. This line was used for the electronically amplified Celesta-Chime. The chime preamplifier fed the signal to the line at 500 ohms and zero level. The front side of the pad fed the signal at 500 ohms direct to either of the loudspeaker systems described below, and the back side of the pad fed through a patch cord to one of the console inputs at 250 ohms and -55 db. Normally the patch panel operated with only one cord in use as all other circuits were normalled through.

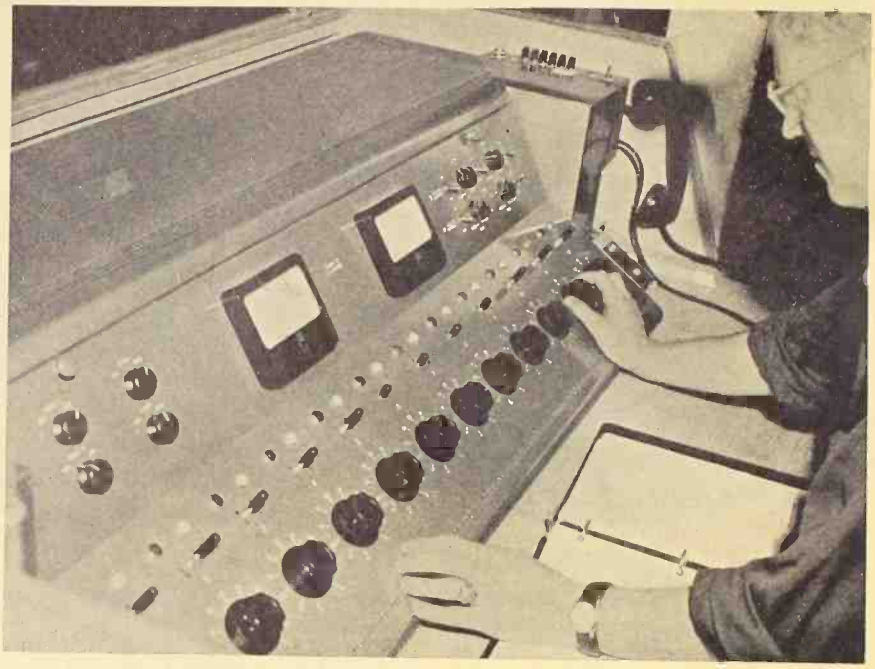
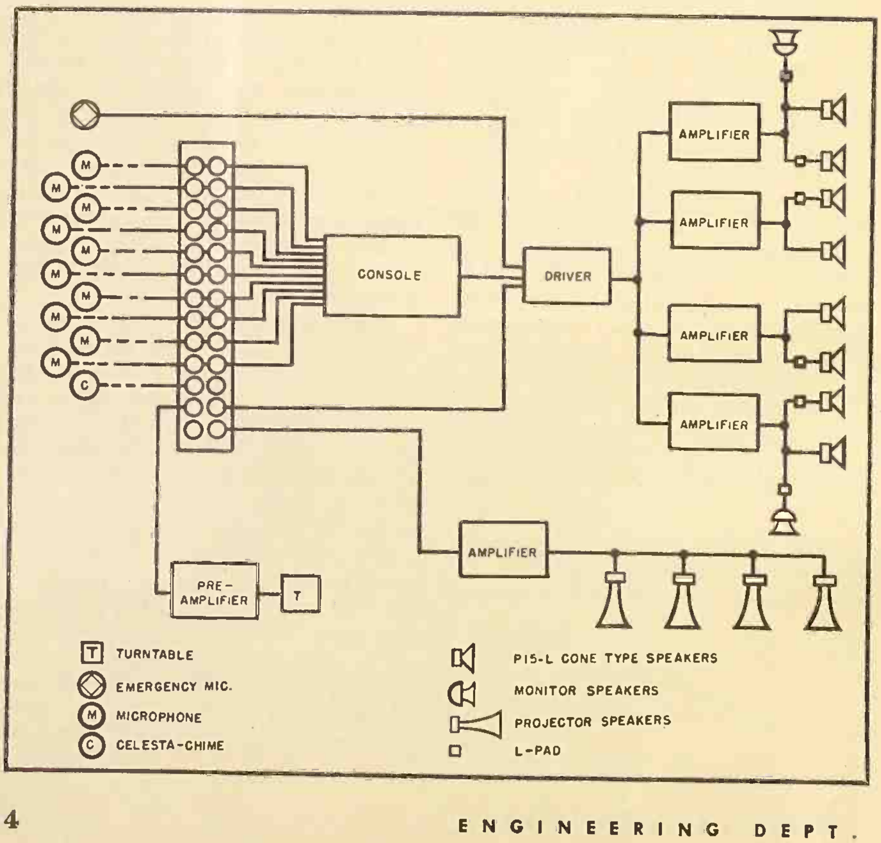


Fig. 3. Ten position, dual channel custom-built control panel in operation. One of the telephone-type intercom masters is visible at the extreme right.

Fig. 4. Block diagram showing essential components of the complete system.



The custom-built console, Fig. 3, contained ten independent preamplifiers feeding through a fader and switch to either of two outgoing channels. Each channel featured independent VU meters, *UTC* tone equalizers and amplifiers, master faders and output switches. Each preamp position had red and white indicator lights to show which channel was being fed. In operation, the "Red" channel was used for music and the "White" channel was used for speech. This enabled the operator to control the fading of several microphones with a single control. In all there were 37 tubes in the console, three in each preamp and the balance in the equalizer and output section. To eliminate any possibility of hum pickup, all filaments and also all indicator lights were operated on 12 volts d.c. The power supply, mounted within the special weatherproof cabinet, consisted of a heavy duty selenium rectifier for the low voltage and a 5U4G for the plate voltage rectifier. The output of the console was fed through the patch panel to the line amplifier-driver located in the main amplifier racks in the director's booth. This driver amplifier had provisions for additional high-impedance microphone inputs as well as the 500 ohm low level input. One of these high-impedance inputs was utilized for the director's break-in microphone for use in case of serious accident or emergency on stage. This microphone could break into the system regardless of the status of any other microphones.

The output of the driver amplifier fed four 80 watt booster amplifiers which were limited to approximately 50 watts output each. These boosters were mounted in one rack and were plug connected so that, in case of a failure, a spare could be quickly interchanged. The need for instant changeover was not deemed necessary as each amplifier fed only one-half of the right or left side of the stage, and failure meant only a slight reduction of volume and would not interrupt the show.

The loudspeakers, by far the most important pieces of equipment in the installation, consisted of eight *Jensen* P15-L cone type speakers specially weatherproofed and treated. These speakers were mounted in special baffles, Fig. 6, and connected two to an amplifier. There are several unique features in the design of these baffles and a study of Fig. 2 will show how each individual motor was oriented to a specific coverage area relative to the grandstand. The pylons in which these units were mounted were not equally spaced and were different distances from the grandstand. This required that the speaker coverage areas be accurately plotted.

Due to the comparatively small difference in angle of each pair of speakers, it was decided that two speakers would be mounted in a common horn and rear enclosure. The loudspeakers had a downward tilt of approximately 10 degrees and the "floor" of the baffles sloped 15 degrees. This angle placed the aural center of the unit at the middle row of the grandstand. These horns were made of three-quarter inch marine grade plywood and finished with flat black paint on the inside so that the speakers were not visible from the audience. The tops of the pylons which supported these large baffles were built with an opening sixteen feet long and four feet high faced with one-quarter inch square mesh screening and treated architecturally with horizontal two by six planks as shown in Fig. 1.

It was evident that with the great variation of distance that the sound had to travel to reach various parts of the grandstand, some control was necessary to drop the level of those speakers operating with the shortest "throw". This was accomplished by inserting a heavy 25 watt "L" pad in the line of each of the two inside speakers of each pylon. In the case of the north pylon, where there was a ladder and platform, these controls were mounted on the speaker enclosures. In the south pylon where the speakers were inaccessable, these controls had to be mounted in a weatherproof box remote from the speakers. Once these controls were set by experimentation and measurement after the installation was completed, they were not changed for the rest of the season. Referring to the block diagram, Fig. 4, it can be seen how these controls were installed. The entire grandstand was covered with a variation of not more than 6 db. This was accomplished by sound pressure measurement and setting each booster amplifier separately and then setting the "L" pads on the individual speakers.

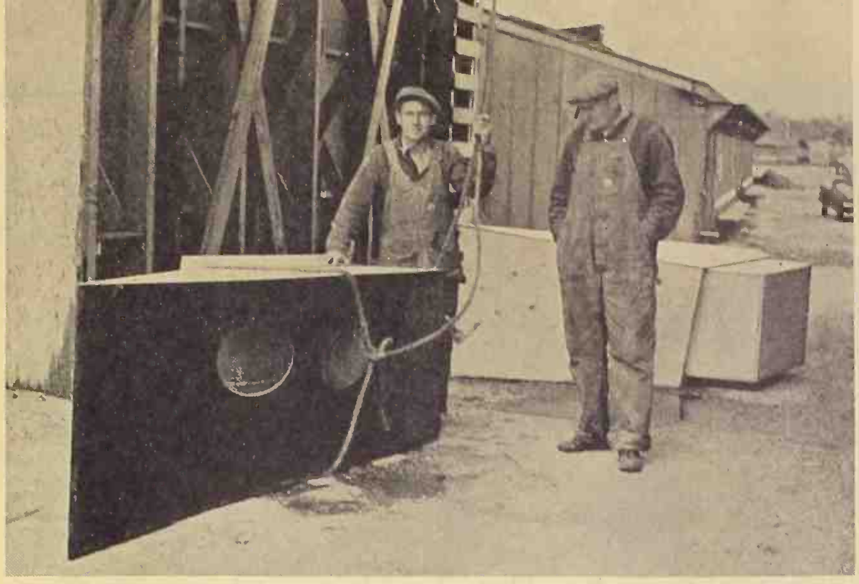


Fig. 5. Two of the special dual loudspeaker housings before being hoisted into place. Flare of horn and mounting of speaker enclosure can be noted on rear horn.

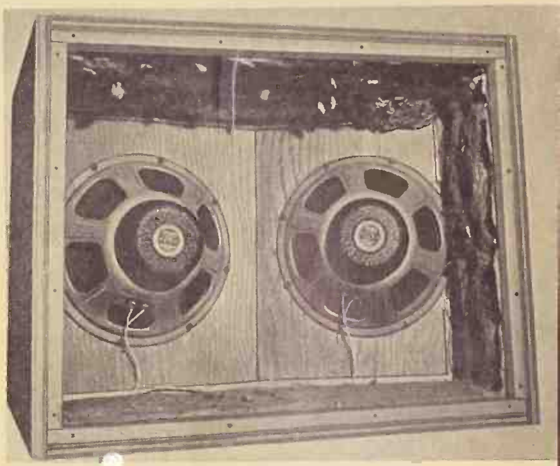
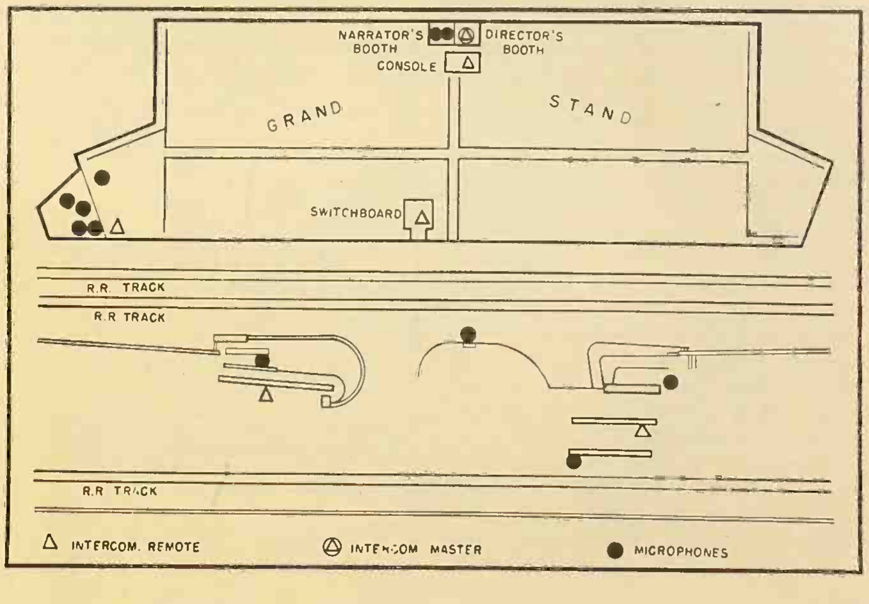


Fig. 6. Dual speaker enclosure showing speakers mounted on individual sections. Note lining of sound absorbent material.

This may seem to be adjusting rather
(*Continued on page* 31)

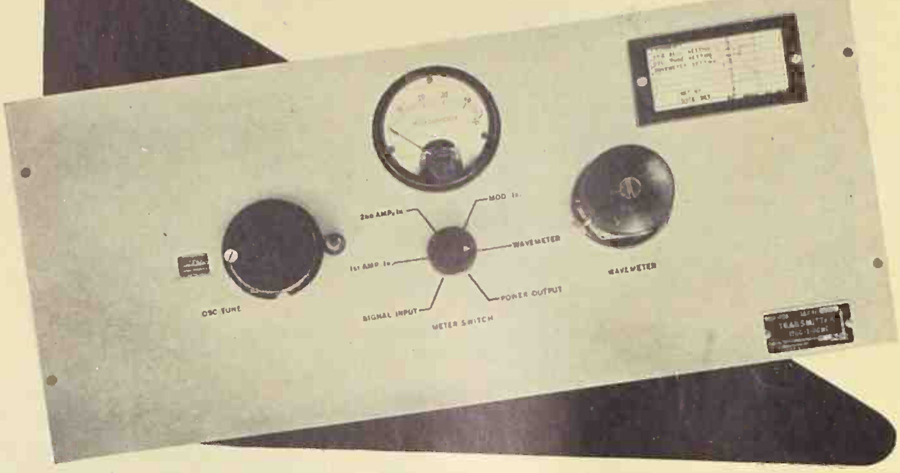
Fig. 7. Drawing showing the location of various components of the system.



MICROWAVE TRANSMITTERS

By J. RACKER

Federal Telecommunication Labs.



Typical microwave transmitter using a lighthouse oscillator.

First of two articles discussing the generation, modulation, and frequency control of microwaves.

THE compactness and simplicity of microwave transmitters is of great interest to all engineers, particularly so to most "low frequency" engineers accustomed to regarding point-to-point transmitters in terms of large size. The main reason for the reduction in size—for equivalent service—is the large power gain obtained by microwave antennas. It is generally true that the higher the frequency, the more expensive and inefficient the generation of power but, due to the increase in antenna gain, less transmitter power is required. It is therefore possible to operate at higher frequencies—all other factors remaining equal—without too much loss in over-all system efficiency, provided directional point-to-point transmission is desired.

Since it is more expensive to operate at microwave frequencies (for a given power output), virtually all transmitters operating in this band employ high level modulation, i.e. one r.f. stage is used with the modulating voltage amplified to a sufficient degree to modulate the final at the desired power output level. Thus in considering the design of a microwave transmitter, we are primarily interested in the final stage—all other stages employing conventional techniques.

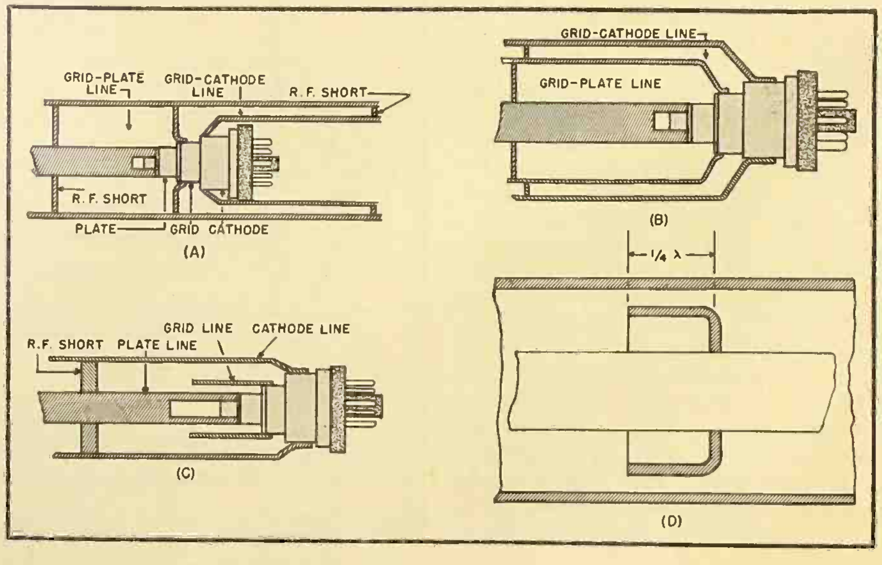
Microwaves can be generated in a number of ways. The most important of these are the lighthouse triode oscillators, klystrons, magnetrons, and traveling wave tubes. With the exception of the lighthouse triodes, the tubes used at these frequencies are as different in concept and design from conventional tubes as amplitude modulation differs from frequency modulation. The important characteristics of these four methods of microwave generators are:

1. *Lighthouse Triodes.* Operating range: up to about 3300 mc.; peak power of the order of kilowatts depending upon frequency and duty cycle; average power of the order of watts. This tube requires an external cavity designed by an engineer. This cavity is permanent and only the tube must be replaced when it burns out (in contrast to other tubes as will be noted), therefore from a long term viewpoint it is the most economical method. Requires relatively elaborate automatic frequency control systems when a high degree of stability is required, can be pulsed, operated c.w. or amplitude modulated. Difficult to frequency modulate.

2. *Klystrons.* Frequency range: 900 to 21,000 mc.; power (peak and average) of the order of 5 watts depending upon frequency; efficiency is poor—less than 10%. Cavity is built right into tube and therefore no special cavity design is necessary unless other design problems are involved. Entire assembly must be replaced if tube goes bad. Can readily be frequency modulated, electronically tuned and frequency controlled (a.f.c.). Requires multiple, well regulated power supplies which are relatively expensive and bulky. Cannot readily be amplitude modulated.

3. *Magnetrons.* Frequency range: 900 to 30,000 mc.; peak power of the order of megawatts depending upon frequency and duty cycle; average power of the order of kilowatts depending on frequency; efficiency is very good—better than 50% depending on frequency. Cavity built right in and entire assembly must be replaced if tube goes bad. Good frequency stability—depends only on expansion of copper. Not easy to

Fig. 1. (A) and (B) Two types of double coaxial resonator circuits. (C) Re-entrant cavity circuit. (D) Quarter-wave choke used as r.f. shorting plug.



a.f.c., amplitude or frequency modulate. Usually used for c.w. or pulsed applications.

4. *Traveling Wave Tubes.* Frequency range: 500 to 10,000 mc. (and higher); power, c.w., is of the order of hundred watts depending upon frequency; efficiency may be as high as 60%. These tubes are particularly noted for their broad bandwidth. At the present time they are largely experimental and show great promise for the future.

This brief summary of the characteristics of the four major methods of generating microwaves indicates the primary considerations confronting the engineer in designing the transmitter. When the power output, carrier frequency, and method of modulation are specified, the choice of the optimum method of generation is usually apparent though other system characteristics such as bandwidth, frequency stability, and equipment size should also be evaluated before the final selection is made.

Transit Time Effect

It is known that a finite period of time is required for an electron to travel from the cathode to the plate of a tube. This period of time is called the transit time and at low frequency ranges it is negligible compared to time required to evolve one cycle. However, starting at about 500 mc., the time required for one cycle of operation is less than 0.002 microseconds which is comparable to the transit time in most conventional tubes.

This fact invalidates many of the characteristics and design principles governing these tubes since the assumption had been made that the electron stream flowing from cathode to grid traveled under essentially static conditions, i.e., transit time small compared to time variation of tube potentials. However, when the transit time becomes comparable to the period time, then the tube conductance is increased and tube efficiency and output decreased.

This can be best understood by considering the action in a triode with an alternating current applied to its grid. When the grid is at zero (a.c.) potential, the electrons travel from cathode to plate at the quiescent velocity. With the grid at a positive a.c. potential, the electrons are accelerated and pick up energy from the grid in the cathode-to-grid path and are retarded and lose energy in the grid-to-plate path. The reverse occurs when the grid is at a negative a.c. potential so that there is no net energy transfer from the a.c. grid potential source to the electrons.

At frequencies where the period is comparable to the transit time, it is possible for the a.c. grid potential to

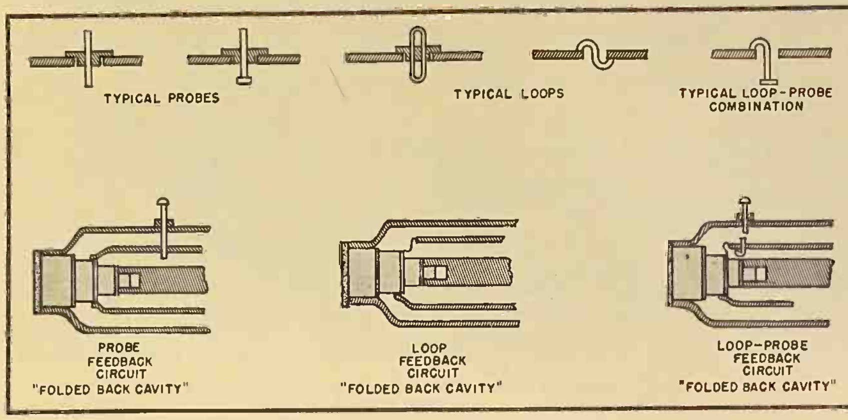


Fig. 2. Typical loops, probes, feedback circuits, and combinations.

reverse its phase before the electrons reach the plate so that, in the positive grid case, the electrons are accelerated and gain energy in both the cathode-to-grid and grid-to-plate regions. Under these conditions, there is a net energy transfer from a.c. grid potential to electrons, resulting in an increase in input conductance. In a similar manner it can be shown that energy can be transferred from the source of a.c. plate potential to the electrons during a portion of the cycle resulting in a decrease in power output and efficiency.

Another factor influencing the operation of tubes at microwave frequencies is the lead inductance of the tube elements (plate and cathode) and the tube interelectrode capacitance. In many cases the resonant frequency of these *L-C* networks falls below or within the range of operation. Hence, microwave tubes must be designed with a minimum of lead inductance and interelectrode capacitance.

The requirements of reducing transit-

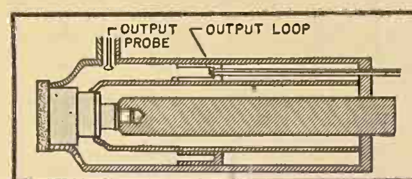


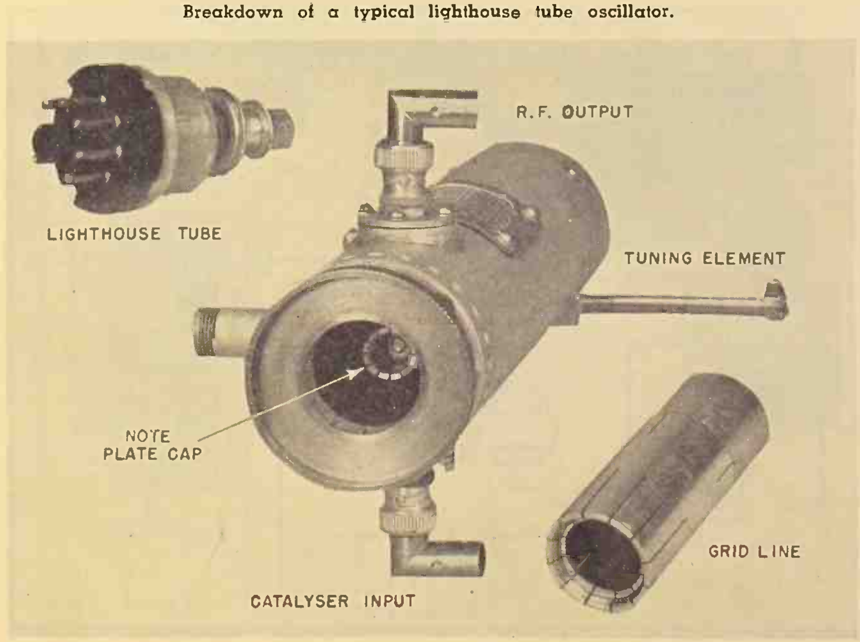
Fig. 3. Alternate methods of output coupling—probe or loop circuit.

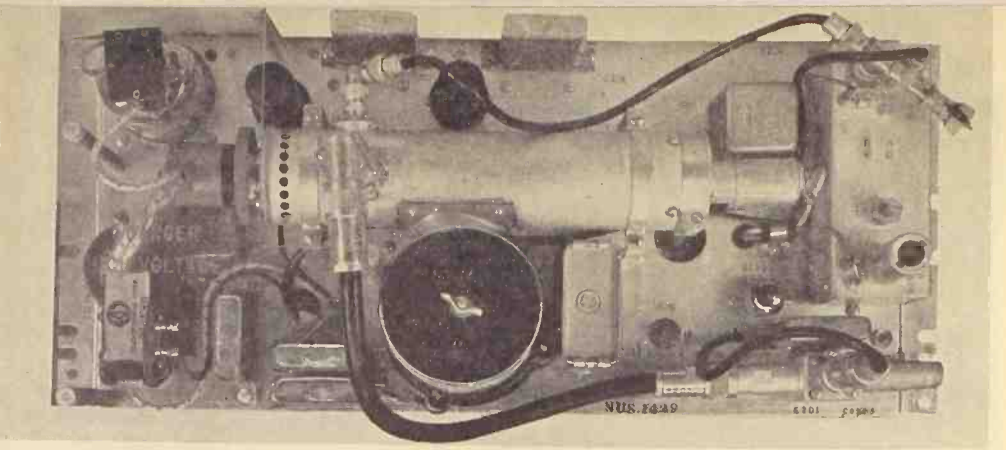
time and interelectrode capacitance are conflicting, necessitating design compromises. Interelectrode capacitance may be reduced by decreasing the physical size of electrodes and increasing the spacing.

However, increasing the spacing increases the transit time, while reducing element size decreases maximum safe power dissipation. Transit time may be reduced by increasing the plate potential, but increased plate potential means greater power dissipation.

It is therefore readily seen that special tubes must be used for microwave frequencies which overcome the limita-

Breakdown of a typical lighthouse tube oscillator.





Microwave transmitter—rear view—showing lighthouse oscillator.

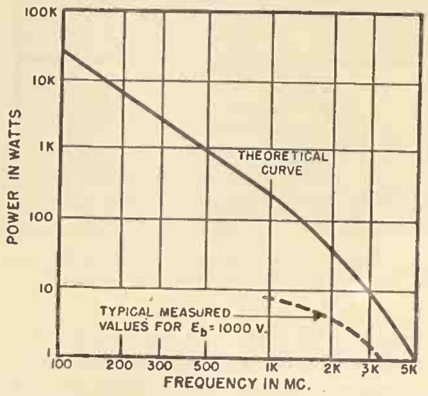


Fig. 4. Theoretical and typical power output versus frequency for a simple triode oscillator.

tion of transit time and capacitance noted above or employ entirely different techniques than conventional tubes. Lighthouse tubes fall in the former category, while klystrons, magnetrons, and traveling wave tubes are examples of the latter category.

Lighthouse tubes are designed for operation in the frequency range of 200 to 3370 megacycles in which coaxial line resonators and elements are used. A typical lighthouse tube is shown on page 7 and as indicated in this picture it has the over-all appearance of a lighthouse—hence its name. These tubes are constructed in this way to permit the use of coaxial elements which fit right over the cathode, grid and plate leads.

This tube has effected reduction in lead inductances by use of disk seals, and employs a parallel plane structure which permits the use of relatively small electrode areas; thereby making possible low interelectrode capacitances while reducing the spacing between electrodes to the point where the transit time is small in comparison with the period at the frequencies indicated.

Design of Lighthouse Oscillators

There are two types of resonator circuits used in conjunction with the lighthouse tube, one is known as the double coaxial resonator, shown in Figs. 1-A and B, and the other a re-entrant cavity, shown in Fig. 1-C. As indicated in Figs. 1-A and B, there are two variations of the double coaxial resonator. In one case, the grid to cathode line is completely independent from the grid to plate line. In the second case, in order to conserve space, the grid to cathode line is folded back over the grid to plate line. This does not alter the fundamental principles of the two resonant circuits inasmuch as the inner surface of the grid cylinder acts as the grid side of the grid-anode line while the outer surface forms the grid side of the grid-cathode line. There is no coupling between these two lines through the common conductor since, as indicated in a previous article[1], the depth of penetration of the electromagnetic field at these frequencies is much less than the thickness of the metal.

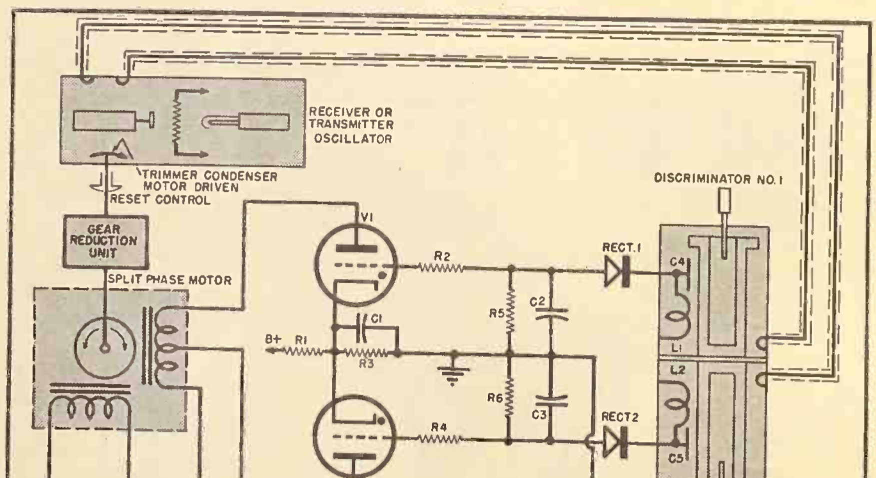
The shorting plugs, shown in Figs. 1-A, B, and C, are actually r.f. shorts and not d.c. shorts. This is done by using a quarter-wave "choke" shown in Fig. 1-D. Since this element is a quarter wave long and open circuited, it presents a short circuit to the r.f. energy. The spacing between the choke and the wall of the tube is determined by the amount of d.c. voltage applied to the plate of the tube. The smaller the spacing the more effective the "choking" or "shorting" action, since Z_o will be small.

In the double coax circuits shown in Figs. 1-A and B, both grid-cathode and plate-grid lines have shorting plugs which can be varied to tune over the desired frequency range. The re-entrant cavity circuit, shown in Fig. 1-C, is similar to the "folded back" grid-cathode line. However, this circuit is not tuned in the same manner that is used in the double coax resonator. Instead of terminating the tuned circuits in "shorting plugs," the grid line is made one half wavelength long and mounted directly on the flange of the tube. The length of the grid line is fixed for any given setup, and thus becomes the frequency determining factor in the circuit.

The shorting plug in the re-entrant cavity, which is mounted on the plate line, does not affect the resonant frequency of the circuit, but functions primarily to optimize conditions for oscillation with a given grid cylinder rather than to change the frequency.

The double coax and re-entrant cavity circuits are basically degenerative and the amount of feedback provided by the interelectrode capacity is insufficient for oscillation, if care is used in shielding the resonant lines from one another. Probably the most frequent cause of insufficient shielding is poor grid contact design. These lines should be slotted to permit a spring contact. Similarly

Fig. 5. Typical automatic frequency control circuit for a microwave oscillator.



the plate line should be machined and slotted so that it grips the plate cap of the lighthouse tube firmly.

There are a number of feedback loop or probe arrangements that can be used to sustain oscillations. Some of these are shown in Fig. 2. A loop is usually placed at a voltage minimum and a probe at a voltage maximum. When broad tuning is desired, a combination of probe and loop feedback elements may be used to cover the entire frequency range.

It will be found that the frequency and efficiency of the oscillator will depend to some degree upon the position (determining phase of feedback) and size (determining the amplitude of feedback) of the probe or loop. A trial and error procedure is used to determine the optimum feedback circuit that provides maximum output and efficiency.

The output of the oscillator is picked up by either a probe or loop as shown in Fig. 3. The exact position of the probe or loop will depend upon the impedance of the load. Again the best procedure is to vary the position until maximum output is obtained across the load. This effect is similar to tapping the coil of a tank circuit to match a given impedance.

The diameters of the lines are determined by the requirements of the oscillator. Using standard theory, the desired Q of the tank circuit is determined considering both the effects of the tube and the load. The Q of a resonant length of coaxial line was given[3] in a previous article as:

$$Q = \frac{2\pi f_o \sqrt{LC}}{2\alpha_T} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

This equation can be simplified by assuming that the attenuation due to the dielectric, α_d, is negligible, in which case $\alpha_T = R_f/2Z_o$, and Eqt. (1) becomes:

$$Q = \frac{2\pi f_o Z_o}{R_f V_T} \quad . \quad . \quad . \quad . \quad . \quad (2)$$

since $V_T = 1/\sqrt{LC} = 3 \times 10^{10}$ cm./sec. for air. Further simplification of this equation can be obtained[3] by writing Z_o in terms of D/d and the r.f. resistance[2] in terms of D, d, and f. Doing this we obtain:

$$Q = 8.39\sqrt{f}\, DHR_1 \times 10^{-2} \quad . \quad . \quad . \quad (3)$$

where D is in cm., f in mc., R_1 is the ratio of the d.c. resistance of the conductor used to the d.c. resistance of copper, and H is defined as:

$$H = 3.6 \left(\frac{D+d}{d}\right) \log_e \frac{D}{d} \quad . \quad . \quad . \quad (4)$$

Eqt. (4) is plotted in Fig. 7. With the help of this figure and Eqt. (3), the Q of a coaxial line for a given D/d ratio can readily be determined.

For the purpose of mechanical simplicity, the plate line diameter is usually made so that it will just slip over the plate cap. The diameters of the grid and cathode rods are then calculated based on this parameter and the Q required. Of course it is desirable to use rods that are of the same diameters as the caps provided or slightly larger.

Earlier in this article it was noted that there is a definite relationship between interelectrode capacitance, electron transit time, and maximum plate dissipation in a tube design, i.e., reducing interelectrode capacitance and transit time also causes a reduction in maximum plate dissipation. Due to the various mechanical factors involved, there is a maximum theoretical (at least at the present time) continuous power output than can be obtained from a single triode employed in the conventional manner. This maximum value has been plotted in Fig. 4 as a function of frequency.

The continuous power output obtained from commercially available tubes is less than this theoretical value. Fig. 4 plots a typical power output versus frequency curve of a 2C43 lighthouse tube operating in a re-entrant cavity. The outputs obtained from other tubes operating with different potentials can be obtained from the tube manufacturers.

At microwaves, where considerable spectrum is available, it is frequently convenient to use greater r.f. bandwidth for greater efficiency and power. This can be done by using one of the several pulse modulation techniques available such as pulse time modulation (PTM), pulse amplitude modulation (PAM), or pulse frequency modulation (PFM)[4]. When the transmitter is pulsed, a much higher peak power can be emitted for a given maximum average power.

It is known that the average value of a pulse series is equal to[5]:

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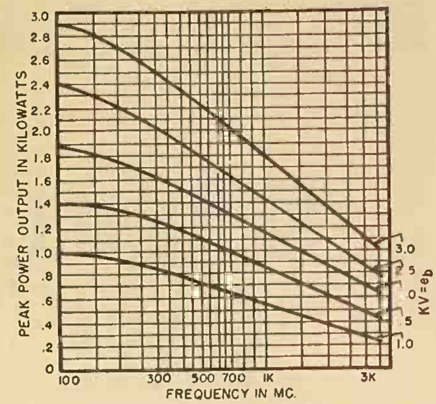


Fig. 6. Plate-pulsed oscillator performance, power output vs. frequency. Pulse rate, 1000 per second; duration, 1 microsecond.

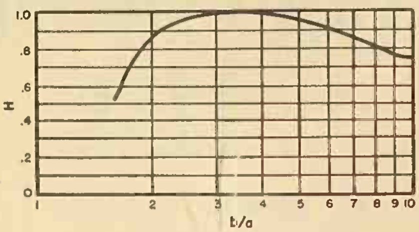
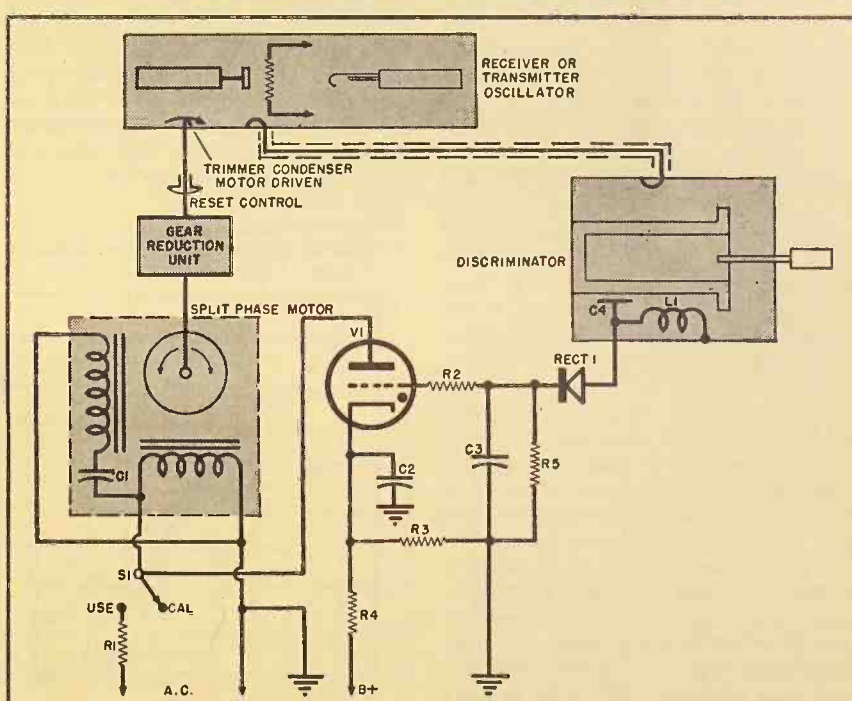
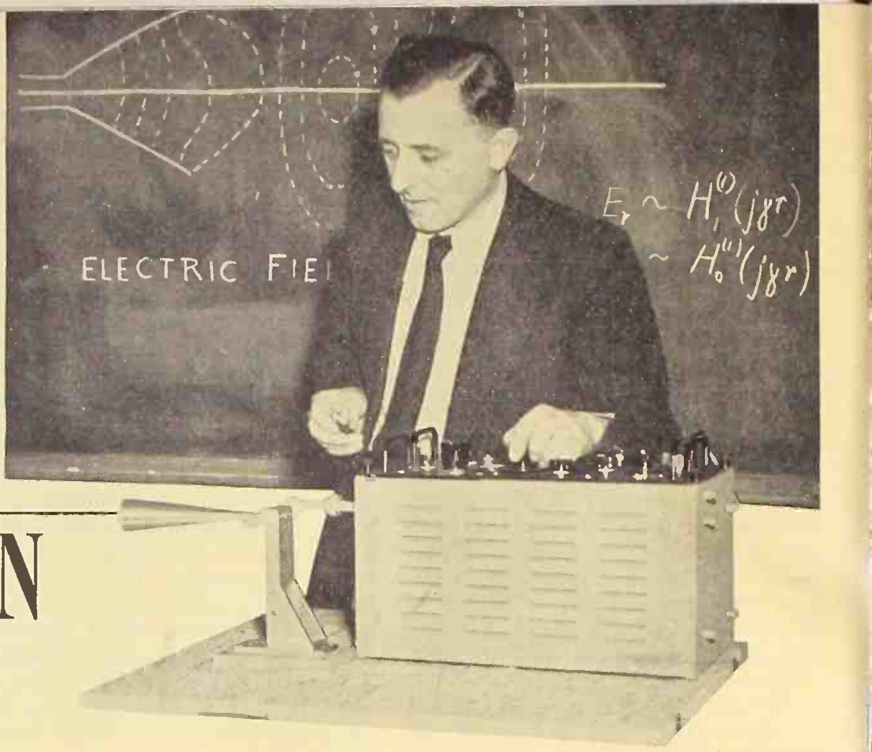


Fig. 7. Variation of H with b/a.

Fig. 8. Another type of microwave automatic frequency control circuit.





Dr. Goubau at the Signal Corps Engineering Laboratories with a model of the simplified "G-String" transmission line which he invented.

SURFACE WAVE TRANSMISSION LINE

By GEORG GOUBAU
Signal Corps Engineering Labs.

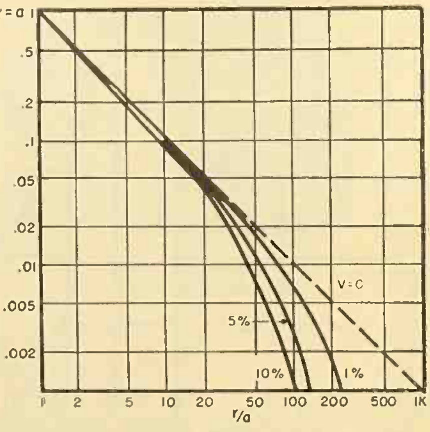
A single insulated wire with special funnel-shaped terminals provides very efficient h.f. transmission.

A PAPER published by Sommerfeld in the year 1899 describes a surface wave guided by a cylindrical conductor of finite conductivity. This wave differs from the wave mode usually excited on long wire antennas in two respects: first, it suffers no radiation loss along its path and second, the field distribution depends to a high degree upon the conductivity of the wire material; and the wave would not exist at all if the conductivity were infinite. Though there is a solution of Maxwell's equations in that case, the solution lacks physical reality since the power transmitted by the wave is infinite if the current in the wire is finite.

Sommerfeld's paper has seen little consideration in the modern literature and there are only a few books in which it is quoted. One of these books is Stratton's well-known "Electromagnetic Theory". It may have been doubtful to many a physicist whether this surface wave can be excited, because not every special solution of Maxwell's equations is realizable. Sommerfeld's wave is a plane wave and therefore the wave could be part of an asymptotic solution, which is valid at a large distance from the power source. In that case, only an infinitesimally small amount of the total power would be converted into the surface wave. To our knowledge nobody has proven as yet that Sommerfeld's surface wave can be excited with finite amplitude. Though we are convinced that this can be done, there are two facts which will make it difficult to obtain good efficiency. The field extends very far from the wire and has a rather complicated shape. An effective launching device would have to be very large, otherwise the radiating wave mode would become predominant.

The reason we were interested in Sommerfeld's wave for a long time is the theoretically low attenuation of this wave. Transmission lines using this wave mode should have less insertion loss than coaxial cables and—provided the diameter of the wire is sufficiently large—even smaller loss than rigid wave guides. This fact made us think about the possibility of concentrating the field closer to the wire; thus, the excitation of the wave would become simpler and besides, the clearance around the wire required for an undisturbed propagation could be kept smaller. A study of cylindrical surface waves showed that the extension of the field of a surface wave depends upon the phase velocity. The more the wave is urged to travel with lower speed, the more the diameter of the field shrinks. A reduction in phase velocity of less than one per-cent is in general sufficient for obtaining reasonable dimensions of the field. Such a reduction can be achieved by a dielectric coat on the wire or by other modifications of the conductor surface like the application of a thread on the wire. The wave mode guided by a conductor, the surface of which is modified as mentioned, differs from Sommerfeld's wave not only in the extension but also in the structure of the field. Both wave modes are mathematically described by Hankel functions; however, in the case of the plain wire, the argument of these functions is complex, while it is purely imaginary in the other case, at least if the losses are disregarded. And here we have an important difference: the finite conductivity is no longer an essential provision for the existence of the non-radiating wave mode.

Fig. 1. Distribution of magnetic field for various reductions in phase velocity λ/(wire radius) α is 100.



Last summer we started experiments with this wave mode on dielectric coated and threaded wires. From the very beginning we observed that this wave mode can be easily excited with an efficiency up to 90%. Furthermore, the experiments showed that the wave mode is very stable with regard to a sag of the wire or to small bends. The measured loss of transmission lines based on this wave mode was in very good agreement with the theoretical expectations.

Before discussing the setup and the results we would like to say a little bit more about the wave mode. The wave mode is a radially symmetrical transverse magnetic mode with a small longitudinal electric component. Fig. 2 shows the electric field lines for a case in which the wavelength is 100 times the wire radius and the phase velocity is reduced by 10%. In general a much smaller reduction of phase velocity is used. We assumed here a large reduction in order to get more field lines on the drawing. All electric field lines start on the wire and return to the wire. The magnetic field lines are circles around the wire.

Fig. 1 shows how the field decreases with the distance from the wire. The ratio of the magnetic field strength at a distance from the wire to the magnetic field strength at the surface of the wire is plotted versus the distance, measured in multiples of the wire radius. Both scales are logarithmic. The broken line indicates a $1/r$ decrease as it would be present in the case of a plain wire with infinite conductivity. In this case the phase velocity would be equal to the velocity of light, and as mentioned before the power would be infinite if the field strength were finite. The unbroken curves show how the field decreases if the phase velocity is reduced by 1, 2, 5 and 10%. The curves follow first the $1/r$ decrease and approach at larger distances an exponential decrease. The smaller the phase velocity is, the earlier the exponential decrease begins.

Fig. 3 is a schematic sketch of a surface wave transmission line. The launching device used in the experimental work consisted of a metal horn connected to the outside conductor of the coaxial feed lines. The inner conductor of that line is connected to the surface wave guide consisting in most of the experiments of a normal enameled wire. The transmitting and receiving ends of the line are alike. The dimensions given in Fig. 3 refer to the curves of Figs. 5 and 6.

The loss of such a surface wave transmission line consists of three parts: the conductivity loss in the wire, the loss in the dielectric coat and the loss, effected by the launching device, due to a partial excitation of the radiating mode. The first two parts, conductivity loss and dielectric loss, can be calculated from the field distribution of the wave. They are proportional to the length. The launching loss is independent of the length of the line and can be determined with fair accuracy by the following consideration. At the receiving end that portion of the wave energy will be received which travels within the area of the aperture of the horn. The wave energy outside this area will be lost. The ratio between received energy and the total energy determines the efficiency of the receiving horn. Because of the reciprocity theorem the efficiency of the transmitting horn must be the same as the efficiency of the receiving horn.

The loss of the transmission line shown in Fig. 3 was calculated for various frequencies and Fig. 5 is a plot of the results. The length of the line is 120 ft., the diameter of the wire 0.2 cm., and the thickness of the enamel layer 5.0 x 10^{-3} cm. The dielectric constant is assumed to be 3 and the power factor 8 x 10^{-3}. The opening of the horns is 13″. The conductivity loss L_c and the dielectric loss L_i increase with frequency while the launching loss L_h decreases. The total loss L_{tot} which is the sum of the three components L_c, L_i and L_h has a flat minimum of about 2 db. The loss was measured for three frequencies. For 1600 mc. it measured 2.2 db.; for 3300 mc., 2.3 db.; and for 4700 mc., 4.5 db. While the deviation between theory and measurement for the lower frequency is comparatively small, it is pretty large for the high frequency. The main reason for this deviation lies in the fact that the angle of the horns was too large for the high frequency and thus the efficiency of the horns was much smaller than expected. For comparison, the approximate loss of 120 ft. of RG-8/U cable is 13 db. for 1600 mc., 22 db. for 3300 mc. and 30 db. for 4700 mc. Though these data refer to a very good cable the losses are about 6 to 10 times as high as the losses of the surface wave transmission line.

The effect of the thickness of the dielectric coat on the efficiency of such a transmission line is demonstrated by curves shown in Fig. 6. Length of the line, diameter of the wire and dimensions of the horns are the same. Only the thickness of the dielectric coat is assumed to vary. The curves refer to a frequency of 3000 mc. For a very thin dielectric coat the insertion loss is large because of the low efficiency of the horns. There is an optimum thickness for which the loss is a minimum. For larger thickness of the dielectric coat the loss rises slowly because of the increasing conductivity and dielectric losses. A curve like this has been verified. The same horns and the same wire length (120′), but a wire of 0.26 cm. thickness were used (this was the only available bare copper wire). The measurements were made at a frequency of 2600 mc. The wire was outdoors for

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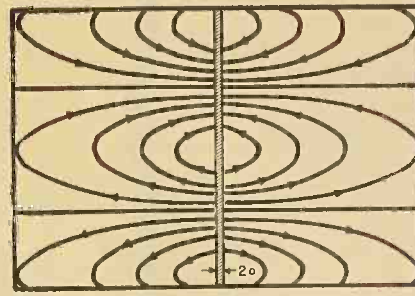


Fig. 2. Electric field lines where $\lambda = 100\ a$ and $v/c = 0.9$

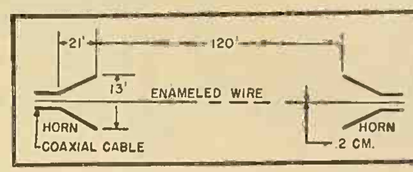


Fig. 3. Schematic sketch of a surface wave transmission line.

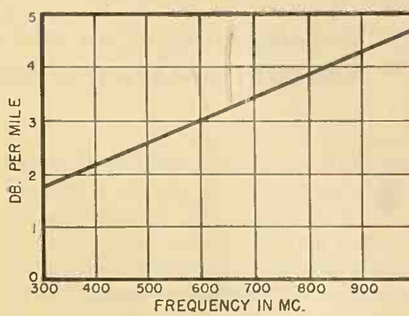


Fig. 4. Calculated attenuation for 2 cm. wire with ½ mm. polystyrene coating.

Fig. 5. Losses for the transmission line of Fig. 3 as a function of frequency.

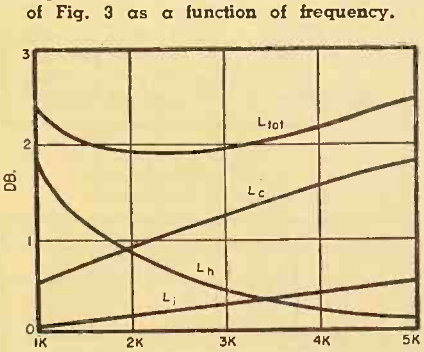
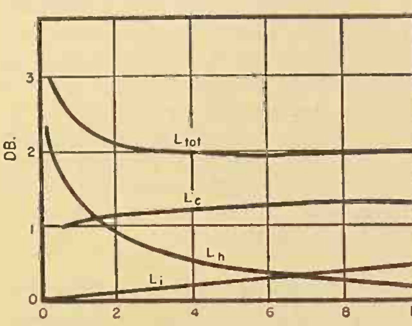
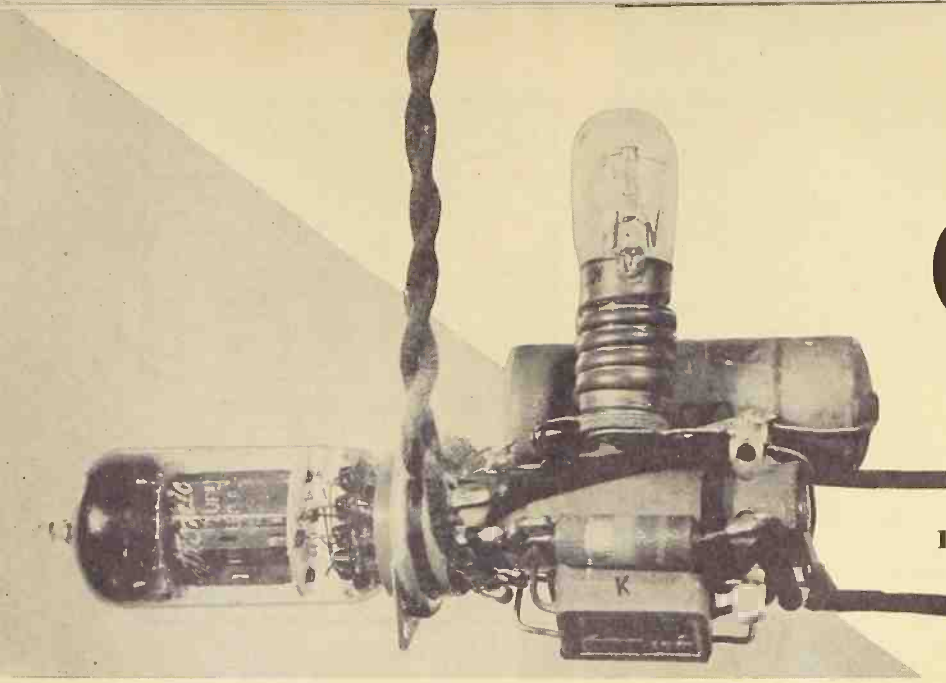


Fig. 6. Losses for the line of Fig. 3 for varying thicknesses of dielectric coating.





A NEW R-C OSCILLATOR CIRCUIT

By **MILTON H. CROTHERS**
Electrical Engineering Dept., U. of Illinois*

Oscillator using a 12AU7 and operating near 9 kc. Note compactness of unit.

Development of an R-C oscillator having greater stability and fewer parts than conventional units.

VACUUM tube oscillators are truly fundamental to the art of modern communications. Vacuum tube oscillators may be grouped in two general classes depending upon the class of operation of the amplifier stage, that is, linear or nonlinear. The first vacuum tube oscillators depended upon nonlinear operation of the vacuum tube to locate and determine the steady state amplitude of oscillation. Nonlinear operation of the amplifier tube must introduce distortion components of the fundamental signal frequency. These distortion components are attenuated by a highly selective tuned circuit.

Resistance-capacitance oscillators have become very popular within the last decade. This paper will introduce a novel form of this oscillator, making use of the standard *R-C* network and a new amplifier, and having several marked advantages over circuits which have been used to date.

Need for Nonlinearity in an Oscillator

Let the general oscillator be described as an amplifier section and a feedback section. The gain of the amplifier will be represented by the symbol K and the transmission of the feedback section will be represented by the symbol β. Either one or both of these sections will include some form of a selective circuit so the gain through the sections will be a maximum at only one frequency, as indicated in Fig. 1.

When the input signal (E_1) and the output signal (E_3) are identical the loop may be closed to form an oscillator. If the net gain around this loop is greater than unity the oscillations will build up in amplitude with each succeeding cycle. The gain around the loop must be self adjusting in order that the oscillator maintains some particular value of oscillation amplitude. The steady state amplitude in an oscillator is defined as that signal level which just makes the loop gain unity. An amplifier operating in a nonlinear region varies its effective gain somewhat inversely with signal amplitude and thereby controls the oscillation amplitude.

Nonlinearity and Circuit Selectivity

An oscillator using any nonlinear element which is instantaneous in nature generates distortion components. The relative distortion depends upon the degree of nonlinearity over the operating range. Attenuation of these distortion components depends upon the selectivity of the circuit. Highly selective circuits permit highly nonlinear operating regions in the amplifier.

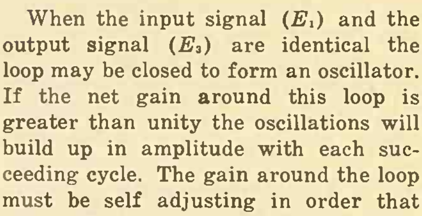
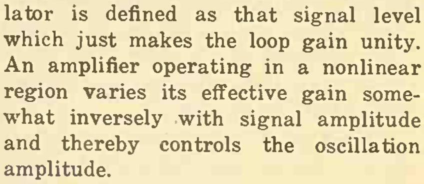


Fig. 1. Block diagram of general oscillator showing selectivity of over-all system.

Fig. 2. (A) Tuned L-C circuit. (B) Tuned R-C network.



The resistance-capacitance network shown in Fig. 2 may be used as a selective network. The Q of the *R-C* network is only ⅓ when equal values of resistance and capacitance are used. A highly selective tuned circuit is compared with the *R-C* network in Figs. 3 and 4.

Very little attenuation is offered to distortion components by the poorer selectivity of the *R-C* network. For this reason there must be control of the loop gain by a method which is not an instantaneous nonlinear element.

Nonlinear Elements Which Are Not Instantaneous

Most of the resistance-capacitance oscillator circuits use thermally responsive nonlinear devices. Generally these elements are low current 115 volt tungsten lamps. Low current fuses are also suitable for higher frequency signals. These devices have a positive temperature-resistance coefficient. Typical characteristics of a number of common nonlinear elements are given in Figs. 5 and 6.

The lamps show linear operation within each cycle for frequencies as low as ten cycles per second. They will not introduce distortion components when operated over the audio range. The small fuses show some instantaneous nonlinearity at frequencies below 100 cycles and therefore are not suitable for use at frequencies lower than 1000 cycles.

A fundamental and useful oscillator circuit using class 'A_1' amplifiers with an *R-C* feedback network is shown in

* *From the thesis submitted in partial fulfillment of the requirements for the degree of Master of Science in Electrical Engineering to the Graduate College of the University of Illinois 1949. 'A Study of Nonlinear Circuit Elements'.*

Fig. 7A. Pentodes are normally used but triodes are shown to illustrate the basic circuit.

A great amount of inverse feedback is included to stabilize the circuit. The inverse feedback circuit includes a nonlinear element R_x which controls the gain inversely as the signal amplitude varies. The gain of this amplifier is indicated in Fig. 13.

The condition required for steady state oscillation is that the loop gain is unity, that is, E_1 and E_3 are identical. This will lead to two general expressions:

$$K = 1 + \frac{R_1}{R_2} + \frac{C_2}{C_1} \quad (1)$$

$$\text{Freq.} = \frac{1}{2\pi\sqrt{R_1 C_1 R_2 C_2}} \quad (2)$$

Adjustment and selection of circuit elements and tube coefficients should be made so the action of R_x limits the oscillation amplitude to class A_1 operation. The nature of control obtained from R_x depends upon the heating by the a.c. signal current. It is to be noted that d.c. quiescent current must also pass through R_x and thus hinders the control action. Oscillator circuits of this general type have been used with great success in a large number of applications over frequency ranges of 1 to 200,000 cycles. The distortion in most cases is of the order of 2 per-cent or less.

These are the general characteristics desired of the amplifier section selected for this oscillator circuit:

1—Gain is stable and constant as a function of supply voltages, component values (in general) and tube changes.

2—Wideband frequency response and minimum phase shift over the working frequency range.

3—Low output impedance compared to values of elements used in the feedback network.

4—Nonlinear element controlling the gain inversely with signal amplitude. This element must not introduce distortion and the signal level must be within the class A_1 region.

The New Amplifier Circuit

The circuit devised by the author and shown in Fig. 7B complies in general with the specifications listed above and offers a number of advantages. A fewer number of circuit components are required with this new "π coupled" amplifier circuit. There is also one fundamental advantage in the manner in which the nonlinear element operates and responds to the signal level. These items will be discussed in detail.

The "π coupled" amplifier is that part of the circuit within the dotted lines and uses one dual triode tube, three resistors and one nonlinear element. This is a total of four circuit elements as compared to the nine components required by the amplifier in Fig. 7A. This smaller number of elements offers decided advantages where cost, space, and weight are to be considered. This is the major reason for advancing this new circuit.

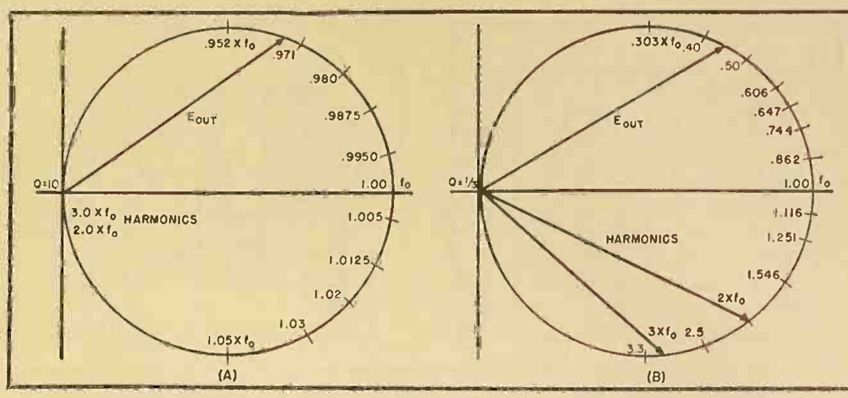


Fig. 3. Frequency response of networks for a Q of 1/3 and a Q of 10.

The second advantage results because the nonlinear element R_x does not carry quiescent current in this new circuit. The element can respond more readily to the a.c. signal component in the circuit and offers a greater sensitivity to signal amplitude. This greater sensitivity offers some compensation for the less stable amplifier characteristics. It is true that any of these circuits could be arranged to impress only the a.c. signal across the element R_x but this new circuit provides such isolation within the basic form. A total of three variations will be given for this circuit. These variations differ only in the manner of balancing out d.c. from the nonlinear element.

The amplifier shown in Fig. 7A has low output impedance because the inverse voltage feedback reduces the output impedance below the value of the amplifier without this inverse feedback. The output impedance of the new amplifier is higher because there is no inverse feedback applied. A value of 5-10 k. ohms is typical of this output impedance. The value of R_1 in the feedback section might be corrected to allow for this impedance.

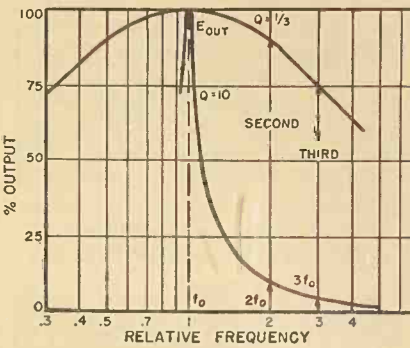


Fig. 4. Selectivity of a network for a Q of 1/3 and a Q of 10.

Examination of the circuit for the "π coupled" amplifier reveals that the first stage is operating as a grounded plate amplifier and the gain is therefore less than unity:

$$K_1 = \frac{\mu_1 R_{k1}}{R_{p1} + R_{k1} + \mu R_{k1}} \quad (3)$$

The gain of this stage is reasonably constant with respect to changes in supply voltages.

Fig. 5. Characteristics of fuses.

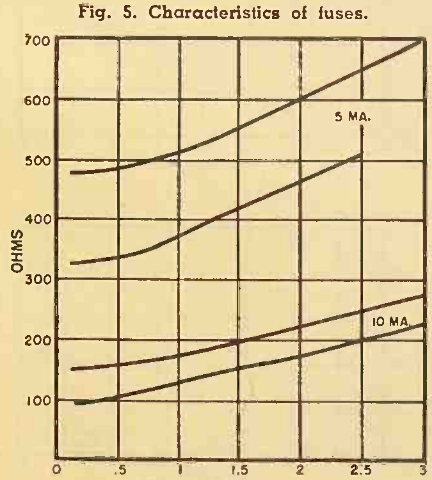
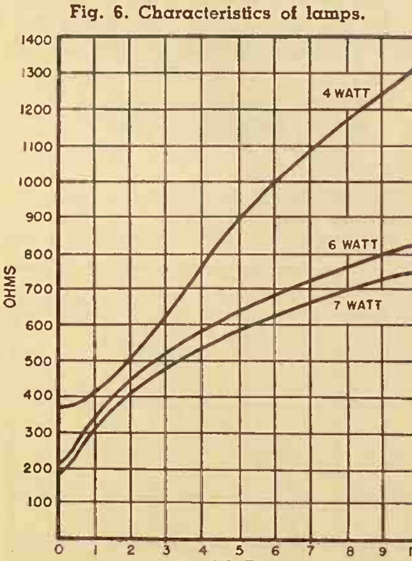


Fig. 6. Characteristics of lamps.



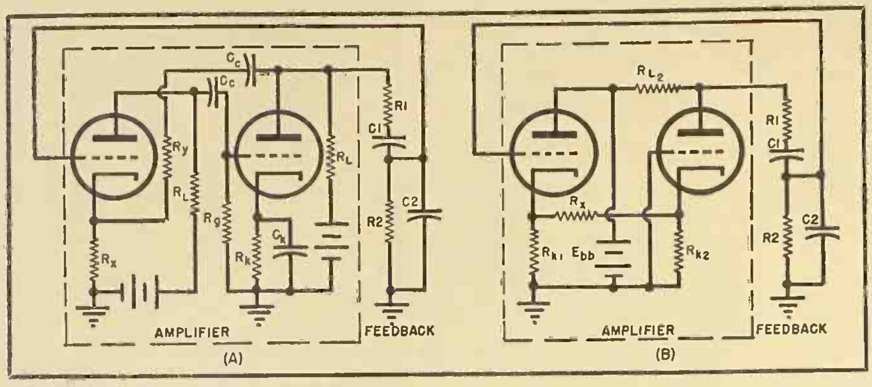


Fig. 7. (A) Fundamental circuit using class A_1 amplifiers with an R-C feedback network. (B) New "π coupled" circuit has advantages over (A).

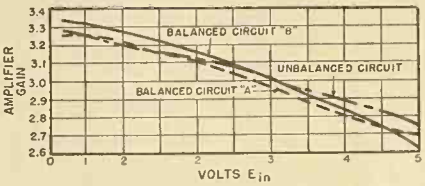


Fig. 8. Variation in amplifier gain with respect to signal level.

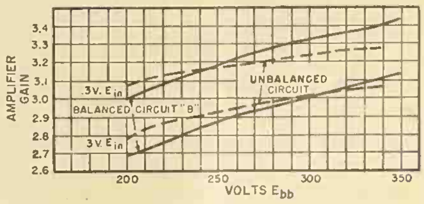


Fig. 9. Variation in amplifier gain with respect to plate voltage.

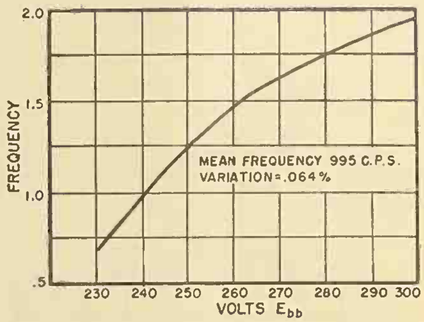
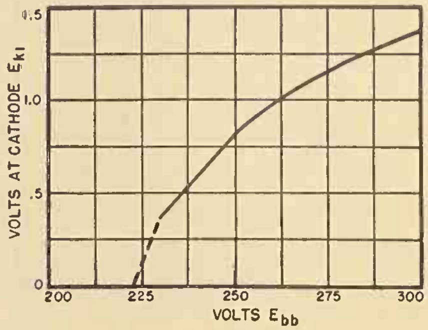


Fig. 10. Frequency stability of typical oscillator plotted against variations in plate voltage.

Fig. 11. Amplitude stability vs. plate voltage using a 115 volt, 4 w. tungsten lamp as R_x. Frequency is 400 c.p.s.



This first stage is coupled to the second stage by means of the π type network between the cathodes. For this reason the name of "π coupled" amplifier has been coined for the circuit. The second stage operates as a grounded grid amplifier, the gain of which is given as:

$$K_2 = \frac{(\mu_2 + 1)\, R_{L2}}{r_{p2} + R_{L2}} \quad . \quad . \quad . \quad (4)$$

Interaction between these stages through the π coupling network gives a net gain different than that of the product of K_1 and K_2 above. The net over-all gain has been calculated and is given in this form:

$$K = \frac{A}{B + CR_x} \quad . \quad . \quad . \quad (5)$$

where A, B and C are:

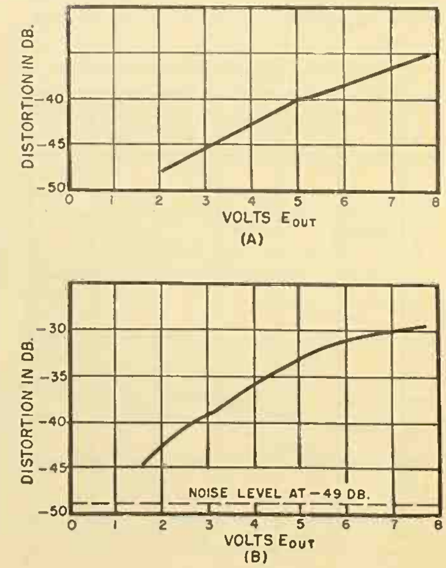
$$A = \mu_1\ (\mu_2 + 1)\ R_{L2}\, R_{k1}\, R_{k2} \quad . \quad . \quad . \quad (6)$$

$$B = (R_L + r_p)\ [r_p R_{k1} + r_p R_{k2} + (\mu + 1) R_{k1} R_{k2}] + r_p R_{k1} R_{k2} (\mu_2 + 1) \quad . \quad (7)$$

$$C = [r_p + R_{k1}(\mu_1 + 1)] \times [r_p + R_L + R_{k2}(\mu_2 + 1)] \quad . \quad . \quad . \quad (8)$$

Loading of the amplifier by the feedback network has not been considered in the calculation of amplifier gain. It is assumed that this may be neglected in most cases by proper selection of the elements in the feedback network.

Fig. 12. Distortion in 400 cycle 6SN7 R-C oscillator with (A) R_x a 4 w. lamp and E_{bb} = 200 v., (B) R_x a 5 ma. fuse and E_{bb} = 250 v.



Typical Operation

Most of these oscillator circuits are used with equal values for the resistances in the feedback network. The gain which is required is then found to be three. The dual section triodes 6SN7 and 12AU7 may be used with these approximate circuit values:

$R_{k1} = R_{k2} =$ 2,000 ohms.

$R_{L2} =$ 10,000 to 15,000 ohms selected for oscillation level.

$R_x =$ one 115 volt, 6 or 7 watt tungsten lamp.

The feedback network values are selected for the frequency desired with these general limits suggested:

$R_1 = R_2 =$ values ranging from 50k ohms to 10 megohms.

$C_1 = C_2 =$ minimum values of 100 μμfd.

These feedback network values will permit operation over and slightly beyond the audio frequency range.

A number of oscillators have been constructed using the basic circuit with fixed elements in the feedback network. These oscillators have been built within the general circuit of some projects to provide various signals at one frequency. Another oscillator has been constructed using a pair of switches to select any one of a number of fixed frequencies. Continuously variable oscillators have also been constructed using either variable resistors or variable condensers in the feedback circuit.

From a number of these different oscillators the curves given in Figs. 10, 11 and 12 have been recorded to illustrate the general operation of this circuit.

Long time measurements are not indicated for frequency stability for this oscillator does not seem to fit in this general class. The variation in frequency with respect to supply voltage is favorable on a short time test. The frequency will be sufficiently constant for most applications in mind for the circuit.

Measurements on a number of different oscillators indicated that an output voltage with 1 per-cent or less total distortion can be obtained operating near the oscillation threshold. Additional measurements indicate that lower distortion was obtained, for any given output voltage level, when the plate supply voltage is increased and the amplifier gain readjusted for the same output level. In making the observation the gain was adjusted by varying the value of R_{k2}. In other circuits the

(*Continued on page* 24)

VIDEO AMPLIFIER DESIGN

By ROBERT C. MOSES

Sylvania Electric Products Inc.

Various types of compensation may be used to give an amplifier with essentially flat amplitude and phase characteristics over the entire video band.

HIGH definition television systems are required to convey information transmitted at rates up to 8,700,000 picture elements per second. The ultimate resolution obtainable in the reproduced image is determined by the ability of the equipment as a whole to respond to the maximum rate of transmission. The latter, in turn, is directly proportional to the bandwidth of the circuits through which the video signal must pass. With the present 525 line, 30 frame transmission standards, it can be shown that adequate response up to a maximum frequency of 4.35 megacycles is necessary for faithful reproduction of the finest detail in the picture. On the other hand, satisfactory rendition of backgrounds and of very large objects requires that the system be capable of reproducing one picture element having dimensions approaching those of the entire scanned area. Proper response of the system down to d.c. is therefore indicated. Since transmission of d.c. in the system is a subject beyond the scope of this article, the following discussion will be limited to amplifiers responding to frequencies down to about 30 cycles.

The necessity for handling such a wide spread in rates of transmission imposes severe requirements upon the video amplifier channel. Not only must the amplifier be capable of passing the entire frequency range from 30 cycles to over 4.0 megacycles with a minimum of amplitude discrimination, but also the transmission time through the system must be as nearly constant as possible for all frequency components within these limits. The latter requires that the curve of amplifier phase shift versus frequency be essentially linear up to the maximum frequency present in the video signal.

The usual resistance-coupled audio amplifier falls far short of these requirements, since both the amplitude and phase characteristics deteriorate rapidly above about 10,000 cycles and below 70 cycles. Some form of compensation for both amplitude response and phase shift is obviously necessary. There are several methods by which such compensation may be accomplished, and for purposes of discussion, the low and high frequency considerations will be treated separately.

Even if the amplitude response of the conventional pentode resistance coupled amplifier were satisfactory at low frequencies, the phase characteristics below 60 cycles would be troublesome. Excessive phase shifts at low frequencies may be attributed to three sources:

1. The time constant of the grid coupling network
2. The screen circuit time constant
3. The cathode biasing network

These factors are indicated in the equivalent circuit of Fig. 2A. The plate supply terminals and d.c. connections are omitted for clarity.

The reactance of the coupling capacitor C_c at low frequencies introduces a leading phase shift and a drop in amplitude in the component E_o of output voltage across the grid resistor of the following stage. These effects become serious below the frequency at which the reactance of C_c exceeds about ⅛ the resistance of R_g, and may be sufficiently great to prevent proper reproduction of square waves of 60 cycles base frequency unless large values of C_c and R_g are used. The magnitude of the phase angle is given by:

$$\varphi = -\tan^{-1} \frac{1}{2\pi f C_c R_g}$$

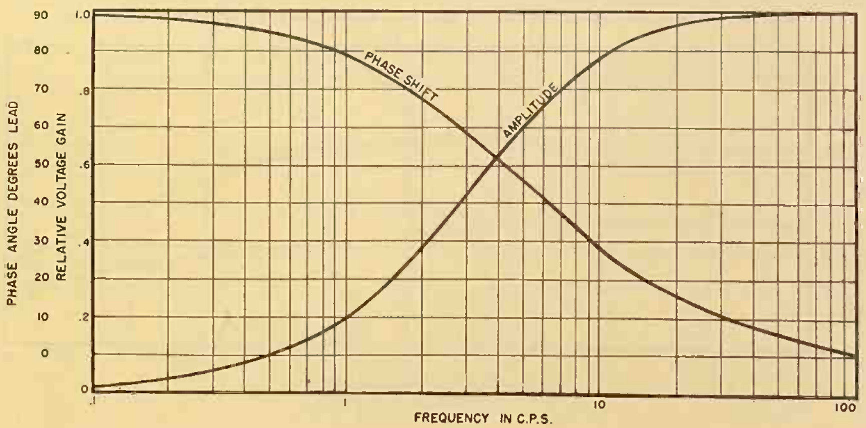
and the relative amplitude characteristic by:

$$\frac{E_o}{E_i} = \cos \varphi$$

where E_o and E_i are the output and input voltage respectively.

While it is theoretically possible to obtain values of C_c and R_g which will keep the frequency of negligible phase shift below 30 cycles, large values of coupling capacitors introduce additional stray capacitances which may affect the high frequency characteristics of the amplifier. At the same time, the maximum value of the grid resistor R_g is usually limited by gas and grid current considerations in the following tube. It is therefore customary to assign reasonable values to C_c and R_g, and

Fig. 1. Low frequency phase and amplitude characteristics of a typical two-stage video amplifier with compensation.



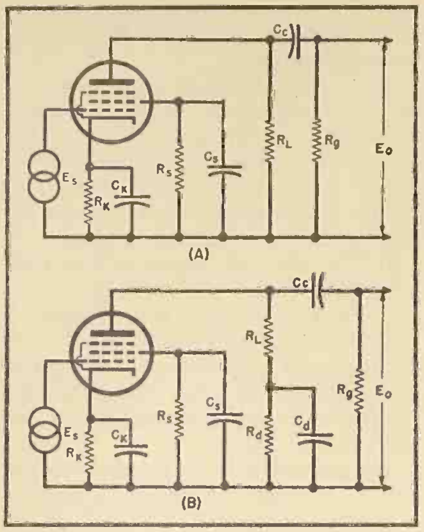


Fig. 2. (A) Low frequency equivalent circuit. (B) Equivalent circuit with low frequency compensation.

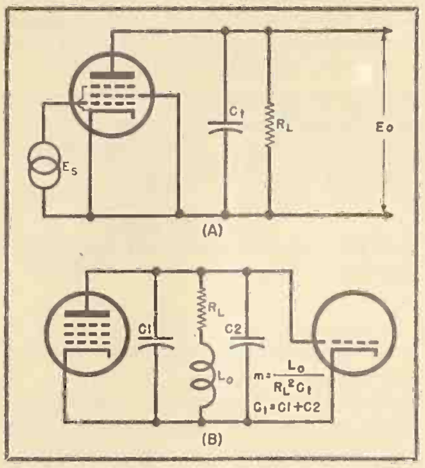


Fig. 3. (A) High frequency equivalent circuit. (B) Shunt peaking network.

provide compensation for the resulting phase shift at some point in the system.

The finite time constant of the screen resistor-capacitor combination R_sC_s causes a shift in phase and a drop in amplitude at low frequencies in exactly the same way as the grid coupling network, and the extent of these deficiencies may be specified in the same manner. However, the phase angle and departure from flat amplitude response introduced by R_sC_s can be kept small with practical component values, and their effect need not be compensated if:

$$R_s C_s \geqq 0.077$$

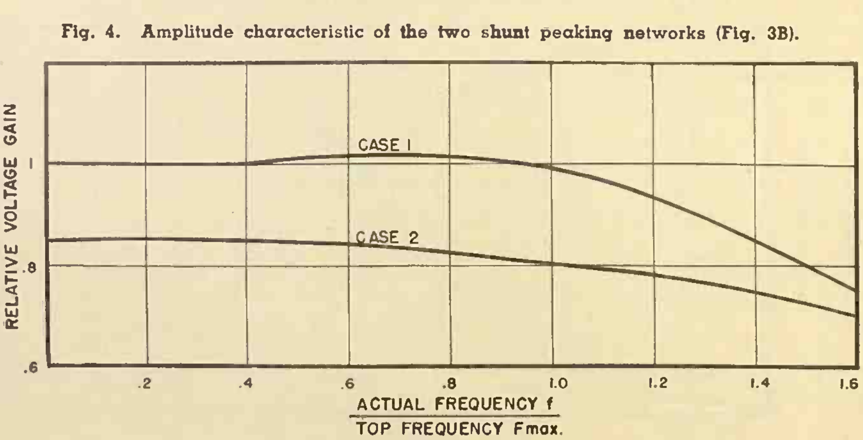
where R_s is in megohms and C_s is in microfarads, a condition readily achieved in most cases. If the above holds, the phase angle due to R_sC_s will not exceed 4° at 30 cycles.

The impedance of the cathode bias network C_kR_k results in degenerative effects and excessive phase shifts at frequencies below that at which the reactance of C_k exceeds about 1/10 the resistance of R_k. In many cases, prohibitively large values of C_k are required if the frequency of near-zero phase shift is to be comparable with the frame repetition frequency. Compensation for low-frequency discrepancies due to C_kR_k can be provided in much the same manner as for the grid coupling network; however, the design constants for cathode-bias impedance compensation are different.

Despite the fact that there are several arrangements for compensating the effects of the grid coupling and cathode biasing networks, one simple form of correction circuit has been widely used. This consists of a parallel resistor and capacitor in the plate circuit of the stage to be compensated, as shown in Fig. 2B.

In this circuit, the additional components C_dR_d in series with the plate load of the tube introduces a compensating phase shift which causes the output voltage E_o to lag in phase by the same amount as the leading phase angle produced by either C_cR_g *or* C_kR_k. At the same time, the reactance of C_d causes a rise at low frequencies in the effective load impedance presented to the tube. The resulting increase in gain counteracts deficiencies in amplitude response. It is usually not possible to satisfy the conditions for correction of both grid coupling and cathode biasing network effects in one such correction circuit. An entirely satisfactory degree of compensation for either can, however, be achieved through a suitable choice of correction parameters.

Fig. 4. Amplitude characteristic of the two shunt peaking networks (Fig. 3B).



Compensation for Grid Coupling Network

For a satisfactory degree of compensation for phase shifts produced by the grid coupling network, the following relations must hold:

1. The time constant of the plate load resistor and compensating capacitor must equal the time constant of the grid coupling network.

$$C_d R_L = C_c R_g$$

2. The resistance of the decoupling resistor must be at least ten times the reactance of the compensating capacitor at the lowest frequency for which correction is required.

$$R_d \geqq 10/\omega C_d$$

It should be noted that this type of compensation for deficiencies in the grid coupling network will not give perfect correction to very low frequencies or to d.c. except in the theoretical and practically unattainable condition of infinite decoupling resistance R_d. For a fixed value of C_d, the lowest frequency for which compensation is exact is inversely proportional to the value of R_d.

The usual procedure in compensating a stage for the effects of the grid coupling network is to select a suitable value for C_d, say 10 microfarads. This, in combination with the plate load resistor as determined from high frequency considerations, gives the time constant to be obtained in the following grid circuit. The decoupling resistor R_d is then made as large as possible consistent with obtaining the required plate voltage at the tube.

Compensation for Cathode Bias Network

Correction for phase shifts and amplitude discrimination due to the cathode biasing network can be effected in a similar manner to the above, except in this case, exact compensation can be achieved down to d.c. with practical circuit values. Perfect correction will be obtained if:

1. The time constant of the compensating network is made equal to that of the cathode bias network.

$$C_d R_d = C_k R_k$$

2. The decoupling resistor is made equal to the product of the midband stage gain and the value of the biasing resistor.

$$R_d = \bar{A} R_k = G_m R_L R_k$$

where G_m is the tube transconductance.

Compensation for All Low-Frequency Effects

It will not usually be possible to obtain values for the correction circuit, such that the requirements for phase and amplitude compensation of both grid coupling and cathode bias networks are simultaneously satisfied. For correction of all low-frequency effects in a multistage amplifier, the preferred procedure is to make the grid circuit time constants as large as possible in stages having a cathode bias network, and compensate the latter in each plate circuit. Then, for every two or three such stages, a stage having no cathode bias network is provided, and the combined phase shifts of all grid couplings compensated in the plate circuit of this stage. This arrangement operates quite satisfactorily provided the total phase shift to be corrected is smaller than about 20° at 60 cycles. This figure is readily achieved if the time constants of the uncompensated grid coupling networks are from 8 to 15 times greater than the period of the lowest frequency to be passed.

For adequate vertical resolution of a 525 line picture, the maximum delay time $\left(\frac{1}{\text{phase shift} \times \text{frequency}}\right)$ variation tolerable at the low end of the video band is on the order of 850 microseconds. With two video stages involved, a phase shift of 4 or 5 degrees per stage at 30 cycles can be considered allowable for design purposes. The amplitude response should be flat within a few tenths of a decibel to well below 30 cycles if the above requirement is to be met. Low frequency phase and amplitude characteristics of a typical two-stage video amplifier with compensation are shown in Fig. 1.

Tube interelectrode capacitances and stray capacitances associated with circuit components and wiring have a shunting effect which produces excessive phase shifts and departure from linear amplitude characteristics at the higher video frequencies. For a fixed value of plate load resistance, the absolute magnitude of these capacitances becomes the limiting factor on the top frequency f_o, at which satisfactory performance can be obtained. Referring to the high-frequency equivalent circuit of Fig. 3A, it can be shown that a phase shift of 45° together with an amplitude drop of 3 decibels will occur when:

$$f_o = \frac{1}{2\pi C_t R_L}$$

where C_t is the total shunting capacitance, and R_L is the plate load resistance (Fig. 3A). For half-power (3 db.) response to 4.5 megacycles, and a total shunt capacitance of 25 $\mu\mu$fd., (a reasonable figure for a two-terminal interstage) the above equation indicates that the load resistor must be 1400 ohms. The stage gain is given by:

$$\overline{A} = G_m R_L$$

For a tube with a transconductance of 9000 ohms the stage gain will be 12.6.

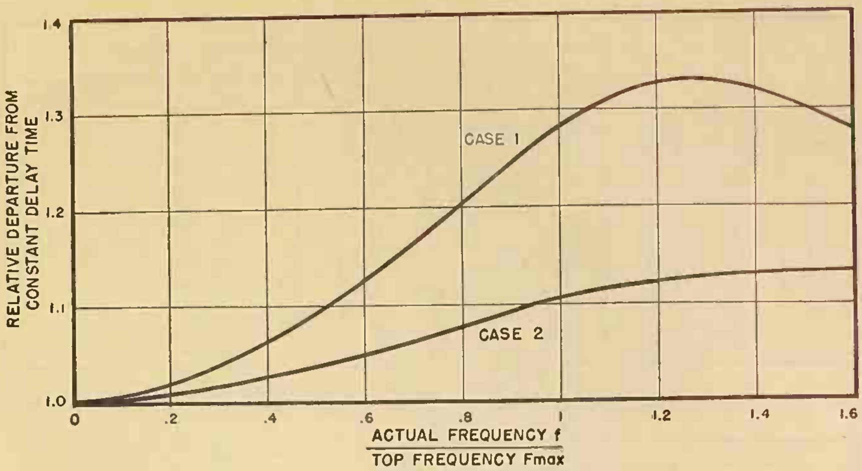


Fig. 5. Time delay characteristics of the two shunt peaking networks (Fig. 3B).

The above considerations lead to a "gain bandwidth" (hereafter abbreviated G.B.) rating for video amplifier tubes, in which the input and output capacitances are compared with the transconductance. For tubes of the same type, coupled by a two terminal network, the input and output capacitances add directly, and the G.B. factor F' is:

$$F' = \frac{G_m}{2\pi\ (C_1 + C_2)}$$

where C_1 and C_2 are input and output capacitances respectively.

With four terminal and certain filter couplings, the input and output capacitances are isolated from each other. The G.B. factor F'' in this case is:

$$F'' = \frac{G_m}{2\pi \sqrt{C_1\ C_2}}$$

Gain-bandwidth factors for several common tube types are given in Table I.

The G.B. factor has the dimensions of frequency, and denotes the bandwidth at which the voltage gain of an ideal stage becomes unity. Thus, a type 6AC7 with a two-terminal interstage coupling would have a gain of 1 at a half-power bandwidth of 90 megacycles; conversely, with a 1 megacycle bandwidth, the stage gain would be 90 times. These figures take into account the tube capacitances only, and do not include the unavoidable strays.

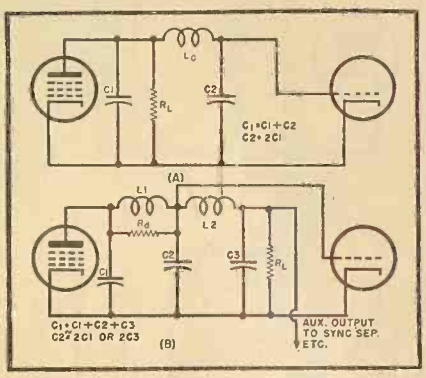
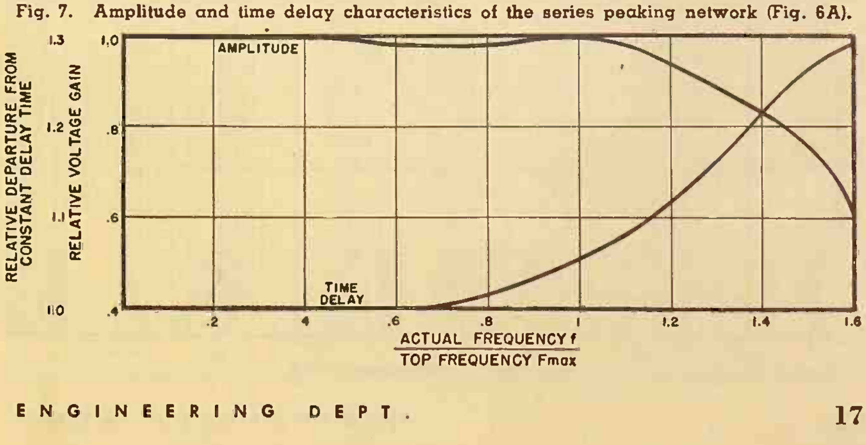


Fig. 6. (A) Series peaking network. (B) Two section filter coupling.

From the above discussion, it is apparent that a tube having high G_m and low interelectrode capacitances (largest G.B. factor) is the most desirable for video amplification. At the same time, the figures given represent the ultimate performance of which the tube is capable, and will not be realized in practice because of the presence of external circuit capacitances which tend to restrict the bandwidth. Furthermore, the G.B. factor is based on half-power bandwidths, and for television video applications, which may involve several amplifier stages, a much more linear phase and amplitude characteristic in each stage is necessary. The G.B. factor does, however, serve to indicate the rel-

Fig. 7. Amplitude and time delay characteristics of the series peaking network (Fig. 6A).



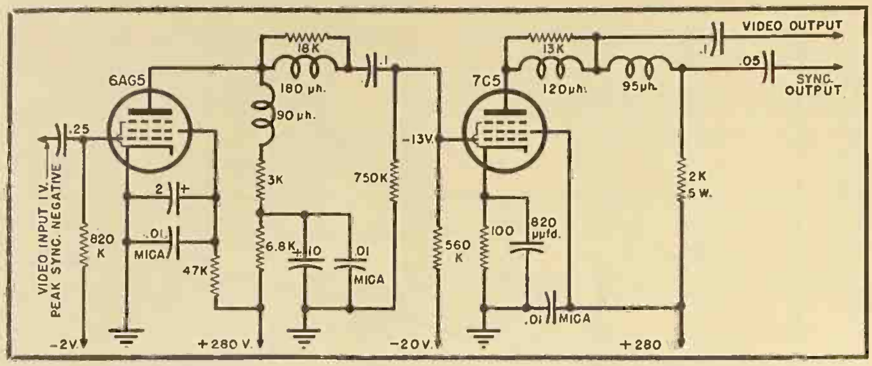


Fig. 8 Typical two-stage video amplifier incorporating high and low frequency compensation. Over-all gain is 90, and bandwidth is 4.35 mc.

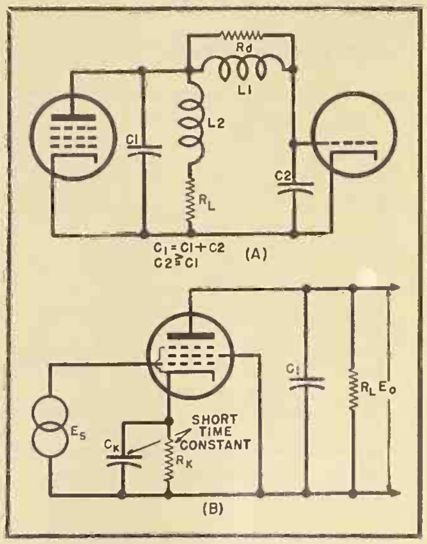


Fig. 9. (A) Modified shunt-series peaking. (B) Cathode peaking.

ative merit of the various tube types in wide-band amplifier applications.

High Frequency Compensation

It is possible to approach more nearly the theoretical performance indicated by the G.B. factor for any tube type through the use of one of a number of two or four terminal interstage coupling networks. Such networks may provide nearly complete phase and amplitude compensation up to a top frequency F_{max} which equals or exceeds

$$\frac{1}{2\pi C_t R_L}$$

This is to be contrasted with the uncompensated case where the departure from ideal phase and amplitude characteristics at the top frequency is very much greater than can be tolerated. In the following discussion of several such coupling networks, the basic circuit parameter for design purposes is the reactance of the total shunting capacitance C_t at the top frequency F_{max}. The figures of merit given in each case apply to the network itself, and indicate its midband voltage gain in relation to the uncompensated case.

Two Terminal Networks

Shunt Peaking, case 1 (Fig. 3B):

$m = 0.50$

$C_t = C_1 + C_2$

$$R_L = \frac{1}{2\pi F_{max} C_t}$$

$L_o = 0.5\, R_L^2\, C_t$

Figure of merit 1.0

Shunt Peaking, case 2

$m = 0.41$

$C_t = C_1 + C_2$

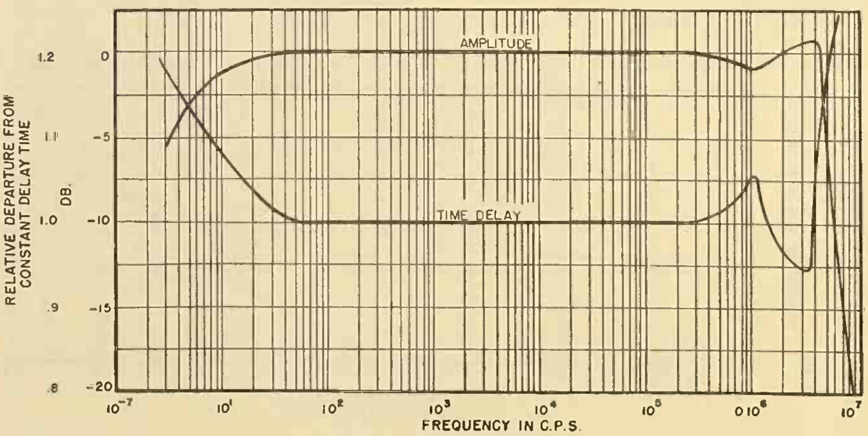
$$R_L = \frac{.85}{2\pi F_{max} C_t}$$

$L_o = 0.41\, R_L^2\, C_t$

Figure of merit 0.85

The shunt peaking circuit of Fig. 3B is the simplest type as it involves only one inductance. It has the lowest figure of merit of any coupling network, but has the advantage of permitting relatively wide tolerances in circuit values with little change in performance. The design coefficient m relates the reactance of the peaking inductor L_o to that of the total shunt capacitance C_t at the frequency of correction, and determines the linearity of the phase characteristic up to the top frequency F_{max}. In case 2, improved phase response is obtained at the expense of gain factor. Figs. 4 and 5 show amplitude and time delay characteristics of the two shunt peaking networks, as a function of the ratio of actual to top frequency.

Fig. 10. Time delay and amplitude characteristics of amplifier of Fig. 8.



Four Terminal Networks

Four terminal networks give improved figures of merit, since the tube input and output capacitances are effectively isolated from each other. On the other hand, four terminal networks are more difficult to adjust, and frequently require certain specified ratios between the capacitances at each end. In general, both the figure of merit and the linearity of phase and amplitude response will be determined by the manner in which the capacitances are distributed; this is particularly true of the more complicated structures.

Case 1 Series Peaking (Fig. 6A).

$C_t = C_1 + C_2$

$C_2 = 2C_1$

$$R_L = \frac{1.5}{2\pi F_{max} C_t}$$

$L_c = 0.67\, R_L^2\, C_t$

Figure of merit 1.5

The series peaking network of Fig. 6A provides a gain of 1.5 over the uncompensated case, at the same time permitting a moderate tolerance with respect to capacitance distribution and circuit values. It is sometimes possible to achieve a slightly greater figure of merit than the above figure would indicate by using higher values of inductance and load resistance, but this procedure may require resistive damping of L_c in order to obtain sufficiently uniform response.

It might be noted that the positions of the source and load may be interchanged so that R_L is at the output end, with no change in network characteristics. This may be found necessary in certain cases where the terminating capacitance C_2 is less than the input capacitance C_1. In any case, the load resistor R_L should be on the low capacitance side of the peaking inductor L_c.

Capacitance ratios other than 2 to 1 or 0.5 to 1 may be accommodated at the expense of uniformity of phase and amplitude response. If, in Fig. 6A, the

(*Continued on page 27*)

NEW PRODUCTS

INFRARED SPECTROPHOTOMETER

A double beam infrared spectrophotometer which will meet the demands of the structural chemist for high resolu-

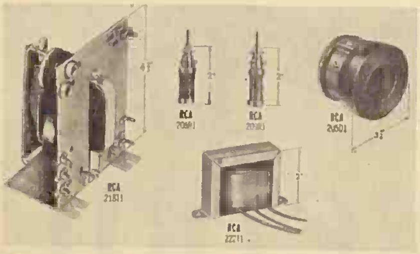
tion and sensitivity, and the analytical chemist for speed and accuracy, is now in production at the plant of the *Perkin-Elmer Corporation*, Glenbrook, Conn.

Designated Model 21, the instrument records directly in percent transmission against a linear wavelength scale on large, easily read charts. The instrument's resolution is up to the limitations imposed by the Raleigh criterion and Johnson noise. Its speed of scanning ranges from 3 minutes to 100 hours for the rock salt region, and the wavelength drive speed can be suppressed automatically when desired. The over-all range of the instrument is from less than 2 microns to 15 microns in the rock salt region.

Chart scales of the *Perkin-Elmer* Model 21 are uniform from 1 to 50 inches per micron by integral factors. The amplifiers and power supplies are external.

PICTURE TUBE COMPONENTS

The Tube Department of *RCA*, Harrison, N. J., recently offered to equipment manufacturers five new components for the RCA 16GP4 deflection sys-



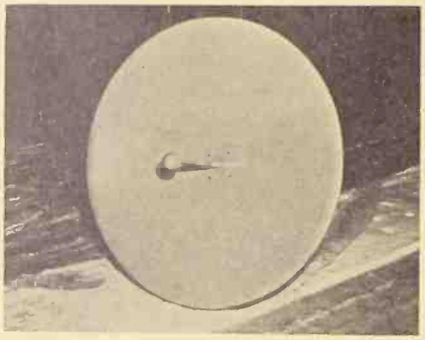
tems designed to use the horizontal-deflection amplifier tube 6CD6-G and the vertical-deflection amplifier tube 6S4.

These new components are as follows: Deflecting Yoke, Type 206D1; Width Control, Type 208R1; Horizontal-Linearity Control, Type 209R1; Horizontal-Deflection-Output and High-Voltage Transformer, Type 218T1; and Vertical-Deflection-Output Transformer, Type 222T1.

PARABOLIC ANTENNAS

Five new parabolic antennas to cover the 5929-7125 mc. frequency band are now available from the *Workshop Associates, Inc.*, 66 Needham St., Newton Highlands 62, Mass.

Each parabola is available in two, four, six, and eight foot diameters, and



mounts are available for all types of installations. The antennas have gains up to 44.9 db., and can be supplied with complete de-icing equipment and junction boxes.

A free booklet giving complete descriptions of these antennas may be obtained upon request.

TEST EQUIPMENT

A new instrument which indicates on the screen of an oscilloscope instantaneous speed vs. time is announced by *Kay Electric Co.*, Maple Avenue, Pine Brook, New Jersey. The Rotalyzer uses a properly magnetized disc and a pick-up to provide a signal whose frequency is proportional to instantaneous speed. An accurate determination of this frequency is made and indicated on the cathode-ray oscilloscope screen. Average speed is read on a panel meter.

This instrument may be used in the study of torsional vibrations in rotating shafts, in the study of "wow" in phonograph equipment, and in the measurement of shaft rotational changes in computer systems. The r.p.m. range of 900 to 7200 r.p.m. may be extended down to 33-1/3 r.p.m. and upward to 50,000 r.p.m. Speed variation is up to 200 c.p.s.

ELECTRONIC TILE SORTER

An electronic tile sorter designed primarily to detect flaws in 4¼" x 4¼"

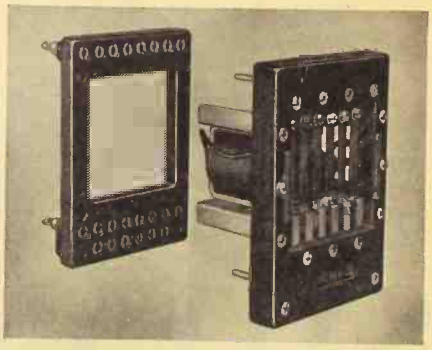
x ¼" double-fired tile when it is in the bisque state has been announced by *Electronic Associates Incorporated*, Long Branch, N. J.

The Model 203 Tile Sorter is adjustable to meet any possible conditions of mixture or firing. Adjustment to any level desired by the tile manufacturer is dependent on the following ratios: 1) When set to reject 100% of the tiles with objectionable air pockets, it will also reject 100% of the cracked tile and will reject all tiles under 200 lbs. on the 3-point break test; 2) When set to reject 95% of the tiles with objectionable air pockets, it will reject 87% of the cracked tiles and will reduce the break test to 110 lbs; and 3) When set to reject 90% of the air pocket tiles, it will reject 80% of the cracked tiles and the break test is reduced to 80 lbs.

This sorter is capable of sorting tiles at any rate up to 10 tiles per second, which is much faster than the tiles can be brought to the machine using present conveyor methods.

HIGH SPEED RELAY

The Autocall Company, Shelby, Ohio, has released for general use their ultra high-speed "HHA" relays which were originally designed and built as a basic component of the Harvard University's



Mark II Calculator for the Navy Department.

These relays, according to the manufacturer, are built for long life with 100 million operations, minimum, at high speed of 6 to 8 milliseconds and incorporate many features including six single-pole double-throw contacts arranged side-by-side to provide 28 different contact combinations; individual contacts and prongs are of one-piece construction, and each contact provides its own prong for jack-connecting to the circuit.

Coils can be furnished for 24 to 150 volts, d.c., with no increase in price for the various voltages.

SOLDERING UNIT

A soldering unit, operating on the conduction principle, and suitable for all types of soldering operations including silver soldering and brazing, is now

available from *Wasserlein Mfg. Co., Inc.*, 7400 3rd Ave., No., St. Petersburg 6, Florida.

The Wassco Glo-Melt has a 24 heat selector to handle lightest up to heaviest work. With capacity equal to a 450 watt heavy electric iron, Glo-Melt soldering is done with a light handpiece weighing only 5 oz., which is said to be as easily and accurately handled as a pencil. According to the manufacturer, this unit is faster and more accurate, consumes less power, and uses solder and flux.

Glo-Melt literature and a use-manual are available from the manufacturer.

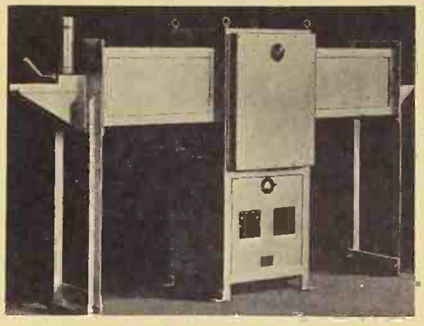
TIME INDICATOR

Electronic Systems Co., 555 East Tremont Avenue, New York 57, N. Y., has announced its Model 632-B Pulse Rise Time Indicator for the accurate plotting of the rise time of rapidly rising positive voltage pulses. The device employs a specially designed delay line of variable length and a vacuum tube voltmeter.

The 632-B features a radical change in the method of analyzing the build-up time of a pulse. However, the most important factor is the range available: .005 to .1 microseconds in 20 steps. Other ranges are available upon request.

PILOT TUNNEL KILN

Harper Electric Furnace Corporation, Niagara Falls, N. Y., has announced a small tunnel kiln for speeding up pro-



duction test runs. Its versatile design permits variations in firing cycles to allow duplication of most firing schedules.

These units are being used in test firing of ceramic powders, steatite, electrical porcelain, insulators, resistors, spark plugs, grinding wheels, newly developed electronic components, and other ceramic products.

Further information may be obtained by writing the company direct.

FM SIGNAL GENERATOR

Boonton Radio Corporation, Boonton, N. J., has announced the Type 202-D frequency modulated signal generator designed for use with telemetering receiver equipment and in other associated applications.

This instrument covers the frequency range from 175 to 250 megacycles and is provided with three continuously adjustable deviation ranges: 0-24 kc., 0-80 kc., and 0-240 kc. In addition, amplitude modulation up to 50% may be obtained using the internal audio oscillator and

modulation to 100% using an external audio oscillator.

The deviation sensitivity of the frequency modulation system is within ± 0.5 db. from d.c. to 200 kc. The ampli-

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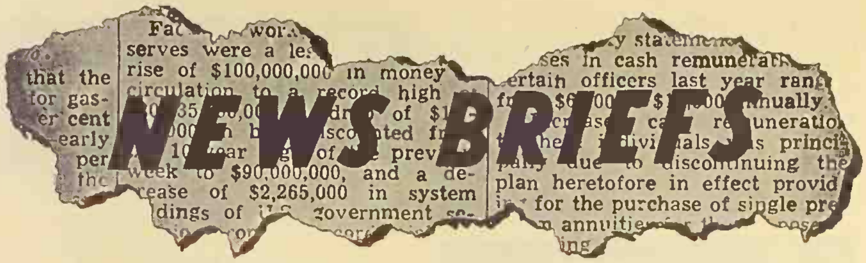


DR. LANGMUIR RETIRES

Dr. Irving Langmuir, regarded as one of the greatest scientists of modern

times, has retired as associate director of *General Electric Company's* Research Laboratory. Dr. Langmuir will continue work in the laboratory in a consulting capacity, but will engage primarily in activities of Project Cirrus, joint weather research program of the U. S. Army Signal Corps and the Office of Naval Research in consultation with *General Electric*.

Dr. Langmuir, who has received the world's top-ranking scientific awards, including the Nobel Prize in chemistry in 1932, was responsible for the development of the gas-filled incandescent lamp, the high-vacuum power tube, atomic hydrogen welding, screening-smoke generator for the military, and methods for artificial production of snow and rain from clouds.

MICROWAVE REPEATER IN PRODUCTION

The *Philco* feedback-type microwave repeater announced at the recent annual session of the Communications Section of the Association of American Railroads is now being manufactured on a production basis.

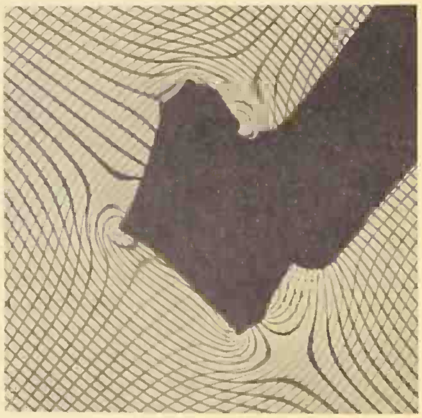
Capable of handling up to 32 two-way voice channels or combinations of voice and coded intelligence, the *Philco* CLR-5 microwave repeater is designed for operation in the 6575-6875 megacycle band, which is available to railroads. This feedback repeater circuit permits multiple chain repeats with very little distortion and makes possible the use of only one microwave oscillator tube for both transmitting and receiving functions of a single-direction repeater.

The Industrial Division of *Philco Corporation*, Philadelphia 34, Pa., will give further information and detailed technical specifications upon request.

ELECTRIC FIELD

A shadow technique, developed by Dr. L. L. Marton and associates of the National Bureau of Standards, now makes it possible to photograph and study quantitatively electrostatic and magnetic fields of extremely small dimensions.

The striking pattern shown is the result of an electric field about a charged crystal of barium titanate when studied by the electron-optical shadow



method. This method offers a powerful tool for exploring fields that have not been susceptible to other methods of investigations.

An electron lens system is used to produce a shadow image of a fine wire mesh placed in the path of an electron beam. From the distortion in the shadow network caused by deflection of the electrons as they pass through the field under study, accurate values of field strength are computed. In the photograph, the distorted shadow image of the wire mesh is superposed on the image of the crystal (center) and its 0.010-inch tungsten-wire support.

NEW SITE FOR NBS LABORATORY

Approval has been given for the development of a site at Boulder, Colorado, for additional Bureau laboratory facilities. To be used initially by the Bureau's Central Radio Propagation Laboratory, the site consists of about 210 acres directly south of the city.

Laboratory facilities at Boulder for research in radio propagation at a cost of about $4,500,000 have been authorized by Public Law 366 of the 81st Congress. Actual construction is expected to start in the summer of 1951 and when completed will maintain a research staff of about 300 people.

NBS TO USE MEMORY UNITS

The first of three *Technitrol* memory units has been obtained by the National Bureau of Standards for use in the latest type of large scale electronic computer developed jointly by funds of the Air Comptroller Department of the Army and NBS.

The memory unit, based on the principle of the mercury delay line originally conceived at the University of Pennsylvania and manufactured by the *Technitrol Engineering Co.*, is an automatic brain which will remember for later recall any sequence of facts or figures once placed on it. It has reached its latest development in the computer to be known as the NBS Interim Computer.

In addition to contemplated work in processing the 1950 census figures, the Interim Computer will be studied in an effort to develop new and better computers for Government and commercial use. It will also be utilized for solving many of the scientific problems that arise in the normal work of the Bureau.

ULTRASONIC GENERATOR

One of the many electronic developments described in "Electrical and Allied Developments of 1949" published in the *General Electric Review* was an ultrasonic generator having a quartz

transducer on top of the power supply and operating in an oil-filled well.

Power at 300, 500, 750, or 1000 kc. is supplied by the equipment. Applications of this ultrasonic generator include assistance to chemical reactions, bacteria control, and the mixing of immiscible liquids such as mercury into water.

IMPROVED MULTIPLIER PHOTOTUBE

Important advances in nuclear research, astronomy, photoelectric spectrometry, and other fields involving work with light at extremely low levels

is expected with the announcement of a greatly improved 1P21 multiplier phototube by *RCA's* Tube Department. Heart of the cyclotron used in radiation research at the University of Rochester is an *RCA* 1P21 multipler phototube shown partially disassembled. The "equivalent noise input" of the improved 1P21 has been reduced to $5x10^{-13}$ lumen at room temperature. This value represents a six-fold reduction in operational noise and permits a corresponding reduction in the lower limit of measurable light intensities.

This extension in the range of the 1P21 makes it an even more valuable aid for astronomers studying light from distant stars, for nuclear scientists studying atomic radiation, and for other laboratory research work requiring measurement of light of extremely low intensity.

INSTRUMENT TO RECORD BRAIN AND HEART ACTION

Equipment for making radar-like pictures of the heart and brain in action was described by Dr. Stanford Goldman, professor of electrical en-

gineering at Syracuse University and former research associate in the M.I.T. Research Laboratory of Electronics, at the recent IRE Convention. Results already obtained indicate that the pictures will be useful for the diagnosis of disease in the heart and brain and in studying the physiology of these vital organs.

The equipment, first conceived and developed at M.I.T., picks up tiny electrical impulses which flow through nerves in the body. These impulses are then converted into a constantly moving map-like picture similar to those made by World War II radar sets.

Equipment shown in the center of the photograph amplifies signals from electrical pick-ups on the patient at the right. Dr. Goldman at the left is shown examining a printed record of these signals while William F. Bantelmann, Jr., M.I.T. Research Assistant in electrical engineering, watches the visional pattern.

NEW LITERATURE

Handling of Radioactive Isotopes

The National Committee on Radiation Protection, established by the National Bureau of Standards, has prepared a handbook giving recommendations for the safe handling of artificially-produced radioactive isotopes in the typical laboratory or small industrial operation.

Handbook H42, Safe Handling of Radioactive Isotopes, gives general safety recommendations which may be modified to suit the control requirements of different types of operations. Specific problems discussed are those of personnel, laboratory design and equipment, hazard instrumentation, hazard monitoring, and transportation of isotopes. The increasing use of radioactive isotopes by industry, the medical profession and research laboratories has made it essential that certain minimal precautions be taken.

This handbook is priced at 15 cents a copy and is available from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

Paper Laminates

A report prepared by the Plastics Laboratory of Princeton University under joint sponsorship of the National Defense agencies describes the development of a low-loss, arc-resistant paper laminate for high frequency applications.

The report, released to the public by the Office of Technical Services of the U. S. Department of Commerce, deals with modifying standard melamine-formaldehyde impregnating resin with varying amounts of hexamethylenediamine to produce a laminated material with high dielectric strength,

(*Continued on page* 25)

Personals

DR. MERLE M. ANDREW has joined the staff of the machine development section of the National Bureau of Standards applied mathematics laboratories. Before coming to the Bureau, Dr. Andrew was engaged in operational research with the Naval Operations Evaluation Group and prior to that time conducted radar research at the Radiation Laboratory of MIT. He is a member of the American Mathematical Society and the American Physical Society.

ARTHUR L. CHAPMAN has been named general manager of the *Colonial Radio and Television Division* of *Sylvania Electric Products Inc.* With headquarters in Buffalo, N. Y., Mr. Chapman will be responsible for production and sales of the division in addition to his duties as general manager of *Sylvania's* newly formed Parts, Wire and Plastics Division at Warren, Pa. Mr. Chapman has been associated with *Sylvania* since 1933.

CURTIS R. HAMMOND has been appointed equipment sales manager of the Receiving Tube Division of *Raytheon Manufacturing Company.* He will be responsible for the sale of radio receiving tubes and cathode-ray picture tubes and will direct activities from *Raytheon's* Chicago warehouse. Mr. Hammond joined *Raytheon* five years ago after many years in sales and sales engineering work with *Ken-Rad Tube and Lamp Corporation.*

DELMAN E. ROWE has been appointed senior test director in the guided missiles research laboratory of the National Bureau of Standards. Mr. Rowe, who received the Naval Ordnance Development Award with Certificate by the Naval Bureau of Ordnance in 1945, is a member of the Institute of Radio Engineers, is the inventor of a microwave switch and attenuator, and has published several technical papers in the electronics field.

ROBERT A. STAREK has been appointed field engineer for the Radio Tube Division of *Sylvania Electric Products* at Emporium, Pa. Formerly commercial engineer, Mr. Starek joined the Engineering staff of *Sylvania* immediately after receiving his B.S. degree in electrical engineering from Iowa State College. He is a member of the American Institute of Electrical Engineers and the IRE and will make his headquarters in the Cincinnati office.

JEROME R. STEEN, director of quality control for *Sylvania Electric Products, Inc.*, New York, has been elected to grade of Fellow by the board of directors of the Institute of Radio Engineers. Mr. Steen joined the engineering staff of the Radio Division of *Sylvania* at Emporium, Pa., in 1931 and since that time has been active in the application of quality control and standardization methods in radio tube manufacture.

New R-C Osc.

(Continued from page 14)

plate load resistor R_{L2} or the alternate cathode resistor R_{k1} may be varied for manual adjustment of amplifier gain. The higher supply voltages give a more linear operating region for the amplifier tubes.

Oscillators of this type have been adjusted to operate with various supply voltages ranging from 100 to 350 volts. When distortion is considered important the circuit should be supplied with higher voltages and adjusted to as near the oscillation threshold as is deemed advisable. The supply should be regulated if possible but satisfactory results can be obtained over quite a range of voltages.

The amplitude of oscillation is a marked function of these supply voltages; therefore, the relative distortion will also vary. Some compensation is possible by using one of the circuit variations.

Circuit Variations

There will be some quiescent current flowing in the nonlinear element if two equal resistors are used in the cathodes. These tubes are not balanced because the load resistor appears only in the plate circuit of stage two. This condition will be known as the unbalanced circuit.

If the cathode resistor R_{k2} is increased in value the d.c. voltages on the cathodes may be made equal and the nonlinear element will not carry d.c. current. This will be known as balanced circuit A and does not require any additional circuit components.

When the circuit is made symmetrical by using a dummy load resistor in the plate circuit of the first stage the quiescent current may be removed from the nonlinear element. This will be known as balanced circuit B. If this dummy load is bypassed for a.c. signals the gain is given by the first expression. If the dummy load is not bypassed the element R_{L1} must be included in calculation of the net amplifier gain in this manner:

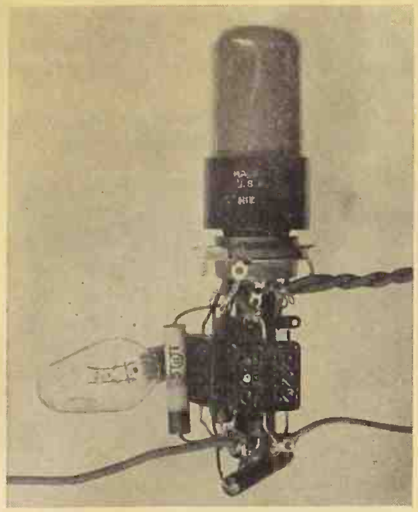
$$K' = \frac{A}{B + CR_s} \qquad (9)$$

where the constant A is the same as Eqt. (6), and B and C are given by:

$$B = (r_p + R_{L2})[r_p(R_{k1} + R_{k2}) + (\mu+1)R_{k1}R_{k2}] + r_pR_{k1}R_{k2}(\mu_2+1) + R_{L1}\{R_{k2}(r_p+R_{L2}) + R_{k1}[r_p+R_{L2}+(\mu_2+1)R_{k2}]\} \qquad (10)$$

$$C = [r_p + (\mu_1+1)R_{k1}][r_p+R_{L2}+(\mu_2+1)R_{k2}] + R_{L1}[r_p+R_{L2}+(\mu_2+1)R_{k2}] \qquad (11)$$

The basic unbalanced circuit causes the attenuation in the "π coupling" network to increase with increasing supply voltages. This tends to counteract



A 6SN7 oscillator of the unbalanced circuit form operating at 995 to 997 c.p.s.

the increase in tube amplification at higher supply voltages in much the same manner as the circuit of Fig. 7A. This unbalanced circuit offers this inherent compensation; however, where the greatest a.c. sensitivity is desired the amplifier should be balanced for equal d.c. voltages on the cathodes.

The variation of amplifier gain with respect to signal level has been examined for the unbalanced and two balanced circuit variations. Typical results are indicated in Fig. 8.

The basic unbalanced circuit is shown to be less responsive to supply voltage variations in the results given in Fig. 9.

The circuit presented is well suited to fixed frequency oscillators. The values required may first be calculated and then trimmed on a trial and error basis. A great deal of care is required to locate the frequency to less than 1 per-cent using fixed elements. These rules may prove helpful:

1—To raise the frequency pad additional resistance across R_1 or R_2 in the feedback circuit.

2—To lower the frequency pad additional capacitance across C_1 or C_2 in the feedback circuit.

3—The choice between the two resistors (or the two condensers) is made considering the present level of oscillation. When additional resistance is placed across R_1 or additional capacitance is placed across C_1 the signal level of steady oscillation will increase. If some adjustment of amplifier gain is provided this need not be considered.

A number of fixed frequencies may be obtained by suitable switching of fixed components in the feedback network. Both the resistance and the capacitance elements may be switched for frequency ranges greater than one decade. There is a relationship between the values selected; these general rules may help in adjusting these values.

1—Begin with the larger values of resistance and capacitance first.

2—Complete the selection of the set of elements (R's or C's) which are greater in number. Attempt to maintain nearly constant output signal amplitude for these values.

3—Complete the selection of the remaining set of elements. This set will generally be used as multipliers of the frequencies given by the first set.

A very useful audio oscillator may be constructed using two sets of element values for the feedback network. An example of the different frequencies which might be obtained is as follows:

The resistor elements are selected with a two pole, nine point switch to give these base frequencies: 10, 15, 20, 25, 30, 40, 50, 60 and 80 cycles per second. The condenser elements are selected with a two pole, three point switch to give these multiplier factors: Times 1, Times 10 and Times 100.

Continuously variable frequency oscillators may use ganged resistance or capacitance pairs with the alternate elements switched for different multiplier factors times the basic frequency range. Ganged resistors may be used in series with fixed resistors to minimize tracking errors—a frequency range of 3:1 may be suggested here. Variable resistor elements will cause the loading of the feedback network upon the amplifier to vary as the oscillator is tuned to different frequencies. Ganged condensers may be used for a range of 10:1 as tracking is much easier and the loading upon the amplifier is constant. If the minimum capacitance is to be 100 $\mu\mu$fd. the condenser pair would require a maximum of 1000 $\mu\mu$fd. per section for a decade of frequency variation. With care a smaller minimum capacitance might be used.

Advantages of this new oscillator are primarily in design and economics,

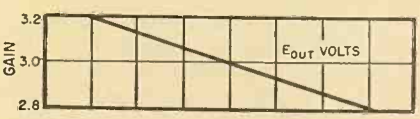


Fig. 13. Gain characteristics of the amplifier shown in Fig. 7A.

where cost, weight and space are important. Fixed frequency applications include bridge test sets, intermodulation test sets, modulation signals for signal generators, telemetering systems, tone generators, and simple musical instruments.

Semi-variable and continuously variable frequency oscillator applications are many and varied. This oscillator will best serve in those applications where it has not been possible to use other circuits because of the cost, space and weight factors.

News Briefs

(*Continued from page* 23)

very low dielectric loss and good punching characteristics.

PB 99220, Low-loss Arc-resistant Paper Laminates, sells for 50¢ a copy. Orders should be addressed to the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C., and should be accompanied by check or money order payable to the Treasurer of the United States.

Welding Control Equipment

Westinghouse Electric Corporation now has a booklet describing in detail its electronic resistance welding control equipment, both synchronous and non-synchronous.

Basic equipment in the control circuit includes a sequence-weld-timer panel and a means to fire the ignition tubes in the power circuit. Auxiliary control panels can be readily added to meet specific job requirements.

A copy of this booklet, B-4309, may be obtained by writing *Westinghouse Electric Corporation*, P. O. Box 2099, Pittsburgh 30, Pa.

TV Tube Booklet

The Radio Tube Division of *Sylvania Electric Products Inc.*, Emporium, Pa., is offering a new 20-page booklet providing television picture tube and general purpose cathode ray tube characteristics; replacement tube data; base diagrams; suggestions for tube handling; and a concise description of cathode-ray oscilloscope use in television servicing.

Information contained covers 165 tube types with faces ranging from two to twenty inches maximum dimension utilizing electrostatic or magnetic deflection systems. A nomenclature chart explains the meaning of type number letters and figures and applications for different types of tube screens.

The booklet is available through Authorized *Sylvania* Distributors.

Geiger-Muller Counter

The nature, construction, and use of the Geiger-Muller Counter is presented in a new booklet just published by the National Bureau of Standards.

In addition to the treatment of the counter itself, the booklet discusses methods of detecting counter pulses, applications of counters to quantitative measurements, proportional counters, and the preparation and filling of Geiger-Muller Counters.

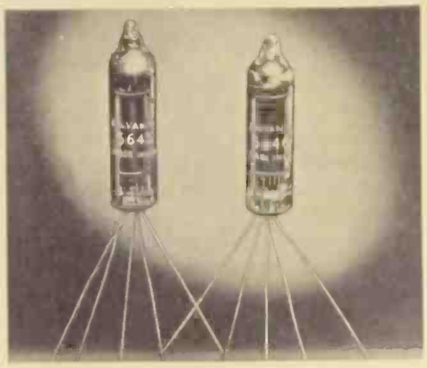
Circular 490, The Geiger-Muller Counter, may be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., 20¢ a copy.

NEW TUBES

SYLVANIA TUBES

Subminiature Tubes

Two additional *Sylvania* subminiature tube types, a medium-mu triode



and a high-mu triode, are now available. Type 5645 medium-mu triode is a T-2 suitable for Class A amplifier applications. Under typical operating conditions the tube will have a transconductance of 2700 micromhos and an amplification factor of 20. Maximum rated plate dissipation is 1 watt and plate resistance is 7400 ohms.

Type 5646 high-mu triode is a T-2 suitable for Class A amplifier or resistance coupled amplifier applications. Under typical operating conditions the tube will have a transconductance of 2400 micromhos, an amplification factor of 70, and a plate resistance of 29,000 ohms. Maximum rated plate dissipation is 0.3 watt.

Both subminiature types have 6.3 volt, 150 milliampere heaters, and flexible leads for direct wiring to circuit.

Beam Power Amplifier

A high-perveance beam power amplifier designed for use as a horizontal

deflection amplifier in high efficiency deflection circuits for television receivers is another of the many new tubes announced by *Sylvania*.

Features of type 6AU5GT include: low-mu, high plate current at low plate voltage and high operating ratio of plate current to number 2 grid current. The tube has a 6.3 volt, 1.25 ampere heater. Under typical operating conditions, it will have trans-conductance of approximately 6000 micromhos; mu approximately 5.9. The maximum plate dissipation is 8 watts; peak positive pulse plate voltage is 4500 volts; and maximum d.c. plate voltage is 450 volts.

Picture Tubes

An 8½-inch, all-glass, direct-view television tube with electrostatic focus and deflection for use in television sets designed for 7-inch viewing tubes is now available through authorized *Sylvania* distributors.

The tube, type 8BP4, is interchangeable with type 7JP4 and offers the advantages of 50% increase in useful area. Deflection sensitivity is provided for full scan in circuits designed for full scan

with 7-inch tube types. Operating voltages include: 6000 on number two anode; 1620 to 2400 on number one anode; and zero to —72 to —168 volts on number one grid.

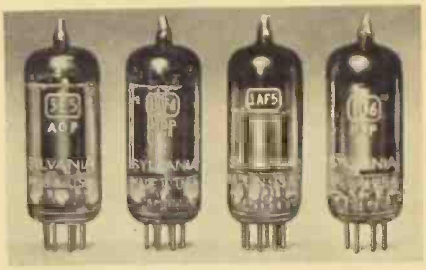
Type 16LP4 is a 16-inch, all-glass, direct-view television tube supplied with an external conductive coating which acts as a filter capacitance when grounded. The tube is supplied with neutral or clear face plates and employs an ion trap gun for use with external magnet.

Typical 16LP4 picture tube operating conditions include: maximum usable face diameter, 15⅞ inches; heater volts, 6.3; heater current, 0.6 amperes; focus coil current, 110 milliamperes; ion trap magnet current, 120 milliamperes; number 1 grid volts (for cut-off) -33 to -77; number 2 grid volts, 300; and anode volts, 12,000.

Miniature Portable Radio Tubes

A new line of miniature radio tubes for portable radio receivers, said to be the first announced in this country in more than ten years in which filament current has been reduced below 40 milliamperes per tube, is now being offered by *Sylvania*.

The new tubes require only 25 milliamperes per tube which means that "A"



batteries used in portable receivers will last approximately three times as long as with previous tubes. All are supplied with 7-pin miniature button bases.

Types included are: 1U6, a heptode converter with oscillator anode as a separate element; type 1AF5, a diode pentode; type 1AF4, a sharp cutoff r.f. pentode; and type 3E5, a beam power output tube. Power required for a complement of the new tubes in a typical battery-operated superheterodyne is only 2.1 watts. They will also operate satisfactorily over a range of 1.4 volts to 1.1 volts.

Miniature Rectifier Tubes

The Radio Tube Division, *Sylvania Electric Products Inc.*, 500 Fifth Ave., New York 18, N. Y., has announced a miniature high voltage half wave rectifier designed for television receiver pulse rectifying systems and voltage doubler circuits for magnetically deflected 10 and 12-inch viewing tubes.

The tube, type 1V2, has a peak inverse plate voltage of 7500, a peak plate current of 10 milliamperes, and an av-

erage plate current of .5 milliamperes.

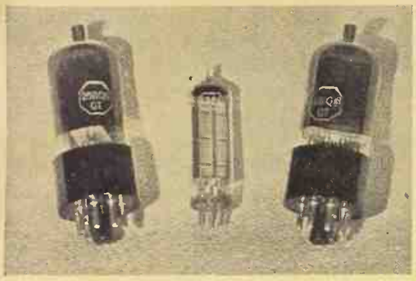
Type 1X2 is a double ended miniature high voltage rectifier tube designed for use with r.f., fly-back, and 60-cycle types of power supply for television picture tube anodes. This type is for use in power supplies where voltages up to

15,000 volts d.c. are required. The 1X2 is mounted in a T-6½ bulb and has an over-all height of 2 11/16 inches.

RECEIVING TUBES

Three new receiving tubes designed mainly for television receivers have been added to *General Electric's* tube production lines.

The 6AS5 is a beam-power amplifier of miniature construction intended for use as the audio power-output tube in television receivers and small radio receivers. When operating Class A1, with



a plate voltage of 150 volts and an input signal of 8.5 volts peak, 2.2 watts of output power can be realized with 10 per-cent distortion.

The 6BA6-GT and 25BQ6-GT are beam-power amplifier tubes designed to withstand high-surge plate voltages for short periods of time. These tubes are intended for use as horizontal-deflection amplifiers in television receivers. Maximum ratings of the tubes include a plate dissipation of 10.9 watts, a plate current of 100 milliamperes, and a peak positive surge plate voltage of 5000 volts.

Further information may be obtained from the Tube Divisions, Schenectady, New York.

RCA TUBES

Deflection Amplifier Tubes

Two new deflection amplifier tubes designed particularly for use with the

new 16-inch "metal" picture tube 16GP4 have been announced by the Tube Department of *RCA*, Harrison, N. J.

The 6CD6-G shown is a high-perve-

(*Continued on page* 29)

Video Amp. Design

(*Continued from page* 18)

ratio C_2/C_1 is designated by M, the values of R_L and L_c will be modified by:

$$R_L' = \sqrt{MR_L/2}\ ;\ L_c' = ML_c/2$$

where R_L' and L_c' are the new values of load resistance and peaking inductance respectively. The curves of Fig. 7 show amplitude and time delay characteristics of the series peaking network, for the case where $C_2 = 2C_1$.

Two Section Filter Coupling

It is frequently possible to secure uniform response and an improved figure of merit with a four terminal coupling network designed on the basis of a low-pass filter. Although such a network requires rather carefully controlled distribution of capacitances for best results, it does provide a gain factor approaching 2 over the uncompensated case, and has the further advantage that two output terminals are available (Fig. 6B).

The filter structure consists of a T type constant K full section, with mid-shunt terminated half-sections at each end. The full T section is composed of C_2 and the two series impedances $L_1/2$ and $L_2/2$. The load resistance and the cut-off frequency of the filter as a whole is determined by the capacitance C_2. The cut-off frequencies of the end half-sections are determined by capacitances C_1 and C_3. In order to prevent an unwanted rise in the image impedance of the input half-section, it is usually necessary to add resistive damping across L_1. The value of the damping resistor R_d is best determined by experiment, and will generally lie in the vicinity of 5 to 6 times the load resistance.

Due to the larger number of reactive elements, this type of coupling network shows a somewhat greater deterioration of phase characteristics about the top frequency. In order, therefore, to obtain sufficiently uniform response, it is sometimes necessary to design for a higher top frequency than would otherwise be the case. For optimum results, the terminating capacitances C_1 and C_3 should be nearly equal to one half the center capacitance C_2.

$$C_t = C_1 + C_2 + C_3 \quad ; \quad C_2 \cong 2C_1 \text{ or } 2C_3$$

$$R_1 = 1/\pi F_{max} C_t$$

$$L_1 = \left(\frac{1}{2} + \frac{C_1}{C_2}\right)\left(\frac{R_1}{2\pi F_{max}}\right)$$

$$L_2 = \left(\frac{1}{2} + \frac{C_3}{C_2}\right)\left(\frac{R_1}{2\pi F_{max}}\right)$$

Figure of merit $\cong$ 2.0

Case 3. Modified Shunt-Series Peaking

This type of coupling network provides a large figure of merit and at the same time quite uniform phase and amplitude characteristics. It is similar to the two section filter coupling described above, except that the terminating half-section is of the mid-series type. C_3 of Fig. 10 is therefore missing. The shunt-series network may afford a slightly better gain factor than the network of Fig. 6B, and the deterioration of phase response about the top frequency is not so great (Fig. 9A).

The accompanying Table II, taken from a report by O. H. Schade of *RCA* Laboratories gives the information necessary to design a shunt-series peaking network for capacitance ratios of 0.3, 0.5, and 1.0. Since the characteristics of the network are comparatively insensitive to this ratio, it is possible to interpolate between the values given with a fair assurance of success.

Of the six circuit values in the network, only two need be determined in order to completely design the interstage for a given bandwidth or gain. As the capacitances C_1 and C_2 are presumed fixed and reduced to their practical minimums, either the bandwidth or the gain may be used as a design parameter. Once the capacitances have been evaluated, the remaining parameters may be determined as a function of the chosen parameter directly from the table. The necessary design equations are:

$$C_t = C_1 + C_2 \; ; \; R_0 = 1/\pi F_c C_t$$

$$L_1 = R_1^2 \, C_t \; ; \; L_2 = 0.5 \, L_1$$

where F_c is the cutoff frequency of the network and R_o is the load resistance design constant.

The general procedure in designing a network of this type is to determine first the capacitances C_1 and C_2, the capacitance ratio C_1/C_2 and the total capacitance C_t. The ratio of the network cut-off frequency F_c to the top frequency of correction F_{max} for the existing capacitance ratio is determined from Table II (col. 4). The resulting value of F_c is then used to calculate the load resistance design constant R_o. The actual load resistance R_L can then be obtained from column 3 and the values of L_1, L_2, and R_d determined.

Where the video amplifier provides a fairly large gain reserve, it is sometimes possible to make use of selective degenerative feedback to compensate for discrepancies in the over-all high frequency characteristic. One simple method of accomplishing this is shown in Fig. 9B.

The resistance-capacitance network in the cathode circuit of the tube has a short time constant, and provides selective degeneration. C_k is a small mica capacitor, seldom larger than 1000 $\mu\mu$fd.

The voltage gain of the stage at low and medium frequencies where C_k is not effective is given by:

$$\overline{A_L} = \frac{G_m R_L}{1 + G_m R_k}$$

and is, of course, always smaller than the stage gain were R_k omitted. The gain reduction due to R_k is:

$$20 \log \, (1 + G_m R_k) \text{ decibels}$$

This equation indicates the maximum degree of amplitude compensation obtainable with cathode peaking for a fixed value of R_k.

Table II. Table for determining the ratio of the network cut-off frequency to the top frequency of correction for the existing capacitance ratio.

1 C_1/C_2	2 R_d/R_p	3 R_1/R_o	4 F_{max}/F_c	5 Figure of merit
0.3	5.66	1.088	0.96	2.1
0.5	5.66	1.075	0.85	1.8
1.0	18.90	1.00	0.76	1.5

Tube Type	G_m	F′	F″
6AG5	5000	98	232
6AH6	9000	117	238
6AK5	5100	119	242
6AU6	5200	79	157
6AC7	9000	90	184
6AG7	11000	85	177
7B5 (6K6GT)	2300	24	47
7C5 (6V6GT)	4100	35	71
6L6G	6000	40	84

Table I. The gain-bandwidth factor for some of the more common tubes.

At the high frequency end of the video band, the shunting reactance of C_k causes a decrease in the effective cathode circuit impedance, and results in an increased voltage gain. At the same time, the a.c. plate-cathode voltage is shifted in phase in a direction such as to partially counteract the phase shift produced by the plate circuit time constant $C_t \, R_L$. The gain at high frequencies is given by:

$$A_h = \frac{G_m R_L}{1 + (G_m R_k/y)}$$

where $y = [(2\pi f C_k R_k)^2 + 1]^{1/2}$ and the added phase shift by:

$$-\tan^{-1}\left(\frac{1}{2\pi f \, C_k R_k}\right)$$

It is possible to effect nearly complete high frequency phase and amplitude correction with cathode peaking by making the plate circuit and cathode circuit time constants equal, that is:

$$R_L C_t = R_k C_k$$

The figure of merit of a stage using cathode peaking only is generally not nearly so good as will be obtained with the two and four terminal networks described. Cathode peaking is, however, useful in small and controlled amounts to supplement compensation obtained by other means.

The high frequency characteristics of a well designed television video amplifier should be such that the maximum variation in over-all delay time at 4.0 megacycles will not exceed about 0.08 microseconds. This requires that the total phase deviation up to the highest frequency of importance be not greater than 57° per stage in a two stage amplifier. The over-all amplitude characteristic should be flat to within ± 1 decibel or so up to the top frequency of correction.

The schematic diagram of Fig. 8 shows a typical two stage television video channel incorporating high and low frequency compensation, while Fig. 10 indicates its time delay and amplitude characteristics.

TECHNICAL BOOKS

"ACOUSTICAL DESIGNING IN ARCHITECTURE" by Vern O. Knudsen and Cyril M. Harris. Published by *John Wiley & Sons, Inc.*, 440 Fourth Ave., New York 16, N. Y. 457 pages. $7.50.

Architects, builders, acoustical engineers, teachers and students of architectural acoustics will find this book extremely valuable as it covers the entire field of acoustical design in architecture. Practical applications are emphasized and includes design of auditoriums, theaters, school buildings, homes, churches, radio, television and sound-recording studios.

The contents is divided into two sections: (1) a consideration of the general principles and procedures which form a basis for all acoustical designing, and (2) a description of specific applications of these principles and procedures.

Complicated mathematical formulas are translated into physical explanations, and charts and nomographs are given for many formulas. Comprehensive tables give pertinent data on sound-absorptive materials and sound-insulative structures.

This volume will serve as a practical guide to good acoustical designing as well as a handbook in the solution of problems encountered.

"ADVANCES IN ELECTRONICS" Volume II, Edited by L. Marton, National Bureau of Standards. Published by *Academic Press Inc.*, 125 East 23rd St., New York, N. Y. 378 pages. $7.60.

Because of the favorable reception given the first volume of "Advances in Electronics," it was decided that a yearbook of this kind would receive wide acclaim. This second volume covering the latest advances in the field of engineering electronics includes a large contribution from European authors, whereas the first volume was written in its entirety by American authors.

Contributors to this second volume are: Donald K. Coles, *Westinghouse Research Labs;* H. Fröhlich, University of Liverpool, G. F. J. Garlick, University of Paris, France; Gunnar Hok, University of Michigan; G. Liebmann, *Associated Electrical Industries*, Ltd., Berkshire, England; Hilary Moss, *Electronic Tubes Ltd.*, Reading, England; George T. Rado, Naval Research Laboratory; and J. H. Simpson, National Research Council of Canada.

The editors and publishers have expressed a hope that this text will be a definite contribution to the free and international exchange of ideas among scientists.

New Products

(*Continued from page* 21)

tude modulation system is substantially flat from 30 cycles to well above 100 kc. A monitoring meter is used to standardize the output level of the signal generator in order to make the mutual inductance r.f. attenuator direct reading over the range from 0.1 microvolts to 0.2 volts.

The instrument is self-contained, with power supply, and is designed for use on 115 volts, 60 cycles. Complete details may be obtained on request.

COIL WINDER

A manually-operated coil winding machine designed for the production of paper-insulated coils in multiple or "stick" form is announced by *Universal Winding Company*, Providence, Rhode Island.

The No. 108 Winder was designed to supplement the automatic type of machine with a low-cost machine suitable for the predominantly "job shop" type of market characteristic of the ever-changing electrical and electronic parts industries. The machine can be adjusted quickly to accommodate changing requirements of wire size, coil length and diameter.

Power is supplied by a ½ h.p. a.c. constant speed motor driving through an adjustable-sheave speed controller to a multiple-disc friction clutch attached directly to the spindle. Speed range is 400 to 2200 r.p.m.

VOLTMETER

A new model of the Mini-Volt Voltmeter manufactured by *Industrial Devices, Inc.*, Edgewater, N. J., which features an expanded scale centered on the common 110 and 220 line voltages, is announced.

The new adaptation of this voltmeter is known as Model 410A and is accurate to within 2 volts at 110 volts a.c. Practically burnout proof operation is assured by the glow-lamp indicator which is guaranteed for 25,000 hours' operation minimum. 12" flexible test leads are tipped with heavily insulated test prods assuring user maximum safety.

The Mini-Volt can check voltage, a.c. or d.c.; check for continuity; check blown fuses; locate grounded components; warn of live wires and scores of other possible troublesome occurrences known to electricians, refrigeration servicemen, maintenance men, and anyone who works with electrical circuits.

PRECISION POTENTIOMETERS

Technology Instrument Corporation, 1058 Main St., Waltham, Mass., is now manufacturing an improved version of their Type RV3 precision potentiometers. Available in two types, the Type RV3-8 (8 watts) and the Type RV3-12 (12 watts), these potentiometers have been redesigned to include bronze bushings for the rotor shaft, tapped mounting inserts, rotor take-off slip ring, dustproof construction with Bakelite cover fastened with screws, and molded parts of low-loss mica-filled Bakelite.

The Type RV3 potentiometers can be supplied with a wide variety of non-standard features to meet special requirements, and the manufacturer has published an established price scale covering such non-standard features. Complete information may be obtained by writing the company direct.

New Tubes

(*Continued from page* 27)

ance, beam power amplifier featuring low mu-factor, high plate current at low plate voltage, and a high operating ratio of plate current to grid No. 2 current. In suitable horizontal-deflecting circuits, only one 6CD6-G is required to deflect fully any directly viewed kinescope having a deflection angle up to 70° and operating at an anode voltage up to 14 kilovolts.

The 6S4 is a high-perveance, medium-mu triode of the 9-pin miniature type. In suitable vertical-deflecting circuits, the 6S4 will deflect fully a 16GP4 or any other similar kinescope having a deflection angle up to 70° and operating at an anode voltage up to 14 kilovolts.

Sharp-Cutoff Pentode

The 6AS6 announced by *RCA* is a sharp-cutoff pentode of the 7-pin miniature type. Designed so that grid No. 1

and grid No. 3 can each be used as independent control electrodes, the 6AS6 is especially useful in gated amplifier circuits, delay circuits, gain-controlled amplifiers, and mixer circuits.

This tube is also suitable for use as an r.f. amplifier at frequencies up to about 400 megacycles per second.

Surface Wave

(*Continued from page* 11)

several months and had a rather thick corrosion layer. In this state the measured loss was 3 db. Most of the corrosion was then removed by sandpaper. The result was an increase of the loss to about 3.6 db. The expected loss for a wire with perfectly clean surface is much higher. Then we applied several coats of polystyrene solution by means of a brush. The loss went down more and more till a minimum of 1.7 db. was reached. Further increase of the dielectric layer gave a slight increase of the loss by about 0.2 db. The thickness of the dielectric coat varied more than 1:10 along the wire, thus it was not possible to measure the average thickness. This test indicated that the wave is not sensitive to inhomogeneities of the guide.

Finally, a test was conducted with a 600 ft. line which showed that the support of the wire is not a serious problem. The horns used were the same as in the other experiments (opening of 13"). The wire was an enameled wire of 3.2 mm. thickness. The enamel coat was 25/100 mm. thick and the frequency was 1600 mc. The line was stretched along the slope of a hill with the distance from the ground varying between 4 and 8 feet. The wire was supported at intervals of about 80 feet by waxed strings. The calculated loss was 4.5 db. and the measured loss 5 db. For comparison, an RG-8/U cable of the same length has an attenuation of 70 db. The supports and the bends in the wire produced by the supports had no measurable effect in the attentuation.

Fig. 4 shows the calculated attentuation in db. per mile for a line with 2 cm. diameter and a polystyrene coat of ½ mm. in the frequency range from 300 to 1000 mc. In addition to this loss which is proportional to the length we get the loss of the launching devices, which could be kept below 2 db. Considering the wide frequency band which could be transmitted by such a line, the loss is very small.

The question of the effect of weather conditions can be answered only partially and only as far as rain is concerned. An increase of the attenuation was observed only if the wire was covered with rows of big raindrops. The maximum increase measured so far for 1600 mc. was less than 1.5 db. A water film did not increase attentuation. It enlarges the thickness of the dielectric layer and behaves like a good dielectric because the ratio of the power factor to (ε—1) which determines the losses is very small. The effect of snow and ice will certainly be more serious and it may be necessary to prevent the settling down of snow by shaking the wire or by electric heating.

It is a pleasure to acknowledge the close cooperation in Coles Signal Laboratory which made it possible to obtain the present results in a relatively short time. Thanks are due to Mr. J. Hessel and R. Lacy, Chief and Assistant Chief, Radio Communication Branch, to Mr. J. Egli, Chief, Radio Relay & Microwave Section, and especially to the group who cooperated in the measurements, Mr. C. Sharp and L. Battersby and particularly Mr. A. Meyerhoff, who checked all calculations and improved the English of the paper which otherwise would have been much worse.

Microwave Trans.

(*Continued from page* 9)

$$A_{av} = \frac{1}{T}\int f(t)\,dt \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

where T is the time required for one period (shown in Fig. 9).

The integral is merely equal to the pulse area. For the rectangular pulse shown in Fig. 9, this area is equal to the amplitude times the pulse width. The relation between average power and peak power for a rectangular pulse is therefore equal to:

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Thus it is seen that if the period is equal to 1000 microseconds, and the pulse width equal to 1 microsecond, the peak power is equal to 1000 times the average power.

In a lighthouse tube the maximum power rating, for a given frequency, is primarily limited by the maximum allowable plate temperature or the average power dissipated. As a result, a peak power thousands of times greater than the maximum average power rating of the tube can be effected. Fig. 6 is a series of typical power output versus frequency curves for different plate voltages for a 2C43 tube modulated by the pulse series shown in Fig. 9.

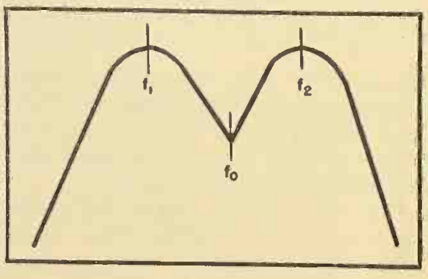
In pulse modulating systems where the timing of the pulse carries the intelligence an effect known as "jitter" may be introduced into the microwave oscillator. The r.f. oscillator is normally keyed "on" at some time during the rise time of the applied pulse — the exact instant of keying depends upon the tube characteristics, circuit constants, and random noise fluctuations. If all of these conditions remained the same, the r.f. oscillation would start at the same relative instant for each succeeding pulse. However, the random fluctuations of the noise signal may cause the tube to fire either slightly before or after the normal repetition rate. Since the timing of the pulse carries the intelligence, this shift in pulse position appears as noise or "jitter" in the demodulator of the system.

This effect can be minimized through the use of a "catalyzer".[6, 7] A "catalyzer", as many readers may remember from their chemistry courses, is an element which aids a reaction but does not react itself. In this case the catalyzer consists of a small oscillator which injects a signal of approximately the same frequency as that of the r.f. oscillator into the microwave cavity. The catalyzer oscillator can be a conventional push-pull oscillator operating in the 300 to 500 megacycle band with a crystal multiplier used to obtain the fifth or sixth harmonic of this signal. The output of the crystal multiplier is then fed to the cavity through a probe or loop. It has been found that the use of a catalyzer appreciably improves the signal to noise ratio of the microwave oscillator. The reason for this effect is probably that the "catalyzer" signal is sufficiently large to "overpower" the noise. Thus random changes in noise have little effect on the oscillator.

There are a number of parameters that affect the frequency of a lighthouse oscillator. Among these effects are the change in cavity length due to a variation in temperature (as a result of expansion and contraction of conductors), change in plate or filament voltage (affecting electron transit time), and cathode to grid interelectrode capacitance (feedback effects). In a typical commercial transmitter, it has been found that the cumulation of all these effects (using "self compensated" temperature conductors in the coaxial lines) results in a frequency stability of about 0.05 per-cent at 2000 mc. In this latter unit no attempt had been made to compensate for the voltage variations through the use of a regulated plate supply.

Where a higher degree of frequency stability is required, some method of automatic frequency control (a.f.c.) must be employed. Heretofore it has been found that the best method of frequency control is an electromechanical system in which a variation in frequency from the desired value is detected by a discriminator, whose output actuates a motor which drives a tuning device until the desired frequency is

Fig. 10. Frequency response of two reference cavities and desired resonant frequency of triode resonant line.



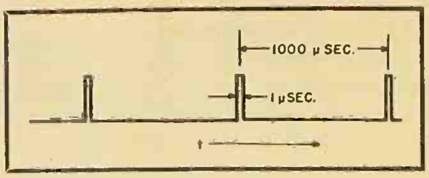


Fig. 9. Typical pulse series.

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Two typical a.f.c. circuits are shown in Figs. 5 and 8. The a.f.c. circuit shown in Fig. 5 uses two reference cavities, one resonant at f_1 and the other at f_2. The desired resonant frequency of the triode line, f_o, is equal to $(f_1 + f_2)/2$. These relations are plotted in Fig. 10.

A small amount of energy from the triode line is picked up by loops shown in Fig. 5 and injected into the two reference cavities. When the frequency of this energy is exactly equal to f_o, the outputs of the two cavities are equal and no discriminator voltage is developed across the grids of V_1 and V_2. V_1 and V_2 are thyratrons which are normally cut off. If the energy obtained from the triode line is not equal to f_o, a discriminator voltage will be developed across the grid of V_1 or V_2 (depending upon whether frequency is above or below f_o) causing one to conduct and operate the split phase motor. This motor will operate the reset control in the direction that will correct for the "off frequency" variation. An a.c. voltage is applied to the plates of the thyratrons, so that they will be extinguished when the discriminator voltage reduces to zero. The cavities used can be made highly reliable by using temperature compensated metals such as Invar or using a temperature controlled oven.

The a.f.c. circuit shown in Fig. 8 is a simplified version of the previous circuit. In this case a single cavity and thyratron are used in conjunction with a motor driven trimmer condenser. The effect of a condenser at the end of a line was discussed in the article, "Microwave Components". The action of this circuit is as follows: The existence of a frequency other than the desired value in the cavity provides a discriminator voltage which fires the thyratron. This operates the motor which drives the condenser rotor. The plate of the condenser turns very slowly, first increasing and then decreasing resonant line frequency. The plate will turn until the discriminator voltage drops to zero. The motor speed is made sufficiently low compared to the time required for the discriminator to extinguish the thyratron, that the condenser will not "overshoot" the desired frequency.

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Custom Sound

(*Continued from page* 5)

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An interesting note is the fact that these speakers and baffles were so efficient in projecting the sound to the grandstand that additional monitor speakers had to be installed backstage in the wings to allow the stage managers and cast to follow the continuity. These were *Jensen* eight inch heavy-duty loudspeakers mounted in weatherproof metal projectors. Each of these monitors had an adjustable "L" pad control to allow the stage managers to set the level to suit their surrounding noise level.

For use before show time and for fill-in in case of rain or accident, a record turntable and its preamplifier was incorporated in the console control cabinet. This unit could be plugged into the patch panel at the output terminal of the console and automatically replace the console program with the recorded program.

In addition to the stage sound system, a second independent amplifier and set of four 24" re-entrant drive unit loudspeakers were installed atop the grandstand roof to cover the outside plaza in front of the grandstand entrances. This system could be fed from the console, the turntable, the Celesta-Chime or from a microphone position located in one of the grandstand entrances. The primary purpose of this system was to advertise the pageant and also to furnish music to the plaza between shows.

The *Deagan* 25 note Celesta-Chime was chosen for this installation because it was found to be the only electronically amplified chime that could be played in tune with an orchestra. The chime rods are tuned harmonically which eliminates the "wow" and "out of tune" effect that is usually associated with chimes and bells. In some instances where the Celesta-Chime was played solo, full chords were used. No other instrument was found that was suitable for such usage. A special low "G" chime was built by *Deagan* to furnish the deep tolling bell effect that was desired for the Lincoln Funeral Train scene. As explained above, these chimes were reproduced through the control console and entire system and could be controlled the same as any microphone.

Since all the sound and control equipment had to be integrated with the entire operation of the show, the design and installation of the intercommunication, buzzer, cue and railroad signal systems was undertaken along with the sound system. The intercommunication system was in two stages; the first stage consisted of an electronically amplified 20 station master unit located in the director's booth connecting with remote stations back stage, ticket booths in the plaza, the orchestra pit, the equipment tents north and south of the stage, the electricians' switchboard room and the sound man. The importance of this intercom system can be realized when it is understood that it was the only two-way communication that the director had with the show from his remote position. It was used to check on equipment, cue in some of the vehicles, instruct the orchestra conductor and deliver the light cues, of which there were 62, to the electricians under the stands. As a supplementary system, there was a battery operated telephone type intercom system with three master stations. One was in the director's booth, one at the sound control console (visible at the extreme end of the control console, Fig. 3), and one in the orchestra pit. This system was used mainly between the orchestra conductor and the sound man.

The buzzers, music cue lights and the railroad signals were all controlled from a small custom-built consolette located in the director's booth. Buzzers were located in all dressing rooms backstage, in the wings and at the extreme ends of the stage extension wings. Each buzzer could be controlled independently or all could be sounded at the same time. The music cue lights were located on the orchestra conductor's stand and consisted of one flashing white warning light and a second steady amber "go" light. The railroad signals were located at the ends of the stage wings and were used to cue the locomotives on stage. These signals had to be visible to engineers on any of the four switch tracks in either the north or south yards both in daylight and dark. The operation of the controls for these buzzers, lights and signals had to be simple and foolproof as a mis-cue during the show could cause a serious accident. Some of the trains that crossed the stage traveled 30 to 40 miles per hour, so it is easy to see that an accident would be hard to prevent in case of a mis-cue.

During the rehearsals before the show opened, a temporary talk-back system was installed in the grandstand to enable the directors to speak to the people on stage and in the equipment and railroad yards. This system utilized the two re-entrant projectors that were on either end of the grandstand roof and additional projectors that were placed in the stands and aimed toward the stage.

Much of the success of the entire installation was due to the intelligent discussion of the entire staging problem by Helen Geraghty, the director of the pageant. By talking over the entire show, and not just the sound system, a clear picture of the entire operation was established before the designing was even undertaken. It must be emphasized here that too many times a sound system of this type will be installed as just a P.A. job and will fail to do the job because "new and unusual" conditions pop up that had not been thought out. One of the reasons that this installation was so successful was that, not only was it a good sound reinforcing system, but a live part of the show. The two operators, who alternated at handling the mixing, worked from a complete technical script that listed all the dialogue, the music cues and the mixing cues. In all they had 260 separate sound cues to handle in one hour and eight minutes.

A good sound installation is noticed and appreciated by those who have to listen to it. Many compliments were received this past summer from musicians, directors and members of the industry who attended the show, and many favorable comments were voiced by members of the audience who had never before paid any attention to the sound systems that they had heard. In every large outdoor amphitheater, stadium or fairgrounds attention should be paid to all the details that contribute to or detract from the coverage and quality of the sound system installation so that high-quality sound systems can become the rule instead of the exception.

PHOTO CREDITS

6, 7, 8. . Federal Telecommunication Laboratories

10. . Official Department of Defense

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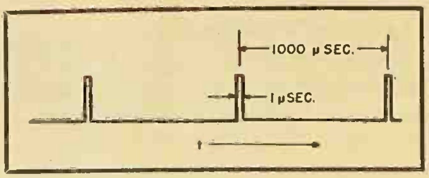

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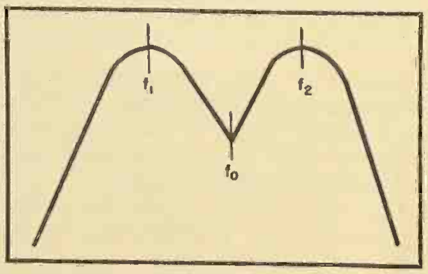

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In addition to the stage sound system, a second independent amplifier and set of four 24" re-entrant drive unit loudspeakers were installed atop the grandstand roof to cover the outside plaza in front of the grandstand entrances. This system could be fed from the console, the turntable, the Celesta-Chime or from a microphone position located in one of the grandstand entrances. The primary purpose of this system was to advertise the pageant and also to furnish music to the plaza between shows.

The *Deagan* 25 note Celesta-Chime was chosen for this installation because it was found to be the only electronically amplified chime that could be played in tune with an orchestra. The chime rods are tuned harmonically which eliminates the "wow" and "out of tune" effect that is usually associated with chimes and bells. In some instances where the Celesta-Chime was played solo, full chords were used. No other instrument was found that was suitable for such usage. A special low "G" chime was built by *Deagan* to furnish the deep tolling bell effect that was desired for the Lincoln Funeral Train scene. As explained above, these chimes were reproduced through the control console and entire system and could be controlled the same as any microphone.

Since all the sound and control equipment had to be integrated with the entire operation of the show, the design and installation of the intercommunication, buzzer, cue and railroad signal systems was undertaken along with the sound system. The intercommunication system was in two stages; the first stage consisted of an electronically amplified 20 station master unit located in the director's booth connecting with remote stations back stage, ticket booths in the plaza, the orchestra pit, the equipment tents north and south of the stage, the electricians' switchboard room and the sound man. The importance of this intercom system can be realized when it is understood that it was the only two-way communication that the director had with the show from his remote position. It was used to check on equipment, cue in some of the vehicles, instruct the orchestra conductor and deliver the light cues, of which there were 62, to the electricians under the stands. As a supplementary system, there was a battery operated telephone type intercom system with three master stations. One was in the director's booth, one at the sound control console (visible at the extreme end of the control console, Fig. 3), and one in the orchestra pit. This system was used mainly between the orchestra conductor and the sound man.

The buzzers, music cue lights and the railroad signals were all controlled from a small custom-built consolette located in the director's booth. Buzzers were located in all dressing rooms backstage, in the wings and at the extreme ends of the stage extension wings. Each buzzer could be controlled independently or all could be sounded at the same time. The music cue lights were located on the orchestra conductor's stand and consisted of one flashing white warning light and a second steady amber "go" light. The railroad signals were located at the ends of the stage wings and were used to cue the locomotives on stage. These signals had to be visible to engineers on any of the four switch tracks in either the north or south yards both in daylight and dark. The operation of the controls for these buzzers, lights and signals had to be simple and foolproof as a mis-cue during the show could cause a serious accident. Some of the trains that crossed the stage traveled 30 to 40 miles per hour, so it is easy to see that an accident would be hard to prevent in case of a mis-cue.

During the rehearsals before the show opened, a temporary talk-back system was installed in the grandstand to enable the directors to speak to the people on stage and in the equipment and railroad yards. This system utilized the two re-entrant projectors that were on either end of the grandstand roof and additional projectors that were placed in the stands and aimed toward the stage.

Much of the success of the entire installation was due to the intelligent discussion of the entire staging problem by Helen Geraghty, the director of the pageant. By talking over the entire show, and not just the sound system, a clear picture of the entire operation was established before the designing was even undertaken. It must be emphasized here that too many times a sound system of this type will be installed as just a P.A. job and will fail to do the job because "new and unusual" conditions pop up that had not been thought out. One of the reasons that this installation was so successful was that, not only was it a good sound reinforcing system, but a live part of the show. The two operators, who alternated at handling the mixing, worked from a complete technical script that listed all the dialogue, the music cues and the mixing cues. In all they had 260 separate sound cues to handle in one hour and eight minutes.

A good sound installation is noticed and appreciated by those who have to listen to it. Many compliments were received this past summer from musicians, directors and members of the industry who attended the show, and many favorable comments were voiced by members of the audience who had never before paid any attention to the sound systems that they had heard. In every large outdoor amphitheater, stadium or fairgrounds attention should be paid to all the details that contribute to or detract from the coverage and quality of the sound system installation so that high-quality sound systems can become the rule instead of the exception.

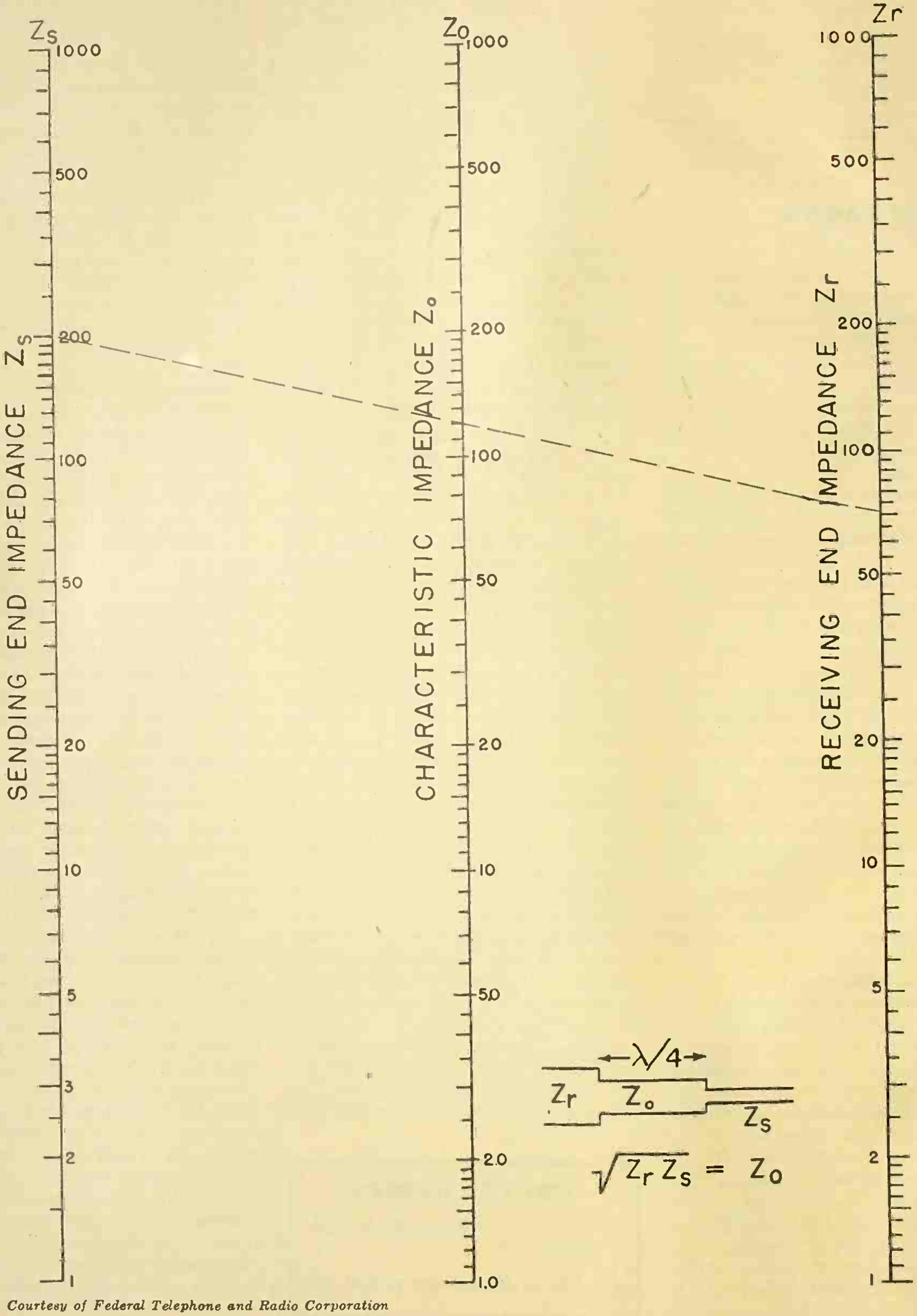
PHOTO CREDITS

6, 7, 8. . Federal Telecommunication Laboratories

10. . Official Department of Defense

QUARTER-WAVE MATCHING SECTION

This chart is used to obtain the surge impedance of a 1/4 λ matching section used as an impedance transformer from one real impedance to another.



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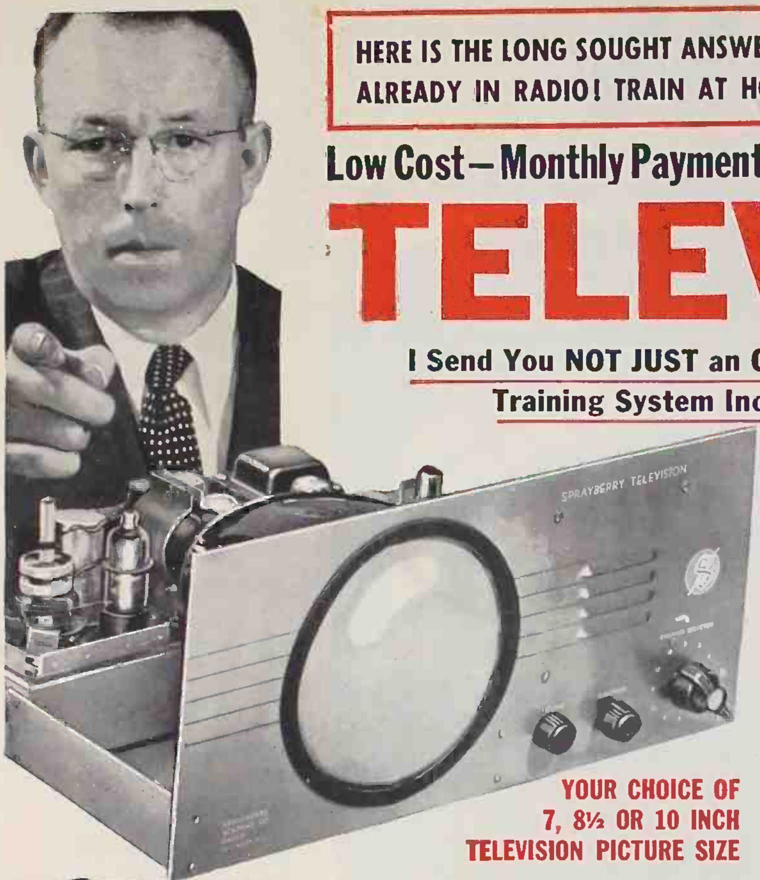

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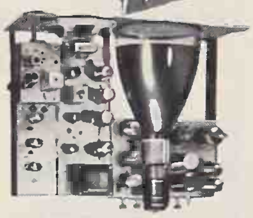

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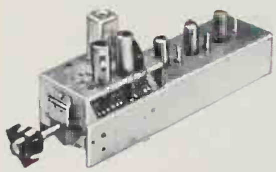
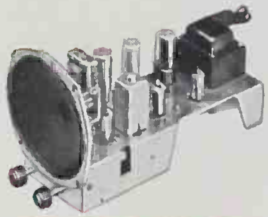


TV Tuner—I. F. Unit

Contains the RF amplified local oscillator, mixer and three stages of broad band IF amplification and the video second dector. The output constitutes the video signal and audio IF signal. For training, it is used to build and test video second detector, and stagger tuned IF amplifier obtaining 4.5 mc band pass. For TV servicing, it becomes a TV calibrator for IF alignment, substitute tuner, IF signal injector and second detector.



Video-Audio Amplifier Unit

Provides 4.5 mc IF ratio detector, low voltage power supply. For TV, it becomes the audio output, including speaker, video output and low voltage power supply for RF and IF stages. For training, it is used to build and test transformer type power supplies, audio, video, IF amplification and FM detection. For TV servicing, it is an audio signal tracer, IF signal tracer, video signal tracer and low voltage power supply.

Video Tube "Scope" Unit

Scope unit contains low and high voltage (6000 V.) power supply for independent operation. For television, it becomes the sync, vertical and horizontal sweep circuits and their power supplies. For training, it is used to build and test most TV power supply, deflection, sweep, oscillator, and sync circuits. For TV servicing, it is a video signal tracer and sweep signal analyzer as well as substitute high and low voltage power supplies.

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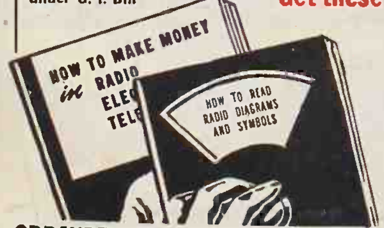

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