MIDLAND TELEVISION

INC.

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT
NO.
7

TELEVISION CAMERAS

LESSON
NO.
1

INVISIBLE FORCES

.....controlled by man.

During recent years, many majestic and awe-inspiring barriers have successfully challenged the flow of great rivers. Mighty dams, created by man, are holding back huge volumes of water, creating lakes or storage basins of pretentious proportions. Tunnels have been blasted and bored through mountains so that the stored-up water could be harnessed and made to create electric power.

The world rightfully acclaims the engineers and construction crews whose devotion to duty and heroism have made such structures possible. The public knows that the purpose of the dams is to conserve water, prevent floods, and to create electricity. But they often fail to acclaim another group of men whose outstanding accomplishments in utilizing this electric power have, in many instances, closely approached wizardry.

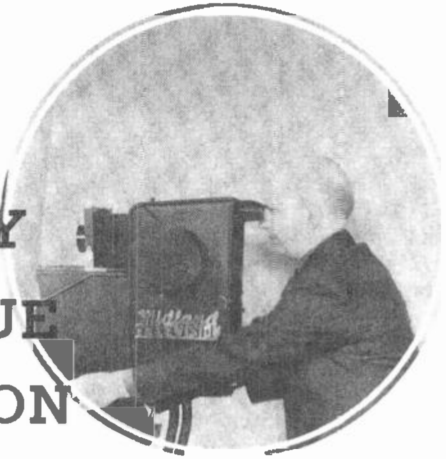
Water can be seen with the eye. Therefore, its action and flow can be followed closely. Electricity cannot be seen---yet man has exerted control over this invisible force in many uncanny ways.

The television camera, which you will study in this interesting lesson, presents an excellent example of delicate man-created control over electrical impulses. Humans give vent to exclamations of awe when they look upon mighty Boulder Dam. Yet those same humans, witnessing a television camera in action, will give no thought to the great amount of money and years of laborious research necessary to the perfection of the camera. Their eyes will be on the cameraman.

Now that YOU are about to reach the conclusion of your training, it is important that YOU fully realize that the perfection of television represents a gigantic, costly, and amazing accomplishment. The opportunity is ready for you, and its proportions are far greater than any towering dam.

Unit Seven

THE THEORY & TECHNIQUE of TELEVISION BROADCASTING



"This unit, consisting of ten or more lessons will be devoted to the study of the transmission of television pictures. In these lessons we shall attempt to give you sufficient information pertaining to the design, lay-out, maintenance, and operation of television transmitting stations, so that you will be qualified to fill the position of television operator at the conclusion of your training.

"The first six lessons in this unit will be devoted to a study of the equipment necessary and the procedure used in picking up a picture from a television studio, and the transmission of that picture through to the antenna system.

"The seventh lesson in this unit will cover the interesting features surrounding stage lighting and scenery design for television purposes. The eighth lesson will serve to supply you with the knowledge necessary in the transmission of motion picture film, since this phase will play an important part in television programs. The ninth and tenth lessons in this unit will cover two subjects not strictly devoted to the transmission of television pictures. The ninth lesson will cover the procedures and theories involved in the reproduction of large television images; that is, pictures 18 inches by 24 inches or larger. The tenth or last lesson in this unit will be devoted to a thorough study of facsimile. This is the transmission of still pictures by wire.

"Even though you do not intend to become associated with a television transmitting station, it is highly important that you be thoroughly familiar with the procedure involved. In this way you will be able to carry on to a more perfect degree in any television work".

Lesson One

TELEVISION CAMERAS

"In an earlier lesson you studied thoroughly the principles of scanning. However, in that lesson, only mechanical scanning was covered in detail. In this lesson you are going to become familiar with electronic scanning and the modern electronic television camera.

"Without the development of electronic scanning as it is today, there would be no successful commercial television. Therefore, the subjects covered in this lesson are extremely important, and you should treat them accordingly.

"The television camera, as it relates to picture transmission, occupies the same place as the microphone does in sound broadcasting. Since it is the pick-up device, the entire balance of the equipment involved would be useless without the camera. Therefore, you should accord it the real importance it deserves in your studies".

1. INTRODUCTION. The television camera is the first link in the chain between the televised scene and the observer of the reproduced image at the receiver. It is the television camera that resolves the required scene into a large number of elemental areas and generates an electrical impulse for each area with a magnitude corresponding to the intensity of the light reflected from that particular element. This process is known as scanning. You recall that scanning is necessary because an electrical communications channel can transmit but one electrical impulse at a time while the eye and its associated nerve channels is capable of transmitting to the brain a large number of impulses simultaneously.

In a preceding lesson you learned how scanning could be accomplished by means of the scanning disc. There were two types of scanning, the direct pickup and the flying spot system. In the direct pickup system the televised scene was under intense illumination. An image of the scene was focused on the scanning disc. A photocell was placed behind the disc. As the disc rotated, light reached the photocell progressively from the picture elements. In the flying spot system, the lights and photocell are interchanged. That is, the televised scene is in the dark and is scanned by a moving spot of light. The light reflected from the televised scene is picked up by banks of photocells arranged to collect all the light reflected from the scene.

The flying spot system was preferable because of the increased amplitude of the picture signal. Also the subject did not have to endure the intense illumination required with the direct pickup system. However, as far as flexibility and utility are concerned, the direct pickup system is best. The flying spot system is limited to studio scenes and motion picture film as the subject must be in the dark. Outdoor and indoor daylight scenes can not be televised.

Pictures with much less than 441-line resolution are not satisfactory as an entertainment medium. It is difficult mechanically to scan a picture with this resolution by means of a scanning disc. Resolution of 90 to 120 lines is the maximum that can be conveniently obtained with mechanical scanning.

Increasing the number of lines in the picture for a given diameter scanning disc means that the holes must be made proportionately smaller. It is very difficult mechanically to punch small square holes with the accuracy required for a scanning disc. This can be remedied by using a larger disc. When the disc size is increased, the power required to produce rotation increases very rapidly as the diameter of the disc is increased unless the disc rotates in a vacuum. A camera containing a high speed disc is difficult to move about the studio because of the inertia of the disc. You are familiar with the action of gyroscope toy tops and how they resist attempts to shift their axes of rotation.

From the foregoing paragraphs it is evident that cameras employing mechanical scanning means are very impractical as far as modern high-definition television pictures are concerned. A scanning process that is electrical in nature is required. Later in the lesson, two methods for electrical scanning will be described.

2. FACTORS CONTROLLING PICTURE SIGNAL AMPLITUDE. As the picture detail is increased, the amplitude of the picture signal generated is reduced proportionately with conventional methods of scanning. The total light energy collected from a picture is the same regardless of the number of picture elements into which the picture is resolved. However, the amount of light energy per picture element varies inversely with the number of picture elements. Therefore, the picture signal amplitude decreases as the number of picture elements increases. The number of picture elements for a picture varies directly with the square of the number of lines.

Another factor which controls the amplitude of the picture signal is the rate at which scanning takes place. The scanning rate depends on two factors, the number of lines in the picture and the number of frames transmitted per second. Let us consider the relation between the scanning rate and the number of lines when the frame frequency remains constant. For example let us consider a picture one inch square and scanned with a hundred-line definition at the rate of 20 pictures per second. Such a picture contains 10,000 picture elements. Therefore, the scanning aperture passes over 10,000 picture elements for each frame. If the picture is scanned with 200 lines, the scanning aperture passes over 40,000 picture elements per frame. If the frame frequency is the same in the two cases, the scanning aperture passes over the picture elements 4 times as fast for the 200-line picture as it did for the 100-line picture. This means that ¼ as much light energy is collected per picture element when it is scanned at 200 lines than when it is scanned at 100 lines. In other words, the amplitude of the picture signal, because of the scanning rate, will vary inversely with the number of elements in the picture.

If the number of lines is kept constant and the frame frequency is varied, the amplitude of the picture signal will vary inversely with the frame frequency. The total amount of light energy collected per second from the picture will be the same regardless of the number of times it is scanned per second. However, the light energy collected per frame will vary inversely with the number of frames per second. Thus, the scanning rate (picture elements traversed per second) depends on two factors, the number of picture elements and the frame frequency. The amount of light energy collected from a picture element will vary inversely with the scanning rate.

From the foregoing it is evident that the amplitude of the picture signal depends on the number of picture elements and the number of elements scanned per second. Putting this statement in the form of an equation, we have:

$$\frac{V_1}{V_2} = \frac{(A_2)^4(R_2)^2N_2}{(A_1)^4(R_1)^2N_1}$$

where V_1 is the average amplitude of the picture signal generated by scanning a picture with A_1 lines, with an aspect ratio of R_1 and a frame frequency of N_1. Similarly, V_2 is the picture signal amplitude for the picture with the constants A_2, R_2, and N_2.

Let us compare the amplitudes of the picture signals produced by scanning a picture mechanically with the best definition and frame frequency that is feasible with mechanical scanning and by scanning the picture with the definition and frame frequency that are considered acceptable for satisfactory entertainment value. It is easily possible to scan mechanically a picture with 90 line definition, an aspect ratio of 1:1, and a frame frequency of 20. The modern standard requires a picture of 441-line definition, with an aspect ratio of 4:3, and with a frame frequency of 30. Let us substitute these values in the above equation. Let V_1 be the amplitude of the picture signal generated by scanning the 90-line picture and V_2 the picture signal amplitude from the 441-line picture. Then:

$$\frac{V_1}{V_2} = \frac{(441)^4(4\div3)^2 30}{(90)^4\ (1\div1)^2\ 20}$$

and

$$\frac{V_1}{V_2} = 1530 \text{ approximately.}$$

Therefore, the acceptable standards system gives a picture with an amplitude $\frac{1}{1530}$ of that obtainable with the mechanical scanning system.

To obtain a better conception of the minuteness of the picture signal amplitudes, it is interesting to express the magnitude of the energy of a picture element in terms of the electron. This will be done on the basis that the direct pickup method of scanning is used. It is the only system that is suitable for televising outdoor as well as studio scenes. If a photographic camera with a fast lens is focused on an outdoor scene of average brightness, the total amount of light falling on the photo-sensitive plate will be of the order of $\frac{1}{10}$ lumen. Let us scan the picture according to the two sets of standards previously mentioned. If we use a vacuum photocell with a sensitivity of 20 microvolts per lumen, calculations show that the average energy per element in the 90 line picture will produce a photocell emission of 9600 electrons and the average energy per element in the 441-line picture will produce a photocell emission of 6 electrons.

It is rather difficult to realize the actual minuteness of these charges. One microampere represents a flow of 6,300,000,000,000 electrons per second. D'Arsonval instruments capable of measuring currents as small as one microampere are very delicate and expensive.

In deriving the electrical energy equivalent of the average picture element for both the 90-line and the 441-line picture, the direct pickup system of scanning was used. This method was unsatisfactory with the old low-definition mechanical scanning systems and the flying spot system was used with its banks of photocells to pick up the light energy reflected by each picture element. This resulted in a greater electron emission per picture element than that given above for the 90-line picture. Also, the intensity of the light source could be made very much greater without discomfort to the subject.

3. AMPLIFICATION OF SMALL SIGNALS. In a preceding paragraph it was stated that the energy of the average picture element in a 441-line picture caused the photo-emission of 6 electrons. Let us assume that the brightest element will produce a photo-emission of 12 electrons. A dark picture element will produce no photo-emission. The maximum voltage generated across the photocell load resistance will occur when the scanning aperture passes from a bright element to a dark element. With a 10,000 ohm load resistance and an electron change from 12 to 0, calculations show that a peak to peak voltage of .00000015, approximately, is produced. The size of the load resistance is determined by the range of frequencies in the picture signal and the shunt capacity to ground.

It is practically impossible to amplify voltages of that magnitude with conventional vacuum tube amplifiers. The noise generated in the first tube of the amplifier by the shot effect may have a greater amplitude than the signal voltage. The shot effect is caused by the fact that the electron stream is a series of particles and not a continuous fluid. Because of this fact, the electron flow to the plate is irregular, that is, the number of electrons arriving at the plate will vary instantaneously. This irregular variation in plate current will add a noise component to the signal. The noise generated by the shot effect covers a frequency range from zero to frequencies much higher than those included in the radio spectrum. The energy content is practically uniform over the entire frequency range. This means that the noise produced by the shot effect will be proportional to the band of frequencies passed by the amplifier. The magnitude of the shot effect is reduced by the presence of the space charge. The space charge smooths out the irregularities in the plate current. Then in order to minimize the shot effect, the emission from the cathode should be sufficient to produce an adequate space charge. In other words, the cathodes must be operated at rated temperatures.

There is noise generated in a photocell by the shot effect. The electron flow to the anode will be slightly irregular for the reason given for vacuum tubes. Since there is no space charge in a photocell, as it is usually operated, the noise generated by the shot effect will be maximum. Thus, there will be a noise component in the weak picture signal before it is amplified. The signal to noise ratio in a photocell will be less for larger electron emission as a greater number of particles will produce a more uniform flow.

Tube noise is also caused by changes in emission over small cathode areas. This is known as the flicker effect. Noise is also generated by the partial neutralization of the space charge through the formation of positive ions either by ionization of the residual gas in the tube or by emission from the cathode.

There is another source of noise that prevents amplification of such small voltages. It is due to the thermal agitation of the electrons in the input resistance of the amplifier. You are familiar with the fact that the free electrons in a conductor are constantly in motion. The magnitude of the motion is dependent on temperature. That is, at higher temperatures, the electrons move faster.

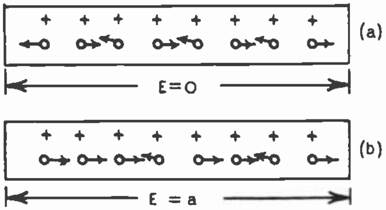


Fig.1 Noise produced by thermal agitation.

Let us consider a conductor as shown in Fig. 1A. The total electric charge of the conductor is zero. There is a balance between the free electrons and positive charges on the atomic nuclei. However, as the free electrons move about the conductor there will be intervals when there are more electrons moving in one direction through the conductor than in the other, Fig. 1B. This means that there will be a voltage developed across the conductor. The magnitude and polarity of this voltage is constantly changing. If the conductor is a coupling resistance between a photocell and the first amplifier stage, the voltage developed across this resistance will form a noise component in the signal. This noise covers the entire frequency range from zero to frequencies greater than the highest in the radio frequency spectrum. The magnitude is fairly constant over the entire frequency range. Like the shot effect, the noise produced by thermal agitation will be proportional to the band width passed by the amplifier. Calculations show that the noise developed across the 10,000 ohm resistance previously mentioned would amount to .000024 volts RMS for a frequency band width of 3.5 megacycles at a temperature of 75 degrees. This voltage is considerably greater than the peak to peak voltage developed across the same resistor when the scanning aperture moved from a bright to a dark picture element.

From the preceding paragraphs, it is evident that the problem of amplification of the small signal voltages produced by high definition scanning is unsolvable by conventional vacuum tube amplifiers. The amplitude of the generated picture varies inversely with the square of the number of picture elements for a given frame frequency and the noise developed in the first stage of the amplifier by the shot effect and thermal agitation increases with the frequency passed by the amplifier.

4. THE ELECTRON MULTIPLIER. One solution to the problem of amplifying voltages that are smaller in magnitude than the noise generated in the first stage of a vacuum tube amplifier by the shot effect and thermal agitation in the coupling resistors is the electron multiplier. The electron multiplier is based on the fact that the secondary emission from some surfaces is much higher than the incident primary electrons. Fig. 2 illustrates the operation of one form of an electron multiplier. The surfaces A, B, C, D, and E have a high secondary emission ratio. Cathode K is the source of the primary electrons such as the cathode of a photoelectric cell. The secondary emission surfaces or targets are maintained at higher and higher positive potentials as they approach the collector P. Let us trace the path of one electron from the cathode

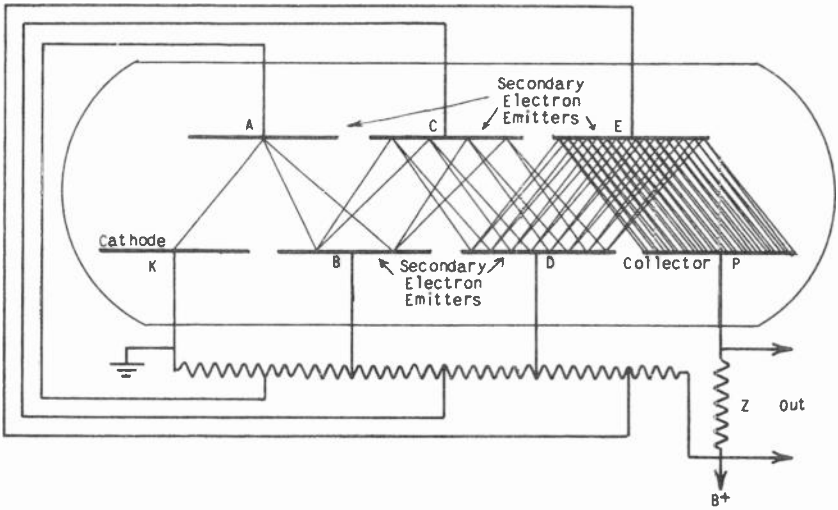


Fig.2 Operation of an electron multiplier.

K through the electron multiplier to the collector P. Also let us assume that the targets have a secondary emission ratio of 2. The electron from the cathode K is accelerated to the target A, as A is positive with respect to K. When the electron strikes A, it will liberate 2 secondary electrons. Since B is positive with respect to A, it will attract the 2 secondary electrons emitted by A. When these 2 electrons strike B, each one of them will liberate 2 secondary electrons from B. This makes a total emission from B of 4 electrons. The same multiplication of 2 will occur at each of the other targets as the electron stream progresses through the tube. The total output current is collected by P. Of course, P does not emit secondary electrons. If a load Z is inserted in the lead to P, a voltage output can be obtained from the electron multiplier.

Let us derive a relation for the gain of the electron multiplier. The electron multiplier shown in Fig. 2 has five stages; that is, the secondary emission multiplication takes place five times. Each stage has a gain of 2. The total gain is 2^5 or 32. If we let R be the secondary emission ratio, n the number of stages, the gain will be equal to R^n. If the input current is I_0 and the output current I, we have:

$$I = I_0 R^n$$

The actual construction of an electron multiplier is more complicated than the simple structure illustrated in Fig. 2. An electron multiplier having the simplified structure of Fig. 2 would not function, as the electrons from the cathode K or any of the targets would go directly to the collector P and there would be no multiplication. In a successful multiplier the targets must have a high secondary emission ratio and there must be a method to focus the electrons on each target and to collect the secondary electrons emitted by each target.

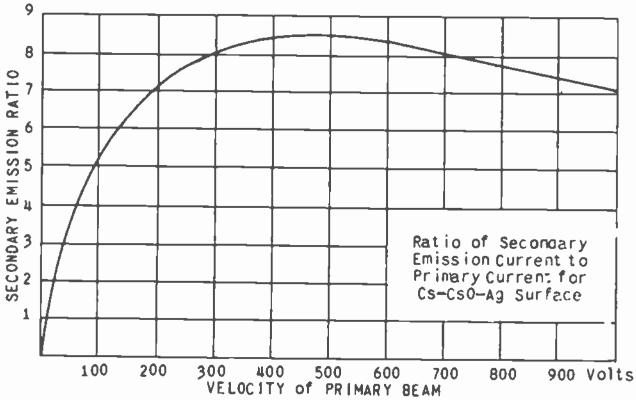


Fig. 3 The variation of the secondary emission ratio with changes in the velocity of the primary electrons.

The most satisfactory targets, like the best photocells, consist of a thin caesium film on oxidized silver. Such surfaces have a maximum secondary emission ratio of eight or ten with primary electron velocities of 400 to 600 volts. Fig. 3 shows the relation of the secondary emission ratio to the velocity of the primary electrons in volts for caesium on a silver-oxide surface.

The efficiency of an electron multiplier is rated in two ways. The first method states the efficiency of the secondary electron emission in terms of amperes per watt of the bombarding emission. The second considers the gain obtainable for a given total multiplier voltage in terms of the number of stages and the gain per stage. Fig. 4 shows the relation between the secondary emission in milliamperes per watt of the incident primary emission, and the velocity of the primary beam in volts for a caesium-silver-oxide target. From this graph it is evident that the secondary emission is most efficient for an electron velocity of 30 volts. For 30 volts, the secondary emission is 58 milliamperes per watt of incident electron energy. Fig. 5 shows the maximum gain obtainable from a caesium-silver-oxide multiplier for a given number of stages as the overall voltage or the voltage per stage is varied. The point of contact between the diagonal line and the curves for the gain of each multiplier is the maximum gain obtainable per volt of the applied voltage. From the graph, the gain per volt of a multiplier having 10 stages (n = 10) is maximum for an overall voltage of 440 or a voltage of 44 per stage. The total gain is 10,000. Examining the other curves, we see that the maximum gain per volt occurs for voltages of 40 to 50 per stage. The stage voltage for the maximum output of the secondary emission in milliamperes per watt and the maximum gain per volt are somewhat different. The maximum gain per volt is the more important from a practical point of view as it is preferable to operate the multiplier with as low an overall voltage as possible.

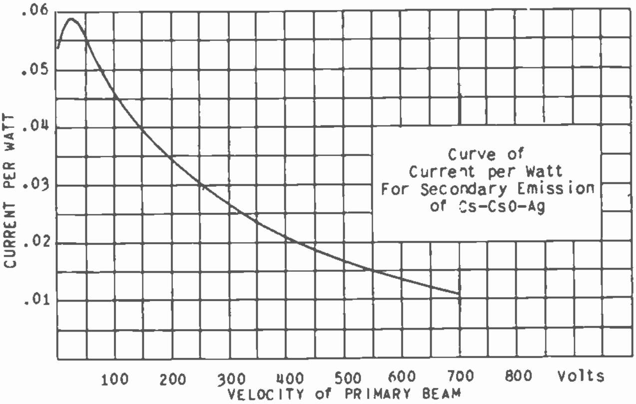


Fig. 4 The efficiency of the secondary emission in milliamperes per watt of the primary electrons for various velocities of the primary electrons.

When R, the secondary emission ratio, is very large, the output signal-to-noise ratio of the multiplier is practically the same as that of the input electron stream. If the source of the input electrons to the multiplier is the cathode of a photocell, calculations show that for the same signal-to-noise ratio, the light falling on the photocell cathode can be $\frac{1}{200}$ of the value of the light needed when the photocell is coupled to an amplifier by means of a resistance. You recall that most of the noise generated in the input stage of an amplifier is due to the thermal agitation in the coupling resistances.

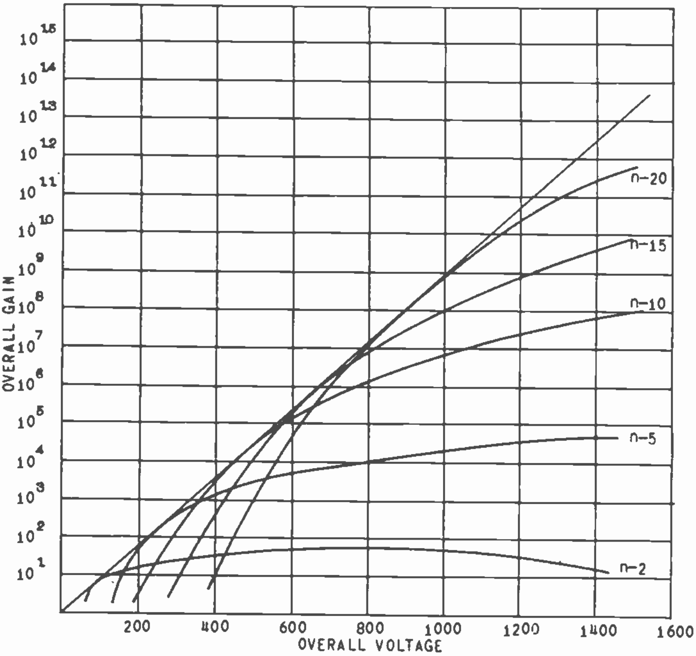


Fig.5 The variation in the overall gain of an electron multiplier as the number of stages and the voltage per stage is changed.

There are two factors that limit the sensitivity of the electron multiplier. The first factor is the emission from the targets. The targets have a very low work function and there is electron emission at room temperatures. This low thermionic emission can be prevented by operating the multiplier at low temperatures.

The other factor that limits the sensitivity is the presence of positive ions. However by proper design of the tube, this source of noise can be minimized.

The frequency response of the electron multiplier is flat from DC to frequencies of hundreds of megacycles.

It was stated in a previous paragraph that some means must be provided in the electron multiplier to collect the secondary electrons emitted by a target and focus them on the next target. Two methods are used. One consists of a combination of electrostatic and magnetic fields to do the collecting and focusing, and the other depends on electrostatic fields to do the collecting and focusing.

Fig. 6 represents the cross section of an electron multiplier that uses a combination of electrostatic and magnetic fields for collecting and focusing the electrons. There are two sets of plates in the tube. The lower set forms the secondary emission targets.

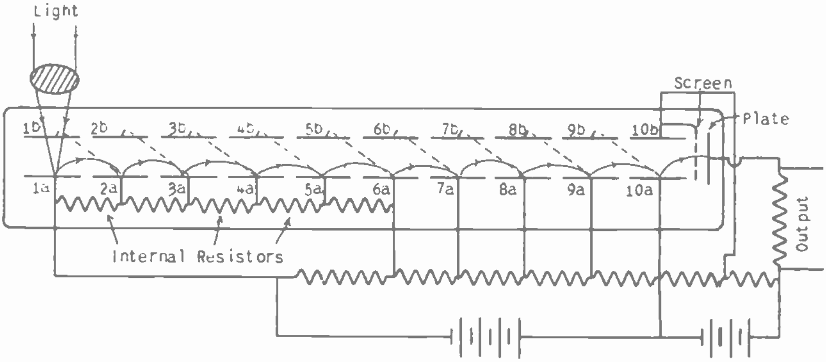


Fig.6 Electron multiplier using a combination of magnetic and electrostatic fields for focusing.

The upper set does not have secondary emission characteristics. The upper plates draw the secondary electrons away from the target. Each upper plate has the same positive potential as the next succeeding lower plate. The dotted line indicates the internal connection between the upper plate and the succeeding lower plate. The two rows of plates are placed as close together as possible so that the electric field will be as strong as possible and draw all the secondary electrons from the target. The spacing between the plates is one-half the distance between centers of the targets. This is the minimum that will prevent the upper plates from collecting some of the secondary electrons. The magnetic field is applied at right angles to both the axis of the tube and the electric field between the upper and lower plates. You recall from the discussion on magnetic focusing in a previous lesson that an electron accelerated by an electric field is deflected at right angles to the direction of the magnetic field and also at right angles to the direction that it is moving. Since the electron is caused to move at right angles to its direction, it will describe a spiral path. In the electron multiplier the intensity of the electric and magnetic fields are adjusted so that the secondary electrons will follow a semi-circular path to the next target as shown in Fig. 6. The secondary electrons knocked out of each target thus follow a semi-circular path to the next target. The magnetic field is usually produced by a permanent magnet.

The electron multiplier shown in Fig. 6 has a shield grid around the collector. This is to prevent changes of the collector potential from reacting on the potentials of the targets and causing feedback and oscillation. To reduce the number of leads to the tube, the voltages for the first few targets are supplied by

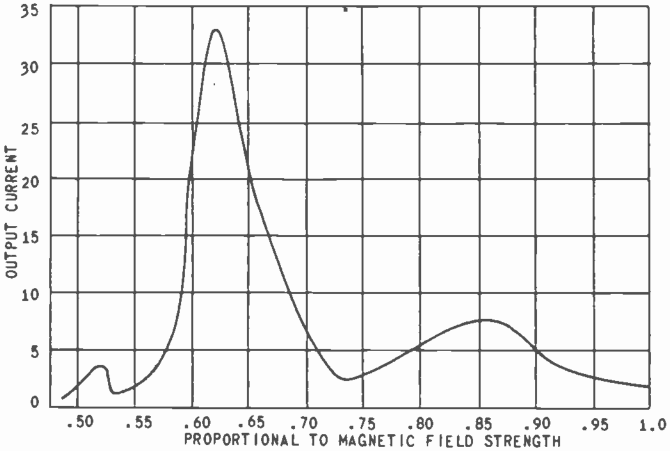


Fig.7 The variation of the output current of an electron multiplier as the magnitude of the magnetic field is changed.

an internal voltage divider. This can be done as the current taken by the first few targets is very small. Fig. 7 shows how the output current varies as the intensity of the magnetic field is varied.

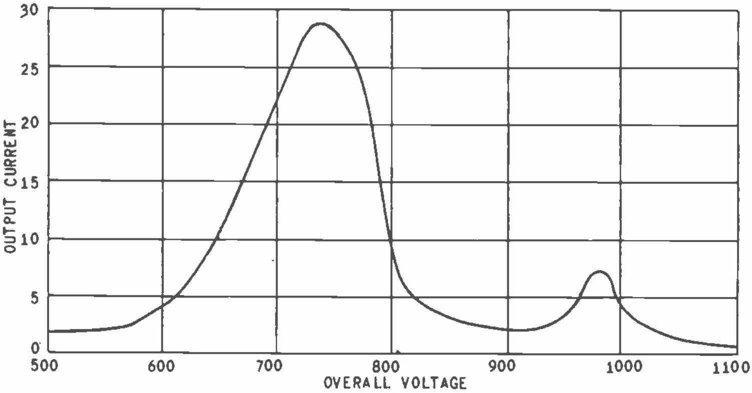


Fig.8 The variation of the output current of an electron multiplier as the overall voltage is changed.

Fig. 8 shows how the output current varies as the overall voltage is changed. The smaller peaks are caused by electrons that have missed some of the early targets.

The construction of a multiplier that uses electrostatic fields only to focus the electrons is more complicated than the magnetically focused multiplier. You recall from your study of a previous lesson that electrons can be focused electrostatically by accelerating them through a series of cylinders at different potentials. This system can be used in an electrostatically focused multiplier but a multi-stage multiplier using cylinders is rather large and bulky. However, by very careful design of the shape of the targets, successful electrostatic multipliers have been built with a structure almost as simple as that of a magnetically focused multiplier.

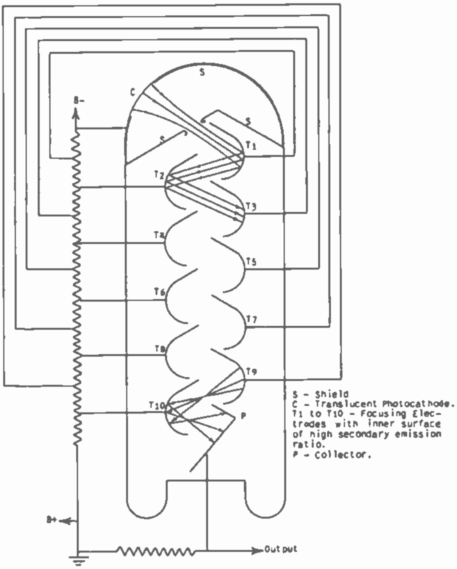


Fig.9 Electrode structure of an electron multiplier using electrostatic focusing. (Zworykin)

Fig. 9 is a diagram of the electrode structure of an electrostatic multiplier photocell developed by Zworykin. The curvature of the plates directs the electrons from one stage to the next. The electrons are focused on the inner surface of each plate which has been treated to have a high secondary emission ratio. Each plate is operated at a higher positive potential than the one preceding. The electrons from the translucent photocathode are collected by the first stage and the resulting secondaries are focused on the second stage etc. The multiplier in Fig. 9 has ten stages and operates with 200 volts per stage. The overall gain is 13,000,000.

Fig. 10 shows the electrode structure of an eleven stage electrostatically focused multiplier, developed by Farnsworth. This multiplier has been designed to operate with a television pickup tube (See Fig. 18). The initial electrons enter the multiplier through the aperture at high speeds. This multiplier, when operated with one hundred volts per stage, has a gain of 20,000.

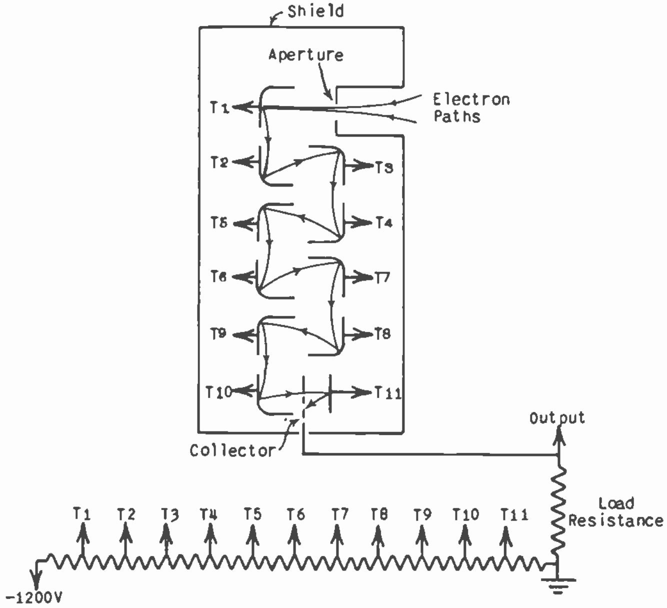


Fig.10 Electrode structure of an electron multiplier using electrostatic focusing. (Farnsworth)

Fig. 11 shows a multiplier phototube of the magnetic type. The magnetic multiplier has ten stages and a gain of several million. The magnetic tube is shown with a type 59 tube so that an idea of its size may be obtained.

There is another form of electron multiplier that has only two secondary emission targets. Multiplication is obtained by sending the electrons back and forth between the two targets. Secondary electrons are emitted at each impact and add to the electron stream. After several round trips, the total emission is collected by an anode. Fig. 12 is a cross section of this type of multiplier. The targets are flat discs at the ends of the multiplier. There is a cylindrical anode in the center between the secondary emission targets. The multiplier is surrounded by a coil which produces a magnetic field for focusing the electrons. An RF voltage is applied to the targets. The center of the coupling transformer is grounded. Therefore, the targets are at ground potential as far as DC is concerned. The intensity of the focusing field can be varied by the rheostat R.

Suppose a few initial electrons are emitted by one of the targets through photo-emission or other means. These electrons are accelerated as they approach the anode and decelerated as they pass the anode and near the other target. The net velocity that is gained through the accelerating voltage of the anode is zero. The velocity that the electrons have when they strike the second target

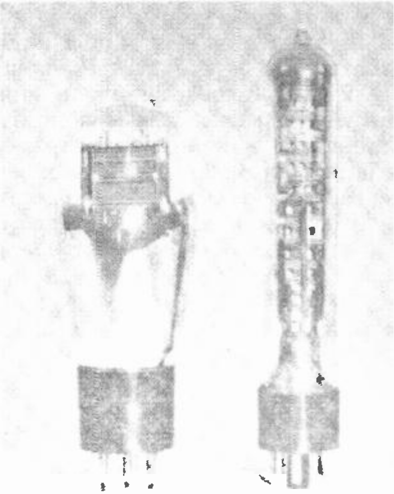


Fig.11 The relative sizes of a 10-stage magnetic electron multiplier and a type 59 tube.

will depend upon the RF voltage. If the RF voltage is positive in the direction the electrons are moving when they approach the second target they may have sufficient velocity to reach the target and knock out some secondary electrons. If the RF voltage is negative when the electrons approach the second target, they will never reach the target but they will reverse direction and start back. Thus it is evident that the phase of the RF at the time the electrons are approaching a target is an important factor in determining whether or not secondary emission will take place. The anode voltage determines the time of transit[1] of the electrons as higher

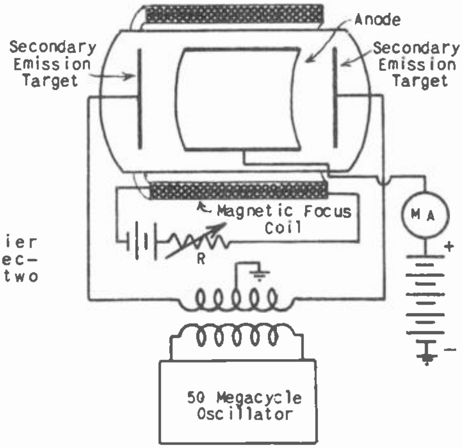


Fig.12 An electron multiplier using an RF field to drive electrons back and forth between two secondary emission targets.

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[1] Transit — Passing over or through.

anode voltages produce more rapid acceleration and deceleration. The focusing magnetic field prevents most of the electrons from reaching the anode until several round trips have been made.

Fig. 13 shows the anode current as the anode voltage is varied from zero upward. The anode current builds up and drops to zero three times as the anode voltage is increased. Each succeeding maximum is greater than the preceding lower voltage maximum. After the third maximum, the current is zero or essentially so for further increases of anode voltage. Varying the frequency of the RF shifts the current peaks along the voltage axis. Shutting off the magnetic field stops the anode current. Increasing the amplitude of the RF causes the peaks to merge into one.

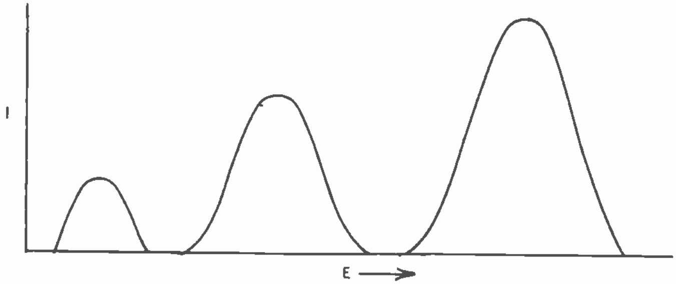
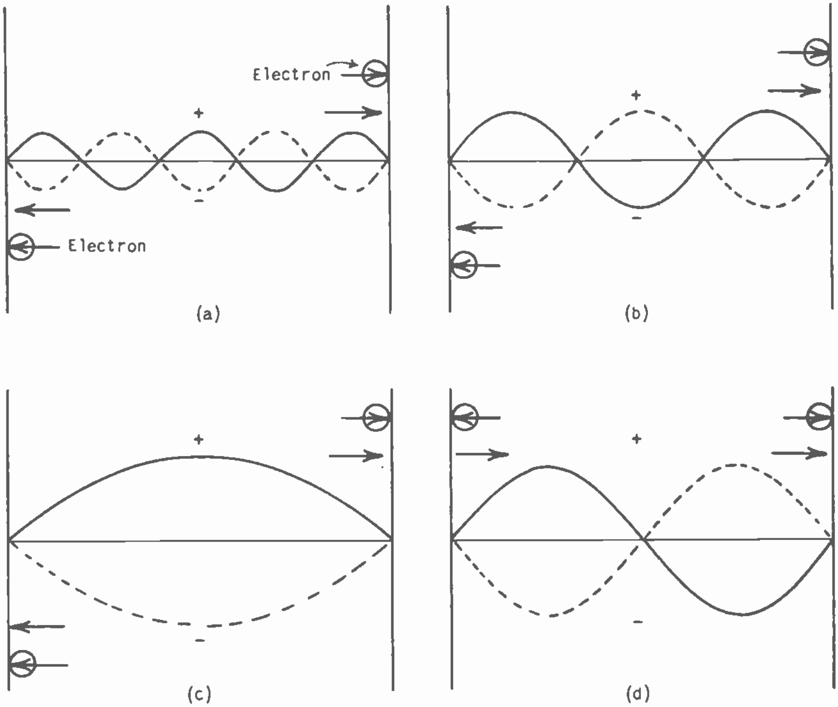


Fig.13 The variation of anode current as the anode voltage is changed.

There will be an anode current only when secondary electrons are emitted from the targets. In order for the electrons to strike a target with sufficient velocity to knock out secondary electrons the RF voltage must be of proper phase to produce maximum acceleration as the electrons approach a target. Maximum acceleration will be obtained if the RF potential on the target reverses just as the electrons strike the target. Then the electrons will have the complete positive cycle to produce acceleration as they approach the target. If the RF potential reverses before the electrons reach the target the negative field will decelerate the electrons so that the impact is insufficient to knock out secondary electrons or may decelerate them to the extent that the electrons will not reach the target. If the RF reverses after the electrons reach the target, again they may lack sufficient energy to knock out secondary electrons as the accelerating field will not have acted on them long enough. Therefore, the RF field must reverse just as the electrons reach the targets at each end of the multiplier for secondary emission to occur. This means that the time of transit between the two targets must be equal to an odd multiple of a half-cycle of the RF. This will be clear if Fig. 14 is examined carefully. There are four conditions illustrated in Fig. 14; (a) transit time equal to five half cycles, (b) three half cycles, (c) one half cycle, and (d) two half cycles. The solid sine curve represents the RF potential conditions for the left to right motion of the electrons and the dotted sine curve for the right to left motion of the electrons. The positive polarity of the sine curve causes electron acceleration to the right, while the negative polarity causes electron acceleration to the left. The direction of the electrons and the accelerating RF field are shown at the time of impact for the four conditions. For the odd number of half cycles the electron direction and the accelerating field direction are the same at both ends of the electron path. For the even number of half cycles, the electron is accelerated before the impact on the right side and decelerated at the left side. Then for the even number of half cycles, the chances for the production of secondary electrons is very much reduced. The three peaks in Fig. 13 show the anode current for the conditions shown in Fig. 14 a, b, and c. The maximum anode current occurs when the transit time is one half cycle. Since the RF field conditions won't be favorable for the production of secondary electrons for transit times less than one half cycle, decreasing the transit time below one half cycle will not cause an anode current flow. Through careful adjustment it is sometimes possible to obtain an anode current for transit times equal to seven and nine half cycles.



(a) (b) (c) (d)

Fig.14 Conditions for secondary emission from the targets in terms of the electron transit time and the period of the applied RF field.

The current that the anode collects for the equilibrium condition depends on the number of round trips that the electrons make as there will be more secondary electrons produced. Equilibrium conditions exist when the anode current is self sustaining; that is, no energy from an external source, such as photo-electrons, is required to maintain anode current. Factors which control the equilibrium anode current are the intensity of the magnetic focusing field, the part of the cathode where the initial multiplication starts (electrons coming from the edges are collected most easily), and the field existing between the anode and targets. Increasing the intensity of the focusing field will decrease the probability of an electron reaching the anode. This means that the electron can make several round trips before being caught by the anode. Thus, there will be many more electrons existing in the multiplier and for the anode to collect. The field existing between the anode and targets is determined by the potential on the anode and the space charge. The space charge will limit the maximum anode current.

If we examine the curve in Fig. 13 we see that the multiplier has both negative and positive resistance characteristics. That is, the anode current over part of the range decreases as the anode voltage is increased and over other parts increases as the anode voltage is increased. Thus the multiplier can be used as an oscillator over the positive or negative sections of its characteristics. If adjustments are made so that the transit time is equal to one half cycle of the resonant frequency of the circuit coupling the two targets, oscillations of sufficient amplitude to destroy the tube are produced if no precautions are taken to limit the number of round trips or multiplications.

If the multiplier is to be used to amplify small currents, the anode current must be completely under the control of the external energy source. Therefore, the number of multiplications must be limited so that the equilibrium condition cannot exist. Also the number of round trips that the original electrons make before they and their resultant secondary electrons are collected by the anode must be a definite number. One way that this can be accomplished is to interrupt periodically the RF field or the focusing field so that all the electrons can be cleared out of the multiplier. If the interruptions are periodic the number of multiplications that occur are fixed. Another method is to use an intense magnetic field so that the first impact of the initial electrons occur near the center of the targets. Each additional impact will occur nearer the edge until a finite number of impacts have been made and the electrons collected by the anode. High amplifications can be obtained in this way before the multiplier becomes self-sustaining or "breaks down". If an intense magnetic field is not convenient or available, the RF voltage that causes the multiplier to break down for a given magnetic field is determined and the multiplier is operated just under the breakdown condition. The maximum multiplication is obtained when the intense magnetic field is used.

From the foregoing discussion it is evident that the electron multiplier is one answer to the problem of the amplification of small currents. The electron multiplier, when operating properly, adds very little noise to the signal. In general the signal-to-noise ratio of the output is very little higher than the signal to noise ratio of the input. One disadvantage of the electron multiplier is that it is a current-controlled mechanism and not voltage controlled; that is, the input must be a current. If a grid were introduced into the multiplier to use voltage control the noise-free amplification of the multiplier would be wasted because the noise generated across the coupling impedance through thermal agitation would over-ride the signal. However, the electron multiplier when used in combination with a photocell is a light-sensitive device of very high sensitivity. In fact, an electron multiplier photocell will produce a satisfactory picture signal when used in a direct pickup system using 441-line definition.

5. THE STORAGE PRINCIPLE. The electron multiplier is one solution to the problem of obtaining an adequate noise-free picture signal for a high-definition television picture. In this section of the lesson will be described the principle underlying another solution to the problem.

The actual electric energy obtained per picture element in high-definition scanning is a small fraction of the available light energy of the picture element. For example when a 441-line picture with 259,308 picture elements per frame is scanned at the rate of thirty frames per second, each picture element is contacted by the scanning aperture 30 times per second but the time that the scanning aperture is on each picture element is:

$$\frac{30}{259,308 \times 30} \text{ second,}$$

or

$$\frac{1}{259,308} \text{ second.}$$

The total number of picture elements scanned per second is 30 × 259,308 and each element is scanned 30 times. For a 90-line square picture transmitted 20 times per second the scanning aperture covers each picture in:

$$\frac{20}{8100 \times 20}$$

or

$$\frac{1}{8100} \text{ second.}$$

Then for the 441-line picture the scanning aperture collects $\frac{1}{259,308}$ of the light energy radiated in one second and for the 90-line picture the scanning aperture collects $\frac{1}{8100}$ of the light energy radiated in one second. This also is a restatement of the fact that the amplitude of the picture signal is reduced as the number of picture elements is increased.

Now if it were possible to store up all the light energy radiated by a picture element between scannings and convert it into electrical energy at the time the scanning aperture passes over the picture element there would be a tremendous increase in the amplitude of the picture signal generated. For the 90-line picture the gain would be 8100 and for the 441-line picture the gain would be 259,308. If only a fraction, say 10%, of the total gain for the 441-line picture could be realized, it would be worth while.

Fig. 15 shows the operation of a television pickup system based on the storage principle. There is a photocell and condenser for each picture element. For a 441-line picture there would be 259,308 photocells and condensers. In the figure let the four photocell-condenser combinations represent the first four picture elements in the first line of the picture. The switch S is the scanning mechanism. There will be 259,308 contacts on the switch, one for each photocell-condenser combination. The switch arm rotates in the clockwise direction. Each condenser is charged through the photocell and resistance R by the battery B. The charge that each condenser accumulated will be proportional to the intensity of the light incident on that particular photocell. When the scanning arm discharges each condenser there will be a positive pulse developed across the resistor R whose amplitude will be proportional to the condenser charge which in turn was proportional to the light intensity on the corresponding picture element. The voltage developed across R will be proportional to the energy radiated by a picture element over the interval between scannings rather than the energy radiated during the time that the scanning aperture is on that particular element. Thus there is a vast increase in the magnitude of the picture signal generated. As stated in a previous paragraph this gain is numerically equal to the number of picture elements. For a 441-line picture this gain is equal to 259,308. Even if only 10% of this gain is realized, there is a gain of 25,930 in the amplitude of the picture signal. The Iconoscope developed by Zworykin of RCA and the Emitron developed by the British E.M.I. Research laboratories are television pickup tubes based on the storage principle. With the present state of perfection, the gain realized by the storage principle in these tubes is from five to ten percent.

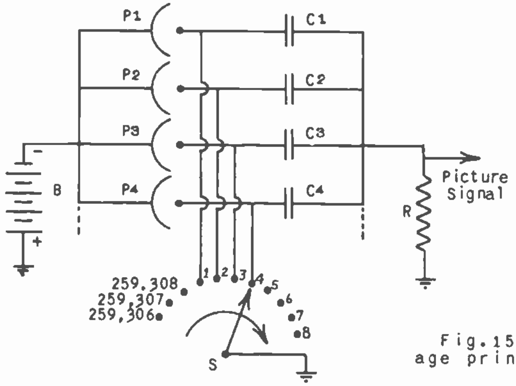


Fig. 15 Illustrating the storage principle.

We have now described two solutions to the problem of obtaining an adequate picture signal when employing modern high-definition scanning. In the next few paragraphs we shall describe solutions to the problem of obtaining a satisfactory method for scanning high definition pictures.

6. THE IMAGE DISSECTOR TUBE. Farnsworth was one of the first to develop a method for scanning electronically. He calls his electronic scanner "The Image Dissector". Fig. 16 shows a functional diagram of the tube. The front inside surface of the glass tube is coated with a translucent cathode K, which is light-sensitive. Near the other end of the tube is a tight fitting silver disc T. In the center of this disc is a small hole O. The size of the hole is the same as a picture element. On the inside of the tube between the cathode K and the disc T is a thin metallic coating. It is in electrical contact with both the cathode K and the disc T. Its resistance when measured between K and T is on the order of two or three megohms. Behind the hole O is a small plate P. Surrounding the tube and placed between the cathode and disc T are two sets of magnetic deflecting coils, one for the horizontal deflection and the other for vertical deflection. The tube and the deflecting coils are surrounded by a magnetic focusing coil F. There is applied a potential of from five to six hundred volts between the cathode K and the plate P, however; most of it appears across the inner metallic coating.

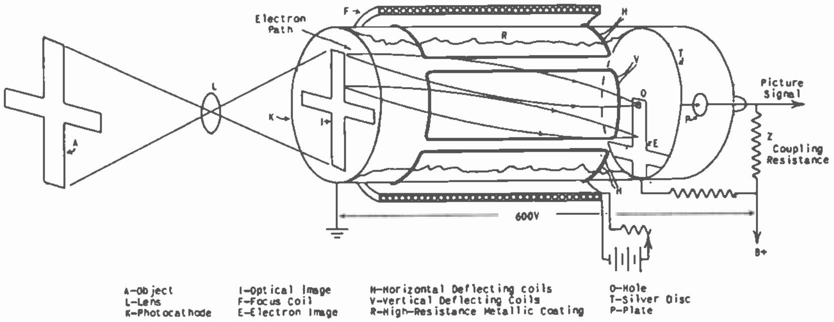
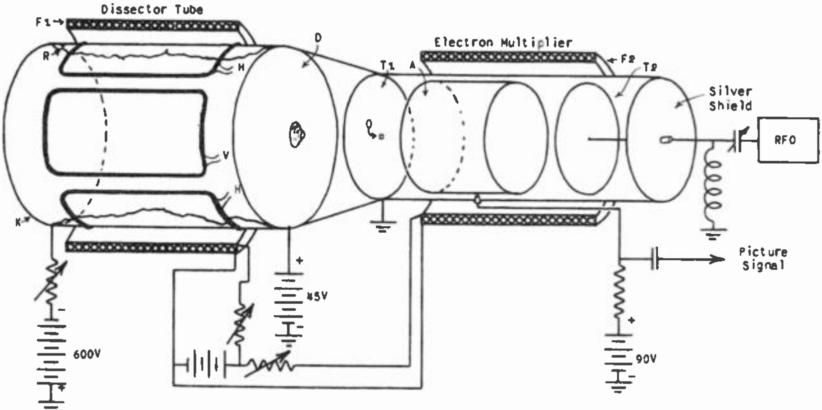


Fig.16 The principle of the Farnsworth dissector tube.

An optical image I of the cross A is focused on the translucent light-sensitive cathode K. This causes electrons to be emitted from the inside surface. The number emitted by any area of the cathode will be proportional to the intensity of the light incident on that area. Since the disc T is at a high positive potential with respect to the cathode K, the electrons emitted from K are accelerated to the disc T. Since electrons repel each other, the electrons coming from a given point of the cathode have a tendency to diverge as they progress through the tube. These diverging electrons are brought to a focus on the disc T by means of the magnetic field produced by the focusing coil F. You are familiar with the principle of magnetic focusing of electrons from your study of a previous lesson. Thus there is produced on the disc T an electron image of the optical image on the cathode K. If we were to replace the disc T with a fluorescent screen, the impact of the electrons in the electron image would produce a fluorescence whose intensity is proportional to the electron density of the corresponding area of the electron image from the original object.

Up to the present we have merely converted the optical image into an electron image. We still have to scan the image so that an electrical communications channel can transmit it. This is the purpose of the deflecting coils mounted around the tube inside the focusing coil. We can think of each electron source or picture element in the cathode K as an electron gun and that we want to scan the disc T with the electron beam originating at that point. As the disc T is scanned by the beam from a given point in the cathode, it will pass over the hole O in the disc T once per frame. The electrons in the beam will pass through the hole and be collected by the plate P. These electrons will produce a voltage drop across Z proportional to the number of electrons in that beam. Similarly the electrons originating from every other picture element source will pass through the hole O and develop a voltage across Z of proportional magnitude. As all the electron beams coming from all the picture elements are deflected parallel to each other, the electron beams will pass over the hole O in the same manner that the holes in a scanning disc pass over every picture element. Fig. 16 the electron beams originating from two picture elements are shown. In considering the scanning disc the picture stays still and the scanning aperture moves. Here, the scanning aperture stays still and the picture moves.



K-Photosensitive Cathode
R-High-Resistance Metallic Coating
F1-Magnetic Focus Coil for Dissector
H-Horizontal Deflecting Coil
O-Hole to admit Electrons to Multiplier
A-Anode of Electron Multiplier

V-Vertical Deflecting Coil
D-Silver Disc
T1 & T2-Secondary Emission Target of Electron Multiplier.
RFO-50 Megacycle RF Oscillator
F2-Magnetic Focus Coil for Multiplier

Fig.17 · The Farnsworth dissector tube with a built-in electron multiplier.

The electrons from each picture element in the picture are collected by the plate P and there will be developed across A a corresponding picture signal. The amplitude of this picture signal will be the same as that produced in a direct pickup system using a scanning disc and a photocell. For high-definition scanning, the picture signal is just as unusable. However, by using an electron multiplier the picture signal can be amplified several thousand times.

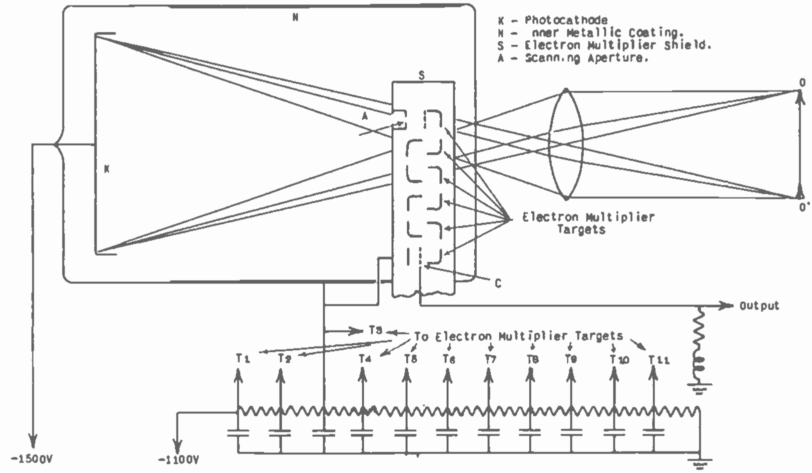


Fig.18 The Farnsworth dissector tube with an electrostatic multiplier. This tube is used for televising moving picture film.

Fig. 17 shows a dissector tube equipped with an electron multiplier. The electron multiplier is of the single stage type (See Fig. 12) where amplification is produced by driving the electrons back and forth between two secondary emission targets and collecting all the resulting secondaries. The electron multiplier is an integral part of the tube. The entire multiplier is enclosed in a grounded metallic cup. The secondary emission target T_1 is also at ground potential. The secondary emission target T_2 is insulated from the enclosing shield. The collector anode is also insulated from the shield. The focusing field for the electron multiplier is supplied with DC from the same source as the focusing field for the dissector tube but the two are adjustable independently of each other. The target D in the dissector tube has an aperture somewhat larger than the size of a picture element. The hole O in the first secondary emission target T_1 is the same size as a picture element. It is through O that the electrons from the photo-cathode K enter the electron multiplier. The RF field is supplied between the target T_2 and ground. The disc D is operated positive with respect to T_1 so that any electrons knocked out of the left side of T_1 will not be drawn into the multiplier. The secondary emission in the multiplier must be controlled by the electrons coming from each picture element on the photo-cathode K.

Fig. 18 shows the construction of a modern dissector tube used for televising motion picture film. An electrostatic type of multiplier (See Fig. 10) is used. The photocathode is not translucent and the image is focused on the inside surface of the photocathode.

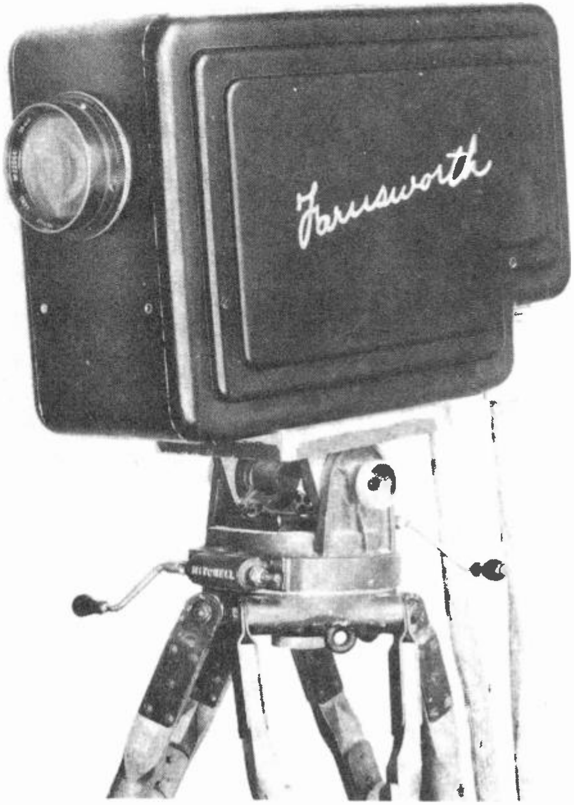


Fig.19 The Farnsworth television camera. (Courtesy Farnsworth Television, Inc.)

This means that the light must travel through the entire length of the tube. The electron multiplier is mounted off the center of the tube so that it will not be in the way of the light rays going to the photocathode. The inside of the tube is coated with a high-resistance coating to serve as an accelerating anode, as in the previous dissector tubes described.

The shield of the multiplier is operated positive with respect to the first target of the multiplier to prevent secondary emission from the shield from being drawn into the multiplier. The input to the multiplier must come only from the electron image produced by the emission from the photocathode. A focusing coil and horizontal and vertical deflecting coils (not shown in Fig.18) are mounted around the outside of the cylinder of the tube. The acanning aperture (input aperture to the multiplier) is a square, .005 inches on a side. With the correct size electron image, 441-line definition is obtained.

Fig. 19 shows a Farnsworth television camera. The cables connect the camera to the control room equipment. They carry the picture signal to control room amplifiers and also carry the necessary operating and deflection voltages to the camera.

7. THE ICONOSCOPE. The Iconoscope, developed by Zworykin, is a television pickup tube employing the storage principle to provide an adequate picture signal. The total energy radiated by a picture element is stored in a condenser during the interval between passages of the scanning aperture over that element. Thus the amount of energy converted into an electrical impulse at the instant of scanning is vastly increased.

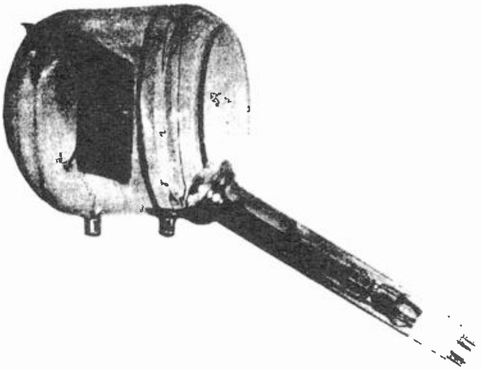


Fig.20 The Iconoscope.

Fig. 20 is a picture of the Iconoscope. Fig. 21 is a schematic diagram. The picture to be televised is focused on the mosaic M. This mosaic consists of a very large number of photo-emissive particles on a thin mica sheet that is approximately four by five inches. Each particle is insulated from every other particle. The other side of the mica sheet is coated with a metallic coating. This metallic coating is called the signal plate (S in Fig. 21). Each photo-emissive particle and the signal plate form a small condenser. In other words, we have the cathode of a photocell in series with a condenser as in Fig. 15. In Fig. 15 each photocell represented one picture element. However, in Fig. 21 it takes several of the small photocells to cover the area occupied by one picture element. This is necessary in order to minimize the effects of variations in the sensitivity and size of the photo-emissive particles. The anode for these minute photocells is a metallic coating A_2 on the inside of the neck of the tube. Scanning or the progressive discharge of the condensers is accomplished by bombarding the mosaic with an electron beam. The electron beam is projected from an electron gun located in the neck of the tube. The coating A_2 forms part of the second anode of the electron gun. The beam is deflected magnetically. Scanning is electronic as in the dissector tube.

The operation of the Iconoscope is considerably more difficult to understand than that of the dissector tube. We shall start with a simple but incomplete description of the production of a picture signal by the Iconoscope. The image of the object being televised is focused on the mosaic by a lens. The photo-emissive particles making up the mosaic emit electrons in proportion to the intensity of the light on them. The photo-electrons are collected by the anode A_2. The condenser formed by any one of the small photo-emissive particles and the signal plate is charged proportionately to the intensity of the light incident on that particle. The particle

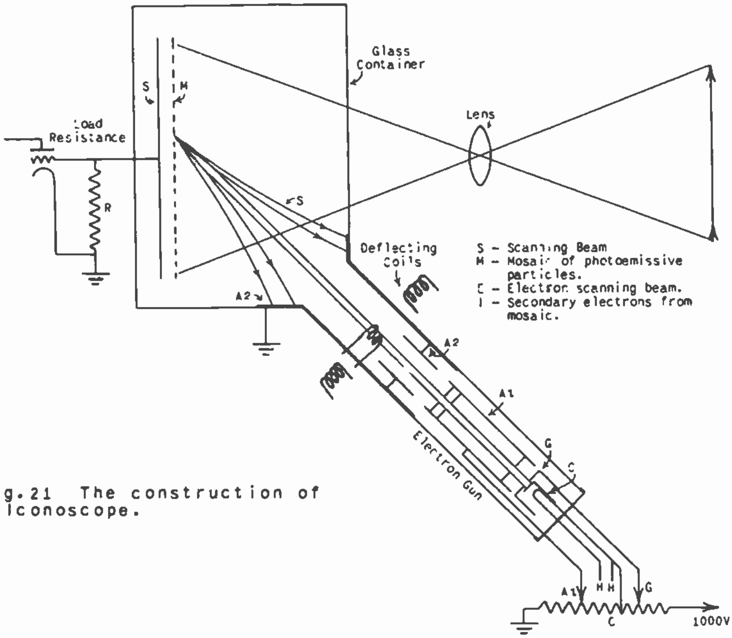


Fig. 21 The construction of the Iconoscope.

assumes a positive charge through the loss of photo-electrons to the anode A_2. Since the signal plate and particle constitute the two plates of a condenser, the electrons collected by the anode will flow to the signal plate via ground and the load resistance R. When the electron beam passes over each particle, it restores to the particle the electrons it lost through photo-emission, or it discharges the condenser formed by the signal plate and the particle. The discharge current of the condenser develops a voltage across R proportional to the charge on the condenser. For a 441-line picture, the condenser discharges for $\frac{1}{259,308}$ of the time during each frame, but charges during the complete frame. Since all the condensers are charged slowly and together through R, the voltage developed across R by the charging current is DC. But since the condensers are discharged several at a time (it takes several particles to cover one picture element) and at a high rate of speed, the discharge current produces a varying voltage across R that corresponds to the light and dark shades in the picture. This varying voltage is the picture signal. The discharge electron current flowing through the resistance R causes a voltage to be developed across R which makes the signal plate end of R negative with respect to ground. It will be least negative for the dark and most negative for the light parts of the picture. Therefore, the polarity of the picture signal is negative.

As stated before this explanation is not complete but will help you in understanding a more complete explanation which will be given later.

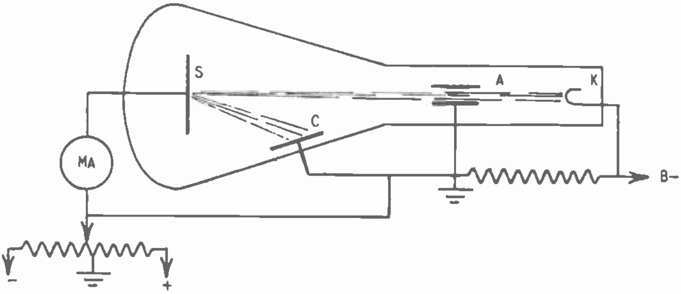


Fig. 22 A tube for finding the factors which control the secondary emission of a target.

The operation of the Iconoscope is complicated by the fact that the particles making up the mosaic have a high secondary emission ratio. As these particles are completely insulated, the potential that they assume with respect to the second anode, before and after bombardment by the electron beam, is governed by the energy of the secondary electrons. In the usual Iconoscope tube the secondary emission ratio of the mosaic is from seven to ten. Let us consider the potential that an insulated body, under electron bombardment, assumes with respect to the electrode that collects the secondary electrons. In Fig. 22 is shown a simple electron ray tube. The cathode K, and the anode A, form a simple electron gun and project a beam of electrons on the secondary emission target S. Let us assume that S has a secondary emission ratio of 10. C is an electrode to collect the secondary emission from S. C and A are operated at ground potential. The cathode is operated at three or four hundred volts negative with respect to the anode. Electrons having that velocity produce the highest secondary emission ratio. The potential of the secondary emission target S can be varied above and below ground.

If the current to the secondary emission target is plotted against the voltage applied to the target, curve A of Fig. 23 results. Curve B is the corresponding collector current. For moderate negative potentials on S, all the secondary electrons are collected by C (secondary emission is saturated) and there is an electron current from ground through the meter to the target that is equal to the difference between the beam current to the target and the secondary emission from the target. As the target potential becomes less negative with respect to the collector, the secondary emission current to the target is reduced due to the formation of a space charge. The secondary emission ratio remains the same but the secondaries that do not reach the collector fall back into the target. Therefore the current to the target is reduced. As the potential of the target becomes zero and then positive with respect to the collector there will be one value of potential difference where the beam current to the target and the secondary emission collected by C are equal and the current to the target through the meter is zero. As the potential is made more positive, the secondary emission is completely suppressed and the beam current returns to the cathode through the target as in a conventional diode. When there is no current going to or from the target via the meter to ground the same state of affairs exists, as far as current is concerned, just as though the target were completely insulated. Then for the primary emission to, and the secondary emission from, an insulated target to be equal, the target must assume the potential with respect to the collector as given in the graph for equal primary and secondary currents. Let the external circuit to S in Fig. 22 be disconnected when the potential of S is at ground. For this condition a large proportion of all the secondary electrons knocked out of the target are collected by C. The remainder fall back into

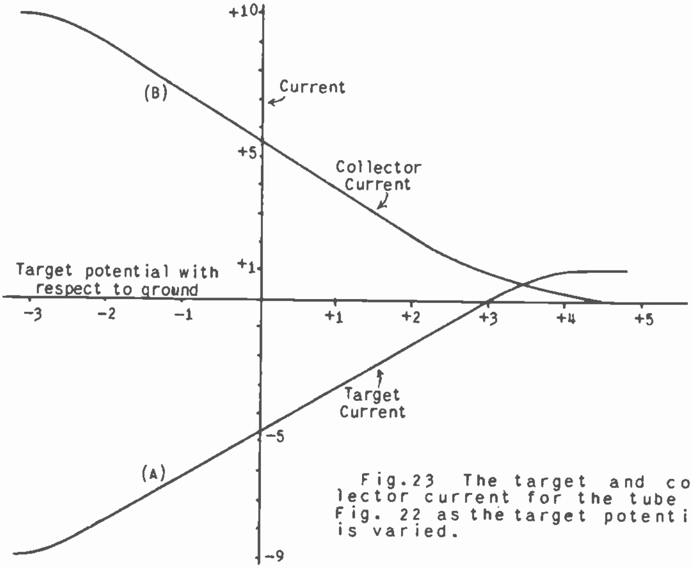


Fig.23 The target and collector current for the tube in Fig. 22 as the target potential is varied.

S. The potential of the targets will become positive with respect to C through the loss of electrons. This will cause a reduction in the magnitude of the current collected by C until the primary current and the collector current are the same. The potential assumed by the target for this condition will be +3, the value on the graph for equal primary and secondary emission to and from the target.

The collector current, curve 3 of Fig. 23, is zero when the target is 4.5 volts positive with respect to the collector or the collector is 4.5 volts negative with respect to the target. This means that the maximum energy of the secondary electrons is equivalent to 4.5 volts.

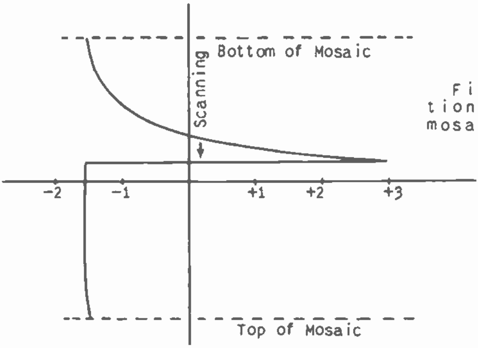


Fig.24 The potential distribution along a vertical line of the mosaic when scanned in the dark.

The student is now better able to follow a more detailed description of the operation of the Iconoscope. If the mosaic is scanned in the dark, the average potential of the entire mosaic with respect to the second anode or collector is between 0 and 1 volt negative. The potential of the element directly under the scanning beam is 3 volts positive with respect to the collector. This is the potential condition for which secondary emission current to the collector is equal to the primary beam current. The potential of the area, before the scanning beam strikes it, is 1.5 volts negative with respect to the collector. The potential of the mosaic changes from 3 volts positive to 1.5 volts negative between passages of the electron beam. When under the beam, the potential changes from 1.5 volts negative to 3 volts positive. Fig. 24 shows the potential distribution of the mosaic when scanned in the dark. This curve shows the potential along a vertical line from top to bottom through the center of the mosaic. The potential rises from -1.5 to +3 when under the electron beam, as +3 is the potential for which the collector current and the primary current are the same. The average value of the collector current for the entire mosaic must be the same as the primary current, while the emission from individual elements can vary slightly around the average. In changing from -1.5 to +3 a large number of electrons are knocked out of each element as they are scanned. Of these, about 25% reach the collector. Of the remainder, part of them fall back into the element from which they were ejected, and the others return to the mosaic. The maximum velocity of this latter group is equivalent to 1.5 volts. This explains why the final potential of the mosaic is -1.5 volts. When any element of the mosaic collects sufficient electrons so that its potential is -1.5 volts, it will not collect any more secondary electrons. About two thirds of the surface of the mosaic is -1.5 volts with respect to the collector and the rest is between -1.5 and +3 volts. The average over the entire surface is between 0 and -1 volt.

When a picture is focused on the mosaic, photo-electrons are emitted by the elements of the mosaic in proportion to the light and dark shades of the image on the mosaic. This means that for an illuminated element the potential before scanning will be slightly more positive than -1.5 volts. Let us assume a value of -1.4 volts as the final potential of a brightly illuminated element before scanning. (This value is exaggerated many times; the actual change in potential due to the loss of photo-electrons is very much smaller than this). When a dark picture element is scanned, its potential changes from -1.5 to +3, and when an illuminated picture element is scanned, its potential changes from -1.4 to +3. With these factors in mind, we wish to derive the polarity of the picture signal developed across the load resistance. (Fig. 21) When the potential of an element on the mosaic changes from negative to positive, there is a surge of electrons from ground through the load resistance R, to the signal plate. This electron surge develops a positive pulse across the load resistance. The positive pulse is greatest when the scanning beam goes over a dark element as the potential change is greater. For example, the potential changes from -1.5 to +3, or a change of 4.5 volts for a dark element, and from -1.4 to +3, or a change of 4.4 volts for a brightly illuminated element. Therefore, white corresponds to a negative voltage in the output of the Iconoscope. (The positive grid swing is less for light than it is for dark). This checks the polarity obtained for the simple explanation of the Iconoscope. Of course there is a flow of electrons from the signal plate to ground when the potential of the elements change from +3 to -1.5, but this is a constant current and the small DC voltage produced will not interfere with the picture signal. Also there is a DC electron flow from ground to the signal plate caused by the emission of photoelectrons from the mosaic. This is the charging current for the condensers formed by the signal plate and the mosaic. The collector current will also vary according to the picture signal. However, the average value of the collector current is the same as the primary beam current. There is also a spurious signal generated due to variations of the potential of the mosaic around the -1.5 volt level. This variation is due to differences in the numbers of secondaries collected by various parts of the mosaic. These spurious signals will be described in more detail in a later part of the lesson.

In the next few paragraphs we shall attempt to explain some of the reasons why the theoretical efficiency of the Iconoscope is not realized. The effective photoemission from the mosaic is only about twenty to thirty percent of its saturated value, because there is no strong electric field to draw the photoelectrons away from the mosaic. The collector has a weak positive field toward the elements of the mosaic when they are 1.5 volts negative with respect to the collector. However, during the interval that elements of the mosaic are positive with respect to the collector (Fig. 24) the field of the collector is a retarding field toward the photoelectrons. Thus, under the most favorable conditions the field of the collector cannot possibly collect a large percentage of photoelectrons. However during a very short interval of each frame there is an intense field available to collect the photoelectrons. This field arises from the fact that the elements just scanned have a potential of +3 and those just in front of the scanning beam have a potential of -1.5. Thus there is a field having a potential of 4.5 volts acting on the photoelectrons of the elements just in front of the scanning beam. This potential is sufficient to saturate the photoemission. This marked increase in sensitivity at the time of scanning can be shown in a rather interesting way.

The image of a continuously run moving picture is projected on the mosaic. Such a projection can be obtained from a standard projector by removing the shutter and intermittent sprocket. The film is moved in the opposite direction to the vertical scanning. If straight scanning (not interlaced) is used, and the film is run through so that the frame frequency is the same as the vertical scanning frequency, two complete frames will be reproduced on the receiving tube. Since the film is constantly in motion before the Iconoscope, the energy stored by all the elements of the mosaic will be identical or equal to the average value of the light from the film. Therefore, the reproduced picture must be from the instantaneous photoemission from the elements during scanning. There will be two complete frames of the picture on the receiving tube because the scanning and the film move in opposite directions and when the scanning beam has covered the upper half of the mosaic, the film has moved up half a frame and thus a complete frame of the picture has passed under the scanning beam.

This increase in sensitivity of an element at the time of scanning is known as line sensitivity. The positive field of the scanned elements is effective in increasing the photoemission for a distance of about two-fifths of an inch into the unscanned area of the mosaic. This increases the stored charge of the elements. For low light intensities, line sensitivity does not add much to the picture signal amplitude. However, for high light intensities, as much as 50% of the signal may come from line sensitivity.

The output versus intensity of light response is not linear for the Iconoscope as might be expected from the fact that the photoemission is not saturated. The space charge formed on the face of the mosaic will prevent the photocurrent from increasing linearly with intensity of the incident illumination. Also the effective field that draws away the photoelectrons will be reduced as the potential of an element rises through the loss of photoelectrons. Brightly illuminated groups of elements, being more positive than their neighbors, will collect more of the secondaries knocked out of the scanned elements and thus will have part of their charge neutralized. Fig. 25 shows the output voltage developed across a 10,000-ohm load resistance for different levels of background illumination as the intensity of the input is varied. Each curve represents the voltage generated in scanning across the boundary of an illuminated area and the background as the intensity of the illuminated area is changed. Curve A is for a completely dark background; curves B, C, and D are for background illuminations of 5, 10, and 30 millilumens per square centimeter respectively. The

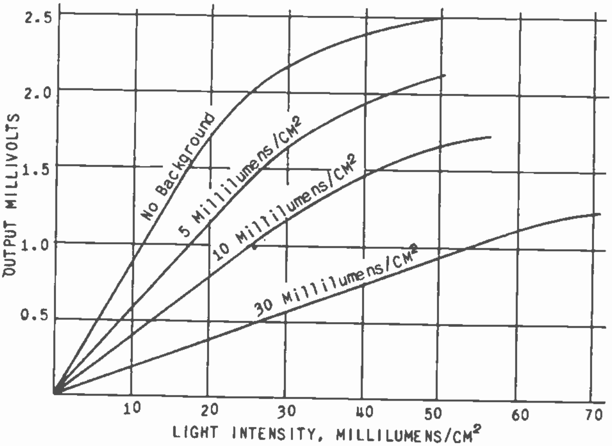


Fig.25 The output voltage of the Iconoscope for different levels of background illumination.

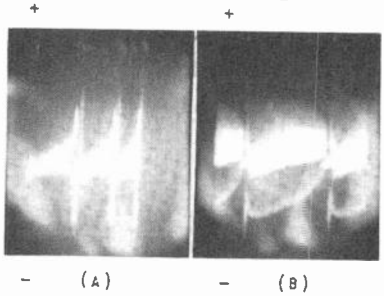
output voltage is plotted along the vertical axis and the intensity of the light producing the illuminated area along the horizontal axis. From these curves it is evident that the output for dark background is greater than for an illuminated background, although the change in illumination from the background to the illuminated area is the same. Also the signal amplitude goes down as the background level increases. All these curves show the non-linearity of the output to the input light intensity. This non-linearity is not disadvantageous as it permits the transmission of a wider range of contrast over an electrical communication circuit. In other words the non-linearity of the Iconoscope serves the same purpose as a compression amplifier in a sound chain.

Only 25% of the charge released from the elements of the mosaic is converted into a picture signal. It was mentioned in a previous paragraph the collector received one-fourth of the electrons knocked out of an element by the scanning beam, and that the remaining three-fourths immediately fall back into the mosaic. The electron current through the load resistance to the signal plate will be the same as the collector current. Thus only 25% of the charge of an element is converted into a picture signal.

Another factor that affects the amplitude of the picture signal is the maximum permissible capacity formed by the elements of the mosaic and the signal plate. The elements must be discharged completely during the instant of scanning; otherwise blurring will result when televising rapidly moving objects. The secondary emission to the collector is equivalent to an ohmic resistance of one megohm. The maximum capacity of the mosaic is approximately 100 mmfds. per square centimeter. The time constant of the discharge circuit is $C(R+R_s)$ where R is the load resistance, R_s is the ohmic equivalent of the secondary emission to the collector, and C is the capacity per picture element. This time constant must be less than the time of scanning for one picture element.

From the preceding paragraphs, it is evident that the efficiency of the present day Iconoscope is very low. The charge stored by the condensers formed by the elements of the mosaic and the signal plate is only twenty to thirty percent of the available energy because of the unsaturated photoelectric currents. Of this charge, only twenty-five percent is converted into a picture signal. Thus the overall efficiency is around five percent. This is a gain of many thousands over the non-storage method of scanning, and a usable picture signal can be obtained for present day television standards.

It was mentioned in a previous paragraph that there is a spurious signal present in the output of the Iconoscope. The spurious signals also exist when the mosaic is scanned in the dark. This signal is caused by small variations in the potential over the surface of the mosaic and in the instantaneous collector current. When there is a picture focused on the mosaic, the spurious signal varies with the light and dark of the picture. The spurious signals result in uneven shading of the picture.



(A) (B)

Fig.26 The spurious signals generated when the mosaic is scanned in the dark. A is the waveform for a line. B is the complete waveform for a frame.

Fig. 26 shows the spurious signals generated when the mosaic is scanned in the dark. Fig. 26A is the waveform for a line, and Fig. 26B shows the waveform for a complete frame. If we neglect the voltages generated during the return time of both the line (these return time signals are blanked out by the blanking impulse introduced in the control room amplifiers) and frame scanning, the output voltage is a combination of an arc and a sawtooth. The output at the end of each line or frame is more positive than at the beginning. When a picture is focused on the mosaic, this combination arc and sawtooth voltage will cause the background of the picture to increase in brightness toward the bottom and right side. When there is a picture focused on the mosaic, the light and dark shades of the picture change the shape of the combination arc and sawtooth waveform.

In order to eliminate the uneven shading produced by these spurious signals, voltages of the same waveform and amplitude, but opposite in phase must be introduced in the control room amplifiers to cancel these spurious signals. Most of the spurious signals can be eliminated by introducing semi-sine and sawtooth waveforms. To correct the vertical shading, semi-sine and sawtooth waveforms with the same frequency as the vertical scanning frequency are required. To correct the horizontal shading, semi-sine and sawtooth waveforms with the same frequency as the horizontal scanning frequency are required. The methods of introducing these compensating voltages will be discussed in the lesson on control room amplifiers.

In the next few sentences we shall attempt to account for the origin of these spurious signals. The potential of the elements of the mosaic changes from -1.5 to +3 volts at the instant of scanning. Through the collection of secondary electrons knocked out of the mosaic, the potential of the elements returns to -1.5 volts. The time required for the potential to return to -1.5 volts is equal to approximately one-third of the scanning time for one frame.

Fig.27 Potential distribution over the surface of the mosaic when the scanning beam leaves the top of the mosaic.

When the scanning beam leaves the top of the mosaic (the image is inverted on the mosaic) to return to the bottom, the potential distribution over the mosaic along a vertical line will have the form shown in Fig. 27. In order for the top of the mosaic to return to a potential of -1.5 volts, it will have to collect secondary electrons knocked out of the bottom and center of the mosaic. Since most of the secondary electrons from the bottom of the mosaic will return to the bottom, the top will require a considerably longer time than one-third frame to return to a potential of -1.5 volts. Thus when the scanning beam reaches the top of the mosaic, the potential of the top has not returned to -1.5 volts, and has the distribution shown in Fig. 28. As the top of the mosaic is slightly more positive than the rest, the effect on the average level of the picture signal is the same as though the background of the bottom of the picture was brighter than that for the rest of the picture.

The waveform generated by the line scanning is probably due to the same condition. When the scanning beam leaves the left side of the mosaic to return to the right (the image is inverted on the mosaic), the left end will be more positive than the rest of the line. Most of the secondary electrons are collected by the elements just scanned by the beam. Thus when the beam jumps from the left side to the right, there will be insufficient electrons to restore the potential to -1.5 volts on the left side of the mosaic. The left side being more positive than the right and center, will cause the background level of the reproduced picture to be brighter on the right side than in the original.

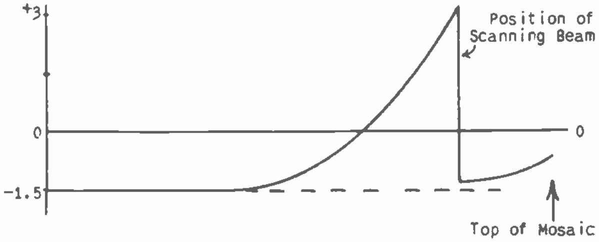


Fig.28 Potential over the mosaic as the scanning beam is approaching the top of the mosaic.

There is a sharp negative pulse generated during the horizontal return time (Fig. 26B). Since this impulse is blanked out by the blanking impulse which sets the black level (explained in the lesson on control room amplifiers), it is not important to know how it originates. The beam current is cut off during the vertical return time. This is necessary to prevent the return lines from appearing in the reproduced picture. If the return lines were not blanked out, the potential of the mosaic would be changed to +3 over the path followed by the beam in returning to the bottom of the mosaic. The return path would show up in the next frame or field as bright lines across the picture.

8. **KEYSTONE CORRECTION.** The electron gun in the Iconoscope is inclined at an angle to the surface of the mosaic (Fig. 21). The surface of the mosaic must be perpendicular to the principal axis of the lens so that every part of the optical image will be

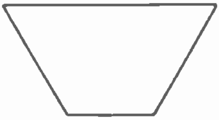


Fig.29 The pattern scanned by the electron beam in the Iconoscope without keystone correction.

in good focus. This requires that the electron gun be placed to one side of the mosaic. The distance from the gun to the top of the mosaic is greater than the distance from the gun to the bottom. Therefore, for the same sawtooth current amplitude in the horizon-

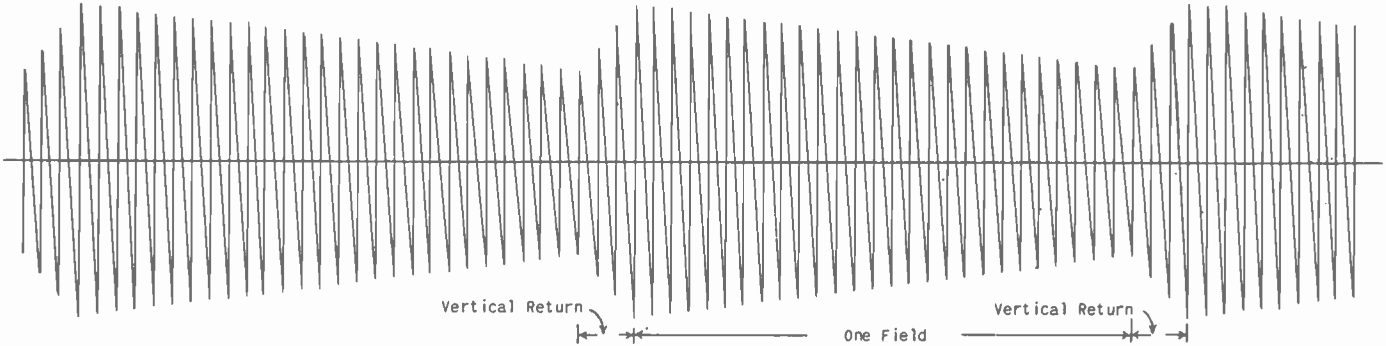


Fig.30 The current waveform of the horizontal sawtooth required to correct keystoning.

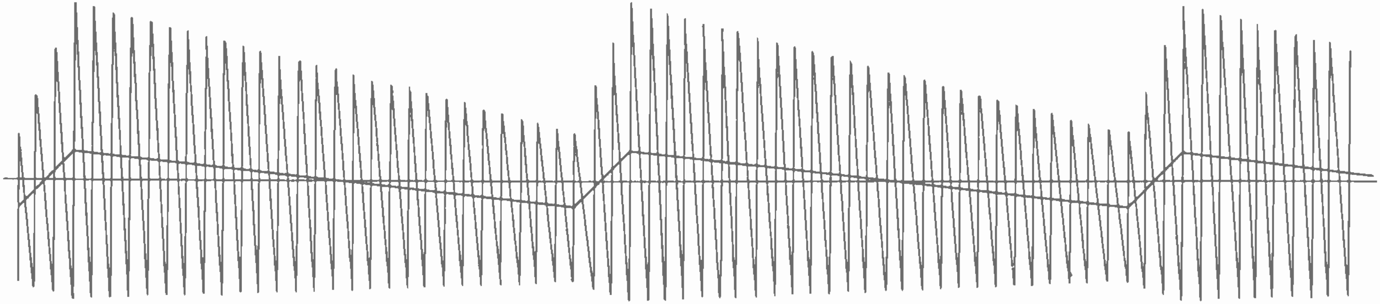


Fig.31 Distortion caused by the presence of modulating vertical sawtooth component in the modulated horizontal sawtooth.

tal deflecting coils, the scanning beam will cover a greater distance at the top of the mosaic than at the bottom. The resultant pattern would have the form shown in Fig. 29. In order to produce a rectangular pattern on the mosaic, the amplitude of the horizontal deflection must decrease linearly as the scanning beam moves from the bottom to the top of the mosaic. The required horizontal current waveform is shown in Fig. 30. The modulation envelope is a sawtooth of the same frequency as the vertical sawtooth. The waveform shown in Fig. 30 is obtained by amplitude modulating the 13,230 cycle horizontal sawtooth with the 60 cycle vertical sawtooth.

It is a rather difficult problem to obtain symmetrical modulation of the horizontal sawtooth. One factor that causes trouble is the elimination of the 60 cycle modulating component from the modulated sawtooth. The presence of the 60 cycle sawtooth component results in the *loss of symmetry of the modulation envelopes* as the vertical sawtooth forms the zero axis for the horizontal sawtooth. This effect is shown in Fig. 31.

Fig. 32 is a circuit of one method of accomplishing symmetrical modulation of the horizontal sawtooth. T_1 and T_2 are the vertical blocking tube oscillator and discharge tube respectively. The output of T_2 is amplified and applied to the vertical deflecting yoke. T_6 is the horizontal blocking tube oscillator. The horizontal discharge tube T_5, is a pentode. The horizontal sawtooth is developed across C_1. C_1 is charged through R_1 and R_2, and discharged through the pentode T_5. R_3 controls the amplitude of the horizontal sawtooth. The sawtooth developed across C_1 is amplitude modulated by the vertical sawtooth. The vertical sawtooth is applied to both the plate and the screen of T_5.

The vertical sawtooth that is used to modulate the horizontal sawtooth is developed across the condenser C_2 in the plate circuit of the discharge tube T_3. The grid of T_3 is connected to the grid of the vertical blocking tube oscillator. Thus the modulating sawtooth will have the same frequency and phase as the vertical sawtooth used for deflecting the electron beam in the Iconoscope but its amplitude can be controlled independently. The potentiometer R_5 controls the amplitude of this modulating sawtooth, and therefore the percentage of modulation of the horizontal sawtooth. The modulating sawtooth is amplified by the modulator T_4.

The output of T_4 is applied directly to the screen of T_5 and to the resistor R_2 through the condenser C_3. The component of the vertical sawtooth applied across R_2 adds to and subtracts from the voltage of the charging source of C_1. Thus the magnitude of the charge part of the cycle of the horizontal sawtooth developed across C_1 is varied in proportion to the amplitude of the vertical sawtooth across R_2. In order to make the modulation symmetrical, the magnitude of the discharge part of the cycle of the horizontal sawtooth must be made equal to the magnitude of the charge part of the cycle. The condenser is discharged through T_5 by the sharp positive grid pulses from the blocking tube oscillator. The amount of the discharge is determined by the plate resistance of T_5 during the discharge part of the cycle. Then, to vary the magnitude of the discharge part of the cycle in proportion to the charge part of the cycle, the plate resistance of T_5 during the discharge part of the cycle must be modulated by the same vertical sawtooth. The plate resistance must decrease when the voltage for charging C_1 increases. The plate resistance of T_5 can be varied by applying the vertical sawtooth to the screen as shown in Fig. 32. Since the vertical sawtooth is applied in the same phase to both the screen and plate circuits of T_5, the plate resistance decreases when the charging voltage of C_1 increases. The relative magnitude of the charge and discharge is controlled by the potentiometer R_6. Then the symmetry of the modulation can be controlled by R_6.

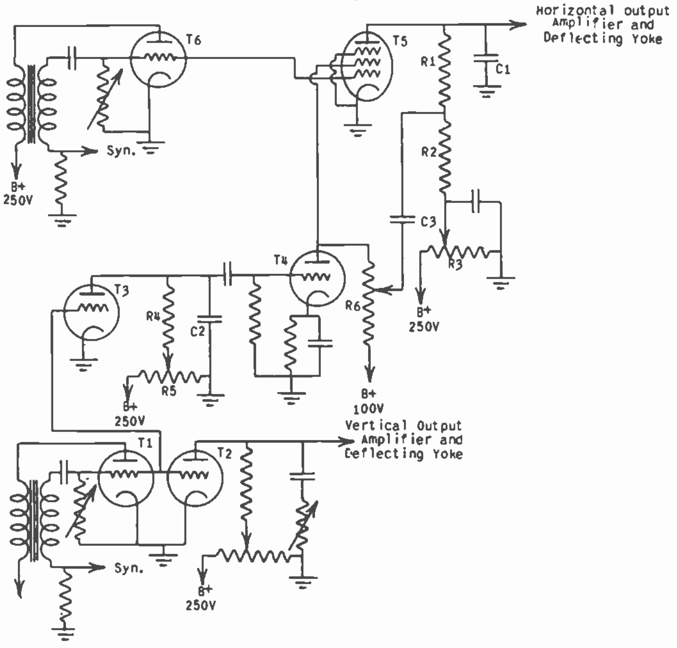


Fig.32 Circuit for amplitude modulating the horizontal sawtooth.

The modulated sawtooth developed across C_1 is amplified and applied to the vertical deflecting yoke. The circuits beyond C_1 are conventional. There is no peaking control for the horizontal in Fig. 32 but it can be added without destroying the symmetry of the modulation. No peaking will be required if the output circuit of the horizontal has a small inductance to resistance ratio.

There are other ways that symmetrical modulation can be obtained. One is to modulate the horizontal sawtooth in a push-pull stage and to combine the outputs of the two halves of the push-pull stage into a single-ended stage by the reverse of phase inversion. The vertical sawtooth component will be eliminated in the combination as it has the same phase in both sides of the push-pull stage.

9. THE TELEVISION CAMERA. A complete television camera is shown in Fig. 33. Fig. 34 is a schematic of the camera. Fig. 35 shows the Iconoscope mounted in the camera. The camera tripod is equipped with a "panning head"; that is, the camera can be tilted up or down and turned from side to side. The cameraman can check the optical focus by viewing the optical image on the mosaic in a mirror mounted in the top front of the camera (See Fig. 34). The

Fig.33 The television camera used by Midland Television.

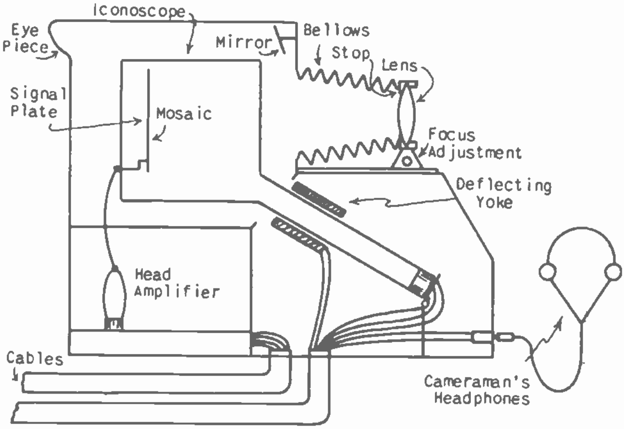


Fig.34 The construction of the television camera.

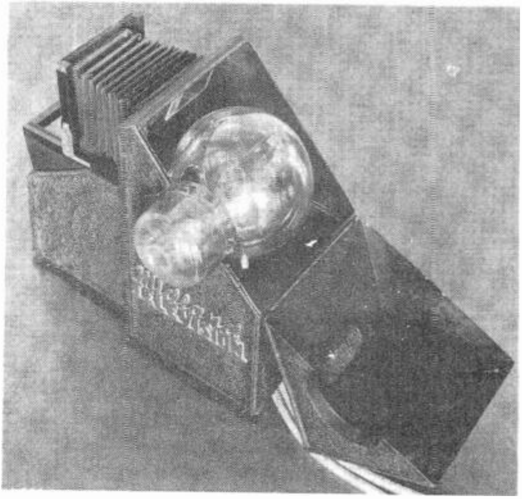


Fig.35 The Iconoscope placed in the camera.

focus of the optical image on the mosaic in this camera is controlled by the same method used in a conventional photographic camera. A three stage video amplifier (called a "head amplifier") is mounted in the lower rear of this camera (See Fig. 36). The output of the older model Iconoscope for a 10,000 ohm load resistance, is approximately one millivolt. Such a signal cannot be transmitted

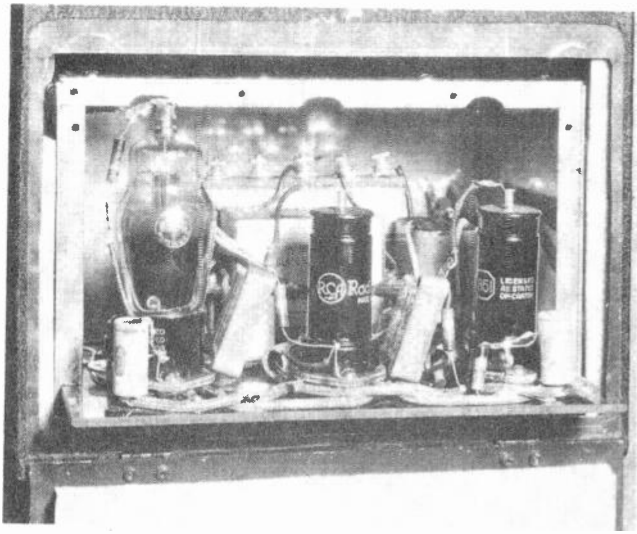


Fig.36 The head amplifier in the camera.

over a line to the control room amplifiers without excessive attenuation and noise pickup. The head amplifier has a voltage gain of 80 to 100. A complete description of the head amplifier will be given in the lesson on control room amplifiers. The head amplifier is very carefully shielded so that any pickup from the deflecting fields will be negligible. Fig. 37 shows the lower front of the camera. The termination of the cables from the control room, the deflecting yoke, and the tube socket for the Iconoscope are located in this compartment. A jack for the cameraman's headphones is mounted in the side of this same compartment. The cameraman receives his instructions from the control room operator via the headphones.

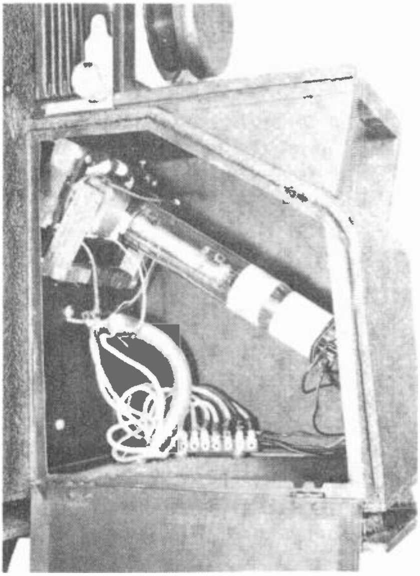


Fig.37 The termination of the cables the the location of the deflecting yoke in the camera.

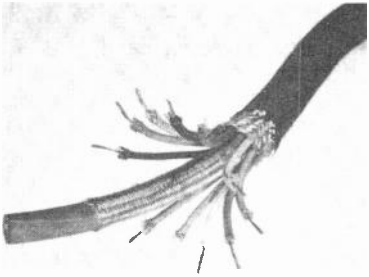


Fig.38 Construction of the cables used to connect the camera to the control room equipment.

Two cables connect the camera to the control room amplifiers. Fig. 38 shows the construction of this cable. Through the center of each cable is a low-capacity coaxial line. Around the central cable are several separate insulated conductors. The entire cable is surrounded by a metal shield and overall fabric cover. For this camera, one of the low-capacity cables conducts the picture signal to the first control room amplifier; the other carries the horizontal deflecting voltage to the yoke in the camera. Low-capacity cables are required for these signals in order to prevent attenuation of the high frequency components. Through the other conductors are transmitted the filament and plate voltages for the head amplifier, the filament and grid, and the first and second anode voltages for the Iconoscope; the vertical deflecting voltage; and the control room operator's instructions to the cameraman.

Fig.39 NBC television camera.

Fig. 39 is a picture of the television camera used by NBC. It is more elaborate than the one described, but the fundamental components are the same. In this camera, two identical lenses are used. One focuses an image of the televised object on the mosaic, while the other focuses an image of the same object on a ground glass screen. Since the lenses are identical and made to operate in unison; and since the image distances are the same, the image on the mosaic will be in focus when the image on the ground glass plate is in focus. The operator can adjust the focus more accurately when viewing the image on the ground glass screen than when viewing the image on the mosaic directly. The output stages of the deflecting circuits are often included in the camera housing if low-impedance deflecting coils are used.

A television camera using the dissector tube is very similar in design to one using an Iconoscope.

The high voltage power supply for the Iconoscope, the horizontal and vertical sawtooth oscillators, the associated output amplifiers, the keystone correction network, and the power supplies for the head amplifier are located in the control room. The operator has controls to vary the amplitude, linearity, and frequency of the sweeps; the symmetry and amplitude of the keystone correction, and the focus, intensity, and velocity of the electron beam in the Iconoscope.

10. THE IMAGE ICONOSCOPE. A new Iconoscope, called the "image Iconoscope", has recently been developed. It is about ten times as sensitive as the best model of the type previously described. The new tube uses secondary emission multiplication to increase the intensity of the output signal. Fig. 40 shows a picture of the new tube, and Fig. 41 is a schematic diagram of it. The optical image is focused on the translucent photoemissive cathode K. The photoemissive layer is deposited on a plane glass disc mounted in the front end of the envelope of the image Iconoscope. N is a thin metallic gauze cylinder of high-resistance material. The front end of the cylinder is connected to the photoemissive cathode K. The other end is grounded. The cathode end is operated at a few hundred volts below ground. The potential increases positively along the gauze cylinder toward the grounded end. The photoelectrons emitted by the cathode are accelerated toward the mosaic I, by this positive field. Around the tube is a magnetic focusing coil M. The magnitude of the current through this coil is adjusted so that an electron image of the photoemission from the cathode K, is focused on the mosaic I. The formation of this electron image is the same as in the dissector tube. The mosaic I, is not made up of photoemissive particles, but consists of a sheet of insulating material such as mica, that has been treated to have a high secondary emission ratio. S is a metallic coating on the back of the insulating sheet I, and is called the signal plate as in the previous model Iconoscope described.

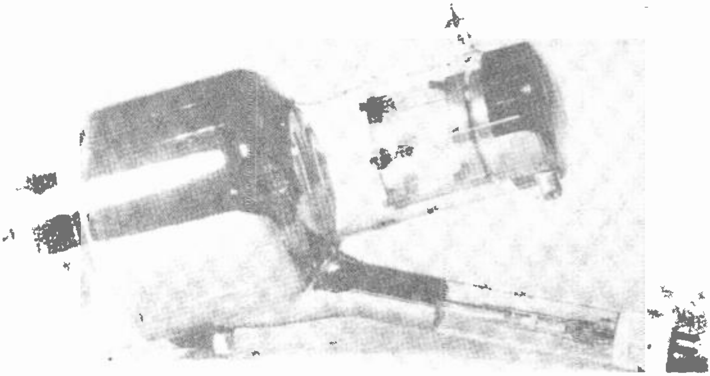


Fig.40 The new image Iconoscope.

The photoelectrons from the photoemissive cathode K, strike the mosaic I, and cause a large secondary emission. The secondary electrons are collected by the metallic coating A_2, on the inside of the tube. Since the mosaic is an insulator, the emission of the secondary electrons leave a potential distribution over the surface of the insulator that is an exact replica of the light and dark shades of the original optical image focused on the photoemissive cathode K. Since the insulator has a high secondary emission ratio, the charge stored on the mosaic is many times greater than the equivalent photoemission. The mosaic is scanned magnetically by an electron beam from the electron gun in the adjoining neck. The potential changes that take place on the insulator are similar to those produced in scanning the photoemissive particle type mosaic. There will be spurious signals generated as in the ordinary Iconoscope. However these signals make up a smaller percentage of the output of the image Iconoscope than in the ordinary type. The picture signal developed across R will also be negative.

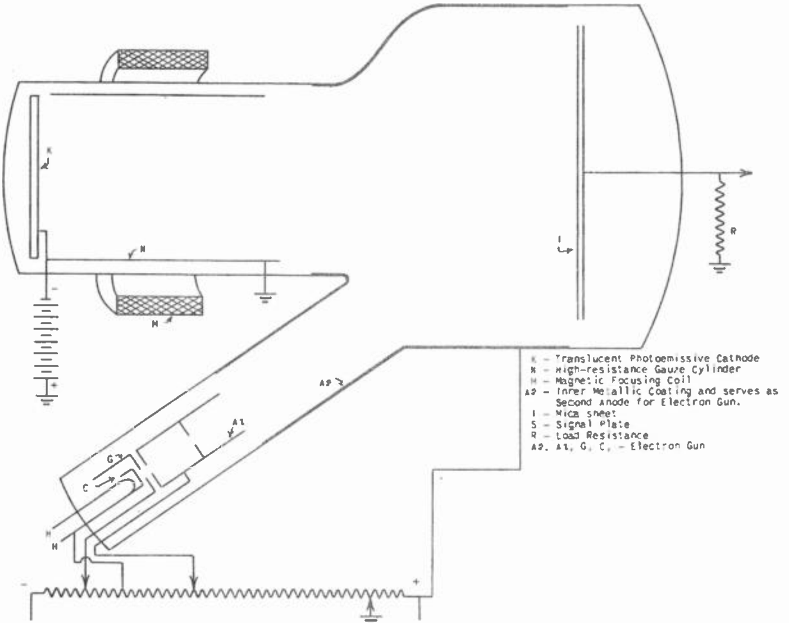


Fig. 41 The construction of the image Iconoscope.

The output of the image Iconoscope is ten times as great as the output of the best Iconoscope of the type previously described. This increase is due to the separation of the functions of photoemission and energy storage. The photoemissive cathode, since it is a continuous surface and not made of discrete particles, has more surface to emit photoelectrons and therefore the emission in microamperes per lumen will be higher than the emission from the particle type mosaic. Also, the photoemission from the cathode K, is saturated. The charge stored on the mosaic will be much higher through the secondary emission multiplication. Since the mosaic is an insulator, there will be no leakage between the particles or elements.

11. LENSES. In Lesson 3 of Unit 5, you studied the simple theory of the operation of lenses. It now becomes necessary to study the application and selection of lenses for use with the television camera. The television camera lens is the first link in the long chain from the televised scene to the reproduced picture on the cathode ray tube at the receiver. The quality of the optical image focused on the mosaic of the Iconoscope, or the photoemissive cathode of the dissector tube or image Iconoscope is determined by the excellence of the camera lens. The television circuits are incapable of compensating for defects existing in the optical image.

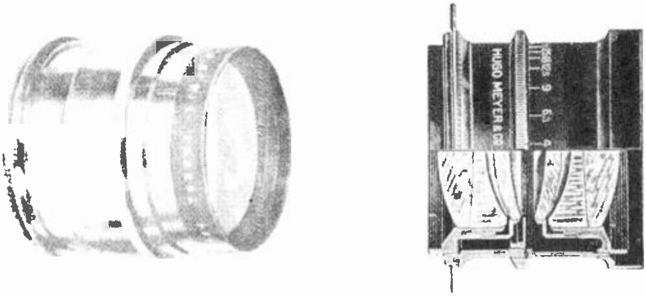


Fig.42 A high quality camera lens.

The type of lens used in the television camera is a converging lens. If the student has forgotten the laws giving the relation of the focal length to the object and image distances, and the relation between image and object size for converging lenses, a thorough review of the sections on lenses in Lesson 3 of Unit 5 is necessary in order to understand clearly the discussion that follows. A good lens is free of chromatic and spherical aberration and the various forms of astigmatism. An aperture of variable size is a standard part of a photographic or television camera lens. This aperture is known as a "stop". Fig. 42 is a photograph of a good lens. Fig. 43 is a cross-section of a standard lens. The lens is divided into two sections with the variable stop S, between them. The ring R, is to adjust the size of the stop. This location of the stop is to prevent barrel shaped and pincushion distortion. Each section consists of two or more lenses designed to correct for chromatic aberration, spherical aberration, and astigmatism. Combinations of diverging and converging lenses are used. As far as the operation of the lens is concerned, the combination can be considered as a single converging lens.

The brilliance, or the average light intensity per unit area of the images produced by lenses when the detail in the images is approximately the same, is a measure of the quality of the lens. The brilliance of the images produced by lenses is proportional to the "speed" of the lenses. The speed of a lens is equal to the ratio of its focal length to the diameter of the largest stop that can be used with the lens and still obtain a good image.

The formula giving the relation between the focal length of a lens and the object and image distances is:

$$\frac{1}{p} + \frac{1}{q} = \frac{1}{f}$$

Where p is the object distance, q the image distance, and f the focal length. When p is very large or is several times f, the image is located very close to the focal point, and f and q are practically equal. The ratio of the image size to the object size is equal to q/p, and for a large value of p, this ratio is equal to f/p. In other words, the size of the images produced by lenses are directly proportional to the focal lengths.

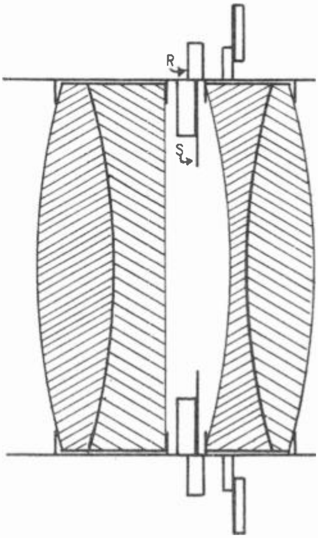


Fig.43 The construction of a camera lens.

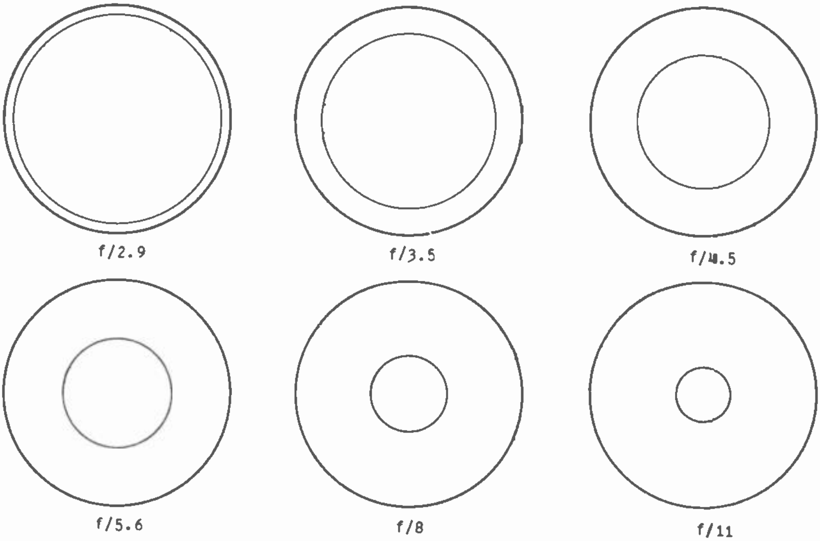
If two lenses have the same stop diameter but different focal lengths, the amount of light going through the two lenses will be the same, but the brilliance of the images will vary inversely with the square of the focal lengths. The same amount of light is being distributed over two different areas and the areas of two similar figures vary inversely with the squares of any two corresponding dimensions. (This statement is proved in any high school geometry). Therefore the brilliance of images vary directly with $1/f^2$.

If a single lens is considered, and the diameter of the stop is varied, the brilliance of the image will vary directly with the area of the stop. The larger stop will let more light pass through the lens. The areas of the stops vary directly with the squares of their diameters. Therefore the brilliance of images is directly proportional to d^2, where d is the diameter of the stop.

It is evident from the preceding discussion that the brilliance of the image for a lens is proportional to d^2/f^2. Therefore the speed of a lens is proportional to d^2/f^2. The speed of a lens is designated by F/a, where:

$$a = \frac{1}{\sqrt{d^2/f^2}} = \sqrt{\frac{f^2}{d^2}} = \frac{f}{d}$$

The d in the formula is the diameter of the largest stop that can be used with the lens and obtain a good image. The lenses used in the television studio of Midland Television have speeds of F/2.9 and F/4.5, and focal lengths of 6½ and 17 respectively, and of course F/4.5 is the slower lens. The speeds of any two lens are inversely proportional to the squares of their F/ numbers. Two lenses having


f/2.9 f/3.5 f/4.5

f/5.6 f/8 f/11

Fig.44 Showing how the relative size of aperture openings affect the speed of a lens.

the same F/ number will produce images of equal brilliance of the same object. The stop diameters for speeds of $\frac{1}{2}$, $\frac{1}{4}$, $\frac{1}{8}$, etc. of the fastest speed of the lens are indicated on every lens. The smallest stop marked is usually from F/45 to F/22. Therefore a lens having a maximum speed of F/2.9 may be stopped down to the slowest speed indicated on the lens (F/22 to F/45...See Fig. 44). It is important that the cameraman know the correct speed lens to use for different conditions. The two main factors controlling selection of the lens is the sensitivity of the Iconoscope or dissector tube and the intensity of the light reflected to the camera from the televised scene. The intensity of the light illuminating a scene is measured in foot candles. No surface reflects 100% of the incident light. The reflected light ranges from near 100% for a good plane mirror, to 0% for a perfectly black surface. The term "surface brightness" is used to describe the intensity of the light reflected from a surface. Surface brightness is measured in candles per square foot. The light reflected from a perfect reflector will have a surface brightness of one candle per square foot when the intensity of the illumination on the reflector is one foot candle. The intensity of a point source of light is measured in candle power. The intensity of the light reflected from a surface is measured in candles per square foot.

Present day Iconoscopes, when used with an F/4.5 lens, require an average surface brightness of approximately 15 candles per square foot to produce a satisfactory picture. A satisfactory picture is one in which interference caused by noise is not disagreeable to view, and one in which the spurious signals can be properly compensated. The average surface brightness required is less when the contrast in the televised scene is high. If an F/2.7 lens is used, the required average surface brightness is about 5 candles per square foot (the brilliance of the image varies inversely with the square of the speed of the lens). The average intensity of the light reflected from a scene can be measured with a foot candle meter such as the Weston model described in Lesson 5 of Unit 5.

The size of the image produced by a lens varies directly with the focal length of the lens. The size of the mosaic in the Iconoscope is fixed. If the object distance is kept constant and several lenses of different focal lengths are used, part of the scene falling on the mosaic of the Iconoscope, or the field of view of the lenses, will vary inversely with the focal length of the lens. Therefore, if the cameraman desires to televise one individual of a group (take a close-up) he will use a long focal length lens, and if he desires to televise the whole group he will use a short focal length lens. Another way of accomplishing the same operation with a single lens is to decrease the object distance to televise a close-up and increase the object distance to televise a group. In many cases this is impossible or inconvenient to do. It would be impractical for the cameraman televising a football game to get close to the scene of action. He must stay on the side lines, and in order to get a close-up of the play, he will use a long focal length lens (telephoto lens).

The image distance also varies with the focal length of the lens. Therefore, the shortest focal length lens that can be used with a television camera is determined by the mechanical construction of the pickup tube. The mosaic in an ordinary type Iconoscope is located three to four inches from the glass wall of the envelope. The focal length of a standard lens is measured from the center of the cylinder. The minimum image distance that can be obtained when using an Iconoscope is from five to six inches. This means that the shortest focal length lens that can be used is around six inches. Tubes like the dissector and image Iconoscope can be used with much shorter focal length lenses, as the photocathode is just inside the envelope of the tube. The long focal length lens required with the dissector tube for televising motion pictures is an advantage, as an enlarged image of the frame of the film is focused on the cathode.

The longest focal length standard lens that can be used with the television camera is limited by physical size of the camera. Lenses with focal lengths of 18 to 20 inches or more are desirable for getting close-ups of distant persons or objects. A 20-inch lens has an image distance of at least twenty inches. Using an ordinary 20-inch lens would make the length of the camera bellows so great that the lens would project several inches in front of the camera. It is difficult to mount such a bellows and prevent it from vibrating mechanically. This difficulty can be overcome by using "telephoto lenses". A telephoto lens is a lens that has an image distance that is shorter than the optical focal length. For example, the image distance for an 18-inch focal length telephoto lens focused on a distant object will be around 13 inches. Thus, there is a saving of five inches in bellows length. Fig. 45 shows a telephoto and an ordinary lens of the same optical focal length. The telephoto lens consists of a short focal length converging lens and a long focal length diverging lens. The diverging lens is inside the focal point of the converging lens. The diverging lens causes the rays of light to diverge slightly and the image is formed a little farther from the converging lens. This results in a larger image being produced. The images for the two lenses in Fig. 45 have the same size but the distance of the image from the converging lens is greater for the simple lens than it is for the telephoto lens. Thus there is a net saving in image distance or bellows length. Fig. 46 shows a picture of a telephoto lens.

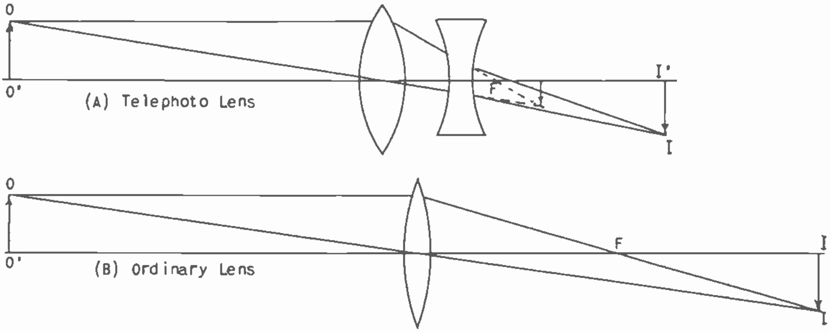


Fig. 45 The construction of a telephoto lens compared to an ordinary lens.

Fast lenses have a small "depth of focus". A lens has a small depth of focus when the images of objects in front and behind the object upon which the lens is focused are not sharply defined. A lens has greater depth of focus when the images of objects in front and behind the object upon which the lens is focused are quite sharply defined. The depth of focus of a lens can be increased by reducing the stop diameter of the lens. Of course stopping down the lens reduces the speed of the lens. Then in order to have the same intensity of the illumination on the televised scene must be increased. Fig.47 shows how the depth of focus increases as the stop size or speed of the lens is reduced. The lens used in obtaining the data for the curves had a speed of F/6.3. The lens was focused on an

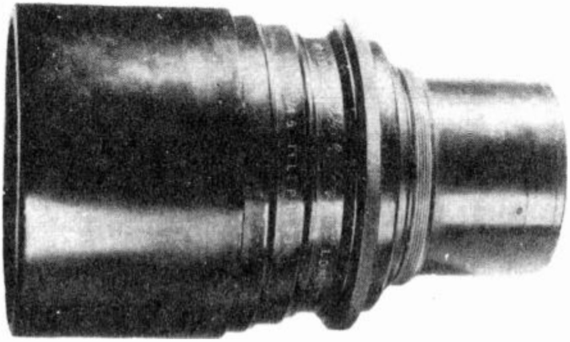


Fig.46 A telephoto lens.

object 3 feet away from the lens. Any object whose distance is greater than the minimum distance designated by the line D_1D_1, or is less than the maximum distance designated by the line DD, will be in focus.

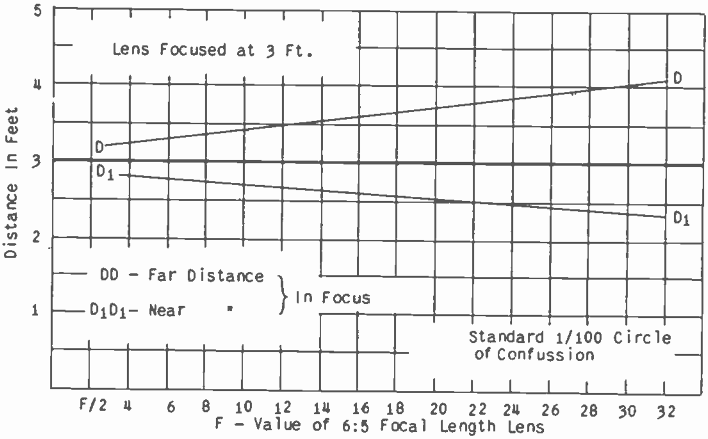


Fig.47 The increase of the depth of focus of a lens when the stop diameter or speed is reduced.

The reason that the depth of focus is increased when the lens is stopped down is shown in Fig. 48. Fig.48A shows the paths of the rays of light through the lens when the entire lens is used. Fig.48B shows the paths of the rays of light through the lens when just the center of the lens is used. In part A, the two rays from O on the object diverge rapidly after leaving point O. These two rays converge rapidly on the other side of the lens to point I on the image. In part B, the two rays diverge slowly from point O and converge slowly to point I. When the rays diverge slowly, their source is not sharply defined. Therefore the rays coming from points nearer the lens and farther from the lens than OO' will focus in the plane of the image II'. Also the focus will be good for the object OO' in planes nearer the lens and farther from the lens than the plane of the image II'.

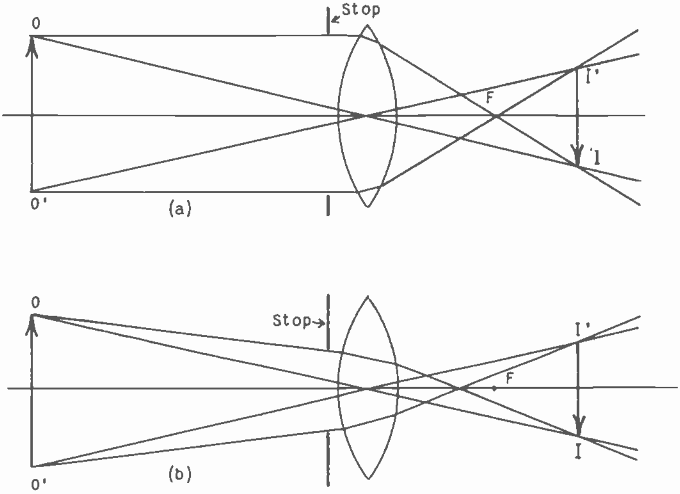


Fig.48 The depth of focus is increased by stopping down the lens.

12. PICTURE SIGNAL GENERATOR TUBES. Before closing the discussion on television pickup tubes, we shall describe a tube that generates a picture signal suitable for testing the operation of the video amplifiers and other circuits of a television system. These tubes transmit but one picture and this picture is incorporated in the tube. Fig. 49 shows the construction of one of these picture signal generator tubes. The picture is printed with a carbon ink on the aluminim plate T. The aluminum is left uncoated for the white parts of the picture. The picture is scanned by an electron beam. Either magnetic or electrostatic scanning can be used. The tube in Fig. 49 uses magnetic scanning. The picture signal developed across the load resistor R is due to the different secondary emission ratios of the carbon coated aluminum surface and the uncoated aluminum surface. The aluminum has a high secondary emission, while the carbon coated surface has a very low secondary emission. The secondaries are collected by the inner metallic coating C. C and the second anode A_2, are operated at ground potential. The target T, is operated at a negative potential with respect to

C so that all the secondaries will be collected; that is, the secondary emission is saturated. There is an electron current through R to the target T of the same magnitude as the secondary emission current from T to C. When the beam scans the aluminum or white, this current will be high, and when it scans the carbon coated aluminum or black, this current will be low. Thus the voltage developed across R will be most positive for white. In other words the polarity of the picture signal is positive.

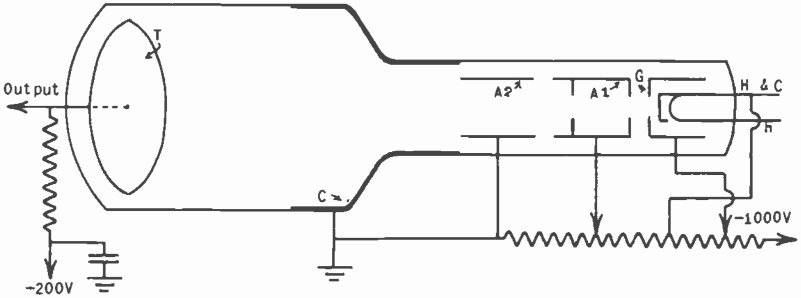


Fig.49 Tube for generating a picture signal from an internally incorporated picture.

These tubes are on the market with many kinds of pictures and test patterns incorporated in them. However, since the picture signal from an Iconoscope is negative, the pictures on the plate of the test tubes used in the Iconoscope for testing the system have the light and dark shades reversed, or the picture is negative. Since the secondary emission from the target T is saturated, there is no spurious signal generated. The picture signal is about as perfect as it is possible to generate.

12. TEST PATTERNS. The quickest and simplest way to check the operation of a television system is by means of a test pattern. Fig. 50 shows a test pattern that is suitable for this purpose. Some of the factors that can be readily checked by the use of such a pattern are: the linearity of the sweeps; aspect ratio; keystone correction for the camera; vertical and horizontal resolution; spot defocusing; and phase, frequency, and amplitude distortion of the amplifiers.

The test pattern is placed before the television camera and the image of the test pattern is focused on the mosaic of photocathode of the tube. The chart is spaced from the camera so that the image completely fills the usable area of the mosaic or photocathode.

If the scanning pattern on the cathode ray receiving tube is in perfect adjustment as to aspect ratio and linearity, the linearity, keystoning, and aspect ratio of the pattern on the plate of the Iconoscope can be adjusted by observation of the reproduced pattern on the receiving tube. Lack of linearity in the camera sweeps shows up as uneven spacing of the fine vertical and horizontal lines forming the squares, and also as distortion in the shape

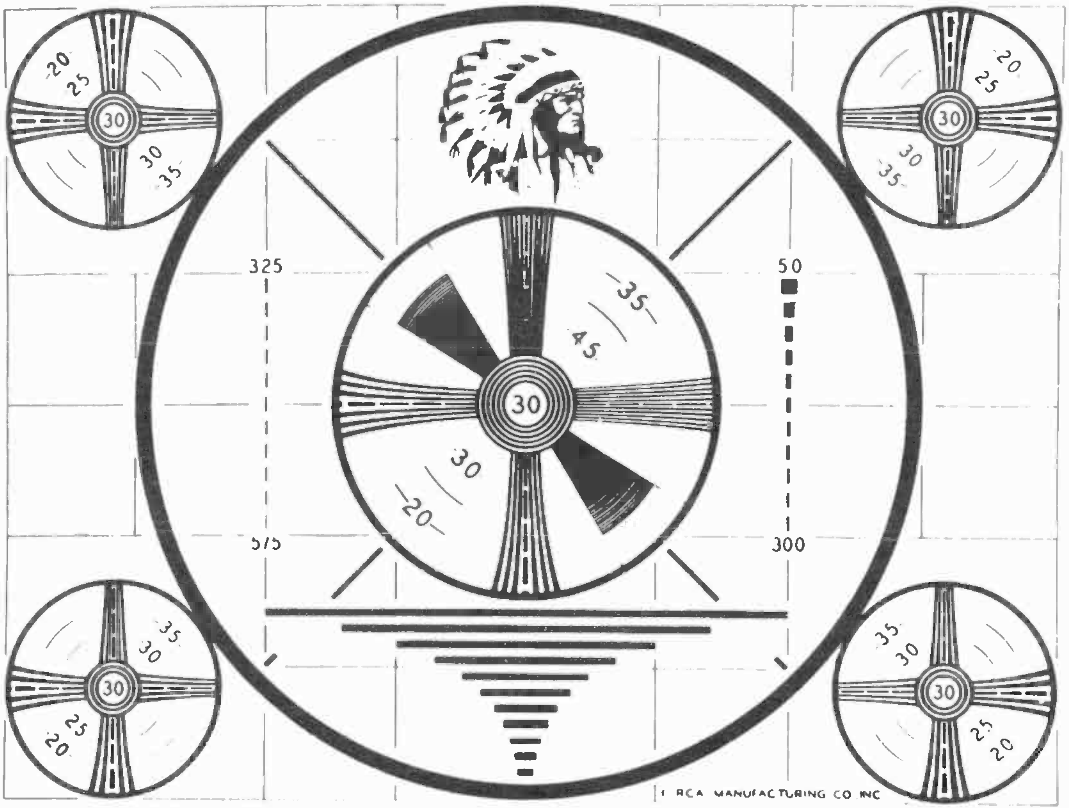


Fig.50 Test Pattern.

of the large circles. If the aspect ratio is incorrect, the squares will be reproduced as rectangles and the large circles as ovals. If the keystoning is incorrectly adjusted, the width of the pattern will decrease or increase toward the top, depending upon whether the keystoning is under or over corrected. Lack of symmetry in the keystone correction will show up as a wedge-shaped pattern that is tilted to one side or the other.

When the camera sweeps are in correct adjustment, the chart can be used to check the overall resolution of the system. The wedges in the center of the chart and in the four circles in the corners are used for checking the resolution in both the vertical and horizontal directions. Each wedge consists of a series of wedge-shaped black and white bars of equal size. The vertical wedges are used to check horizontal resolution and the horizontal wedges are used to check the vertical resolution. When the scanning pattern covers the entire image of the test pattern on the mosaic, the small numbers in the arcs of the circles between the wedges indicate the definition corresponding to the width of the black and white bars in the adjacent wedge. These numbers must be multiplied by ten. For example, the number 45 in the center group of wedges means that the size of the lines in the wedge at that point correspond to 450-line definition of the pattern. The definition or resolution increases toward the small end of the wedge. The resolution of the system is shown by its ability to reproduce the lines in the wedge as the definition increases. The maximum resolution of the system is indicated by a line across the wedge marking the point where the systems is just able to reproduce the separate lines in the wedge. Beyond that point, the wedges are reproduced as a solid color. If the small circles in the center of the pattern are reproduced perfectly, the system has at least 300-line resolution in both the horizontal and vertical directions. The wedges in the corners will reveal defocusing in the corners of the scanning pattern as a loss in resolution in the corners.

The chart can be used to check the transmission characteristics of the amplifiers in a television system. Poor high frequency response is revealed by low horizontal resolution and the lack of sharp vertical boundaries between black and white sections of the pattern. Poor low frequency response shows up as a change in background level from top to bottom of the pattern. Phase shift is revealed by the presence of uneven shading across any vertical black bars in the pattern. Also, the vertical black lines will have a sharp leading edge but a blurred trailing edge, and are followed by a white tail that blends into the background.

Amplitude distortion is revealed by the poor reproduction of the half-tone wedges set at a 45° angle in the center of the pattern.

Through experience with the use of test patterns, the control room operator can gain considerable information as to the condition of the entire system by the reproduced picture of the pattern.

Tubes like those described in the previous section, are available with test patterns incorporated in them. These will reveal any faults in the system exclusive of the television camera.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. What happens to the picture signal amplitude when the number of lines in the picture is increased?

2. Why can't ordinary vacuum tube amplifiers be used to amplify extremely weak signals?

3. What are the two solutions to the problem of obtaining a usable picture signal from a high - definition television picture?

4. What is the principle of the electron multiplier?

5. What is meant by the storage principle as applied to a television pick-up tube?

6. How is scanning accomplished in the dissector tube?

7. Describe briefly the principle of operation of the Iconoscope.

8. What are the essential components of a television camera?

9. What is "keystone correction", and why is it necessary?

10. What is meant by "depth of focus"? How can the depth of focus of a lens be increased?

Notes

(These extra pages are provided for your use in taking special notes)

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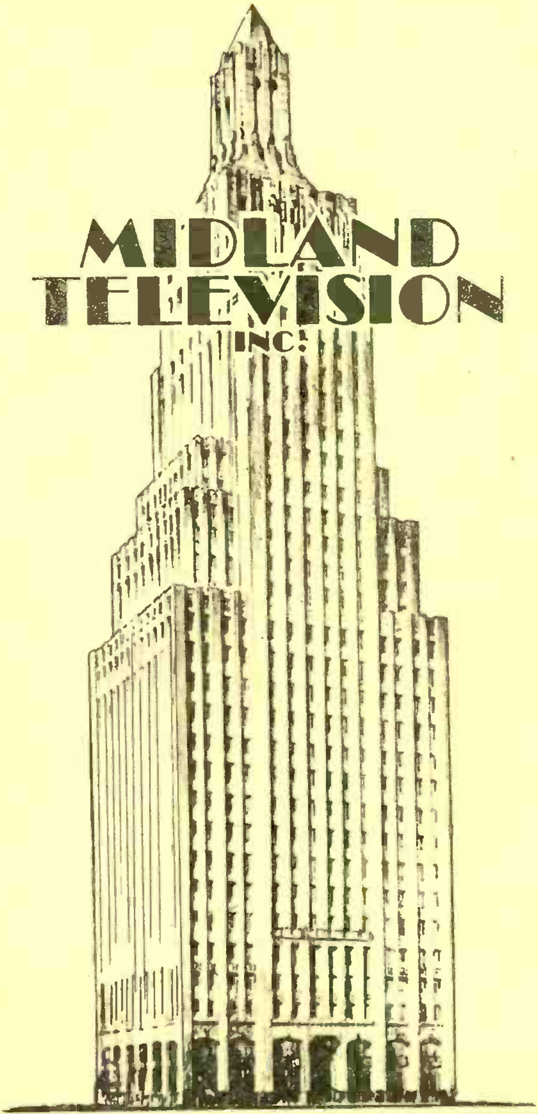


POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT
NO.
7

TELEVISION CONTROL ROOM AMPLIFIERS

LESSON
NO.
2

KNOCKS & BUMPS

.....they discourage some,
.....make others fightin' mad.

During the years that we have been guiding the destinies of ambitious men, we have had an excellent opportunity to study the varied reactions of our students to knocks and bumps.

Some 'knocks and bumps' are really tough. Others are imaginary. But regardless of whether they are 'tough' or 'soft', it is usually the man who is looking for an 'out' who quits when the going gets a little bumpy. The 'man' with a spine gets mad and fights it out.

Several instances of REAL backbone, stand out in our memory. One of these was a man 28 years of age, with a family to support. Misfortune had tracked him for years. His training period proved to be a trial for the stoutest heart. He did not catch on easily....progress was very, very slow. And, to support his family, he worked NIGHTS.

BUT HE HAD BACKBONE. HE GOT MAD AND FOUGHT 'ER OUT.

"What happened to him?" you ask. The same thing that usually happens to every man with courage. He completed his training with honors and A SMILE that was not dimmed in the least by "knocks and bumps". He created a favorable impression upon a prospective employer who visited our school. AND SHORTLY AFTER, HE WAS EMPLOYED AND MADE GOOD.

Here you have just one of MANY instances where the will to fight and stick it out, paid substantial dividends.

Lesson Two

TELEVISION CONTROL ROOM AMPLIFIERS

"Television control room equipment is the most important link in the complicated chain of apparatus needed to produce or transmit a television image; therefore this lesson will be one of the most important in this unit of your studies.

"Since there are as many different set-ups in this type of equipment as there are television stations, we will attempt to describe only a typical installation. In addition, we will cover the fundamentals of such equipment, so that you will be sufficiently acquainted with any installation you might encounter."

1. INTRODUCTION. This lesson will follow the picture signal from the signal plate of the Iconoscope through the control room amplifiers to the television transmitter. In this amplifier chain, the spurious signals generated in the Iconoscope are neutralized; the vertical and horizontal blanking and synchronizing impulses are inserted; and the DC component, which represents the average light level, is also inserted.

Before going into a detailed study of the functions of the various units in the control room amplifiers, it will be necessary to take up a more comprehensive study of the wide range amplifiers than was given in previous lessons. Therefore, the first part of this lesson will be devoted to a discussion of video frequency amplifiers.

2. NATURE OF THE TELEVISION SIGNAL. The waveform of a television signal is quite different from that of speech or music. In speech and music, the total range of frequencies involved is from 16 to 16,000 cycles. In modern high-definition television systems, the range of frequencies is from 0 to 4,000,000 cycles. In sound, the waveform changes slowly, and any particular formation is recurrent for several cycles. In television, the waveform changes very rapidly, and any particular formation may occur but once and not be repeated. In sound, the amplitude changes of the signal take place relatively slowly, while in television, the amplitude changes may take place almost instantaneously. It is far more difficult for communications channels to handle instantaneous voltage changes without distortion than those which take place relatively slowly.

In Fig. 1 is shown an oscillogram of the electrical waveform corresponding to the word "seems". In order to obtain an idea of the time interval involved and the rapidity of the amplitude changes, an oscillogram of a 500-cycle sine wave having a duration of one-tenth of a second is also shown in Fig. 1. The total time interval required to speak the word "seems" is thus seen to be approximately four-tenths of a second. Also, the waveforms corresponding to each letter are repeated for several cycles, and the amplitude changes take place relatively slowly.

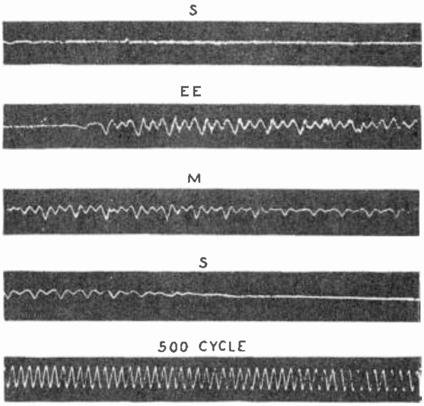


Fig.1 Waveform of the word "seems".

In Fig. 2, is shown a graph of the waveform produced in scanning one line of the picture of a woman's head. The total time interval for the line of the picture and the blanking and synchronizing impulse is slightly greater than 75 microseconds (.000075). This particular configuration is not likely to be repeated for any other line of the picture. The amplitude changes, especially those between black and white picture elements, and those between the background and the blanking impulse, take place very rapidly.

The tonal quality of a sound or musical note depends on the number, amplitude, and harmonic relation of the audio frequencies present in the sound or note. All of these separate frequencies are sinusoidal in form. Thus, any sound consists of a group of sine waves of different frequencies and amplitudes. The quality of the sound depends on the harmonic relation of the component frequencies. Therefore, the waveform in Fig. 1, corresponding to the word "seems", can be resolved into a number of sinusoidal waves of different frequencies. In Fig. 3A is shown the waveform of the note emitted by an organ pipe, and in Fig. 3B is a graph showing the sinusoidal frequencies and their relative amplitudes which make up the waveform shown in Fig. 3A.

The waveforms generated in scanning a television picture can also be resolved into sinusoidal waves of different frequencies and amplitudes. In Fig. 4B is shown the waveform produced in scanning one line of the pattern shown in Fig. 4A. The waveform for the blanking and synchronizing impulse has been omitted for simplicity. If an instantaneous return time is assumed, the television signal will consist of an AC voltage with the waveform shown in Fig. 4B.

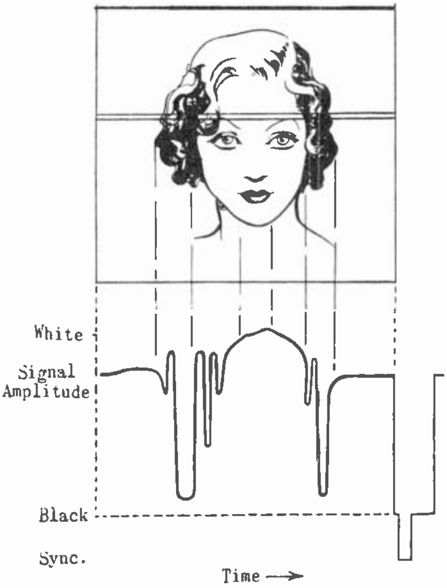


Fig.2 Waveform of one line of a picture.

Fig. 5 shows how this waveform can be built up by combining sinusoidal waves of the correct frequency, amplitude, and phase. Fig. 5A shows the fundamental component, which has the same frequency as the square wave, and the third and fifth harmonics of the fundamental with the correct amplitude and phase that they have in

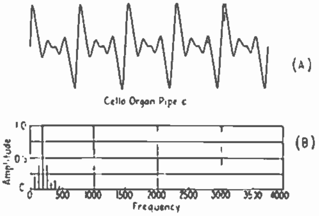
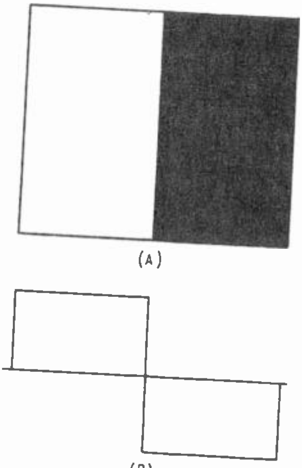


Fig.3 Waveform and component frequencies of the note emitted by an organ pipe.

the square wave shown in Fig. 5D. Fig. 5B shows the resultant waveform when the components in Fig. 5A are added together. You will note that there is considerable resemblance between this and the square wave in Fig. 5D. Fig. 5C shows the resultant waveform produced when the fundamental and the odd harmonics through the 15th are added together with the proper amplitude and phase that they have in the square wave. The ratio of the amplitude of each harmonic to the amplitude of the fundamental component is inversely proportional to the number of the harmonic; for example, the 15th harmonic will have 1/15 of the amplitude of the fundamental for a square wave. Fig. 5C is a fair approximation of the square wave. To construct the square wave exactly, will require the addition of decreasing amplitudes of all the odd harmonics through infinity. However, a very close approximation of the square wave can be obtained by adding together the proper amplitudes of the odd harmonics through the 25th or so. The lack of the higher harmonics results in the waveform having slightly rounded corners and a slight deviation from the instantaneous rise and fall of the leading and trailing edges of each wave.


(A)

(B)

Fig. 4 Pattern and corresponding waveform.

Any AC waveform that has reached steady state[1] conditions can be constructed by adding together harmonically related sinusoidal waves with the proper amplitude and phase. Whether even or odd harmonics or both are required, depends on the form of the wave that is to be constructed. The number of the harmonics that must be included, depends on the waveform that is to be constructed. Waveforms with sharply rising slopes and sharp corners, such as a square wave, or sawtooth with a fast return time, require very high order harmonics to reproduce them exactly. The simple waveform shown in Fig. 3 requires seven harmonics plus the fundamental to produce it exactly. The amplitudes of the lower harmonics may be greater or less than the amplitude of the fundamental (See Fig. 3B). However, the amplitudes of the higher harmonics diminisn as the frequency decreases.

[1] When the switch is first closed in an AC circuit, there are transients produced which modify the form of the current and voltage waveforms. The duration of the transients depends on the time constants of the circuit components. Steady state conditions exist in a circuit when all the transients, caused by the initial closing of the circuit, have died out. Transients are also produced in a DC circuit when it contains inductance or capacitance, or both.

If the waveform is not periodic; that is, it occurs once and is not repeated again, its structure is more complex. The component sinusoidal frequencies not only include the harmonics of the fundamental frequency, but also include bands of frequencies around the harmonic frequencies, bands of frequencies between the harmonic frequencies, and also bands of frequencies lower than the fundamental frequency. Figs. 6A, B, and C show the component frequencies and their amplitudes for three conditions involving square waves. Fig. 6A shows the component frequencies for an AC voltage of square waveform after steady state conditions have been established. Fig. 6B shows the component frequencies and their amplitudes for the condition that the waveform consist of but three square cycles; that is, the amplitude is zero before and after the interval required for the generation of the three cycles. For this condition, the component frequencies range from zero to the maximum recorded in the graph. However, the component frequencies exist in discrete groups or bands. Fig. 6C shows the component frequencies and their amplitudes for the condition that the waveform con-

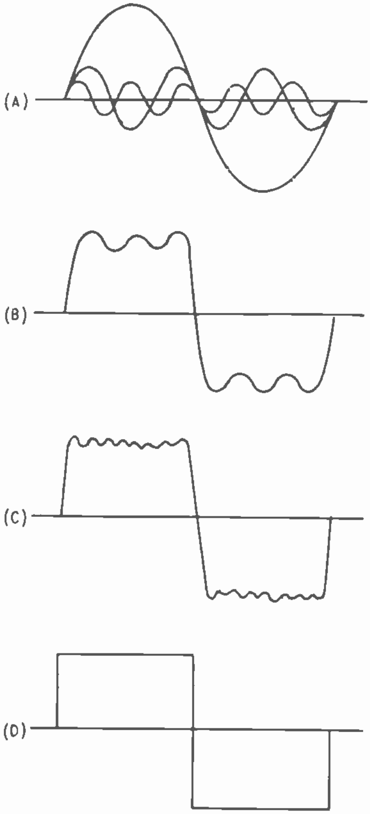


Fig.5 Synthesis of square wave.

sists of but one square wave cycle. The component frequencies range from zero upward, but the distribution of the frequencies is different from that for Fig. 6B. In Fig. 6D is shown the waveform and component frequencies that might be obtained in scanning one line of a television picture. The structure of the component frequency spectrum is quite complex as that particular configuration will probably occur only once per frame, and if the televised scene is in motion, it may never occur again throughout all time!

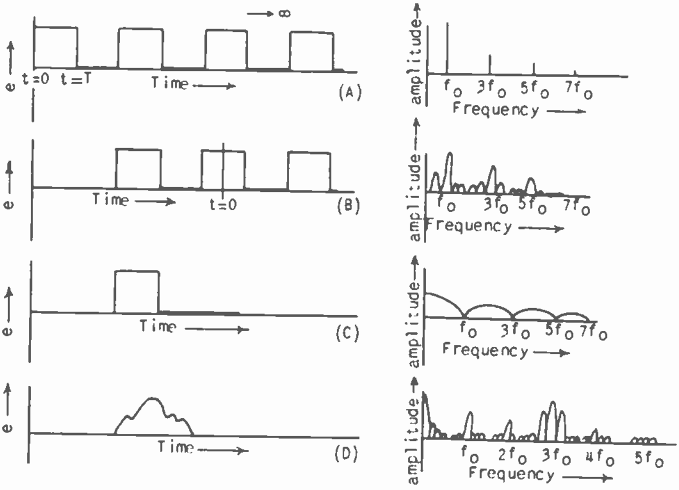


Fig.6 Different waveform types and their sinusoidal frequency spectrums.

The complexity of the component sinusoidal frequency spectrum obtained in scanning a picture, depends on two factors: (a) the number of times that a particular configuration is likely to occur during the scanning of a frame, and (b) the rate that the amplitude of the signal changes during scanning. When the scanned detail becomes of the same order of magnitude as the scanning aperture, the distortion of the resultant waveform due to the finite size of the scanning aperture produces a waveform of much simpler structure. Although the fundamental frequency will be high, the range of the component harmonics is very limited. If the highest frequency transmitted by a television system is limited to the maximum calculated by the formula $F = \frac{1}{2}A^2RN$, the distortion of the scanned waveform will not be more noticeable than that caused by aperture distortion.

In both sound and television, complex waveforms can be resolved into groups of sinusoidal components. The television waveforms have much greater complexity and cover a much wider range of frequencies. *In the case of television, the phase relations between the component sinusoidals is important because the shape of the impulse, and therefore, the detail in the picture, depends on the correct phase relations of the components.* In sound, the phase relations between the components is not important, as the ear detects sound in terms of frequency and amplitude, and not by the shape of the waves. Thus, during the process of transmission of a television signal, it is absolutely necessary to maintain the same phase relations between all component frequencies. This is not necessary in the case of sound, as the ear is unable to detect phase differences of the magnitude ordinarily found in conventional sound amplifiers.

3. DESIGN REQUIREMENTS OF VIDEO AMPLIFIERS. The three types of distortion encountered in amplifiers are phase distortion, amplitude distortion, and frequency distortion. In video amplifiers, phase distortion is far more objectionable than the other two. Very little phase distortion can be tolerated in an amplifier chain, while considerable amplitude and frequency distortion can occur before the effect becomes objectionable.

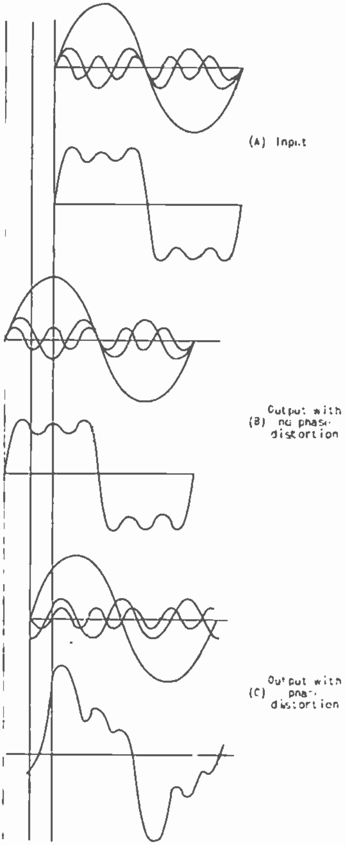


Fig.7 Output waveform of an amplifier with (B) no phase distortion, and (C) with phase distortion.

Phase distortion is minimum in an amplifier when the phase shift is directly proportional to the frequency. In other words, the phase difference between the output and input voltages must be proportional to the frequency. For example, if the phase difference at 1000 cycles is ten degrees, the phase difference at 10,000 cycles must be ten times ten degrees, or one hundred degrees, and similarly, the phase difference at 500 cycles must be one-half of ten degrees, or five degrees. When this condition exists, the waveform is unchanged by transmission through the amplifier chain, provided there is no amplitude or frequency distortion.

Figs. 7A and B show the phase conditions that must exist between the input and output voltages of an amplifier for no phase distortion. Fig. 7C shows distortion of the output waveform produced when these conditions are not fulfilled. Fig. 7A shows the input waveform and its component sinusoidals. These consist of the fundamental and the third and fifth harmonics, with the amplitudes that they have in a square wave (see Fig. 5).

Fig. 7B shows the output waveform when there is no phase distortion. The output fundamental component is leading the input fundamental by 90 degrees. The output third harmonic component is leading the input third harmonic by 270 degrees, and the output fifth harmonic is leading the input fifth harmonic by 450 degrees. In other words, the phase shift resulting from transmission through the amplifier, is proportional to the frequency.

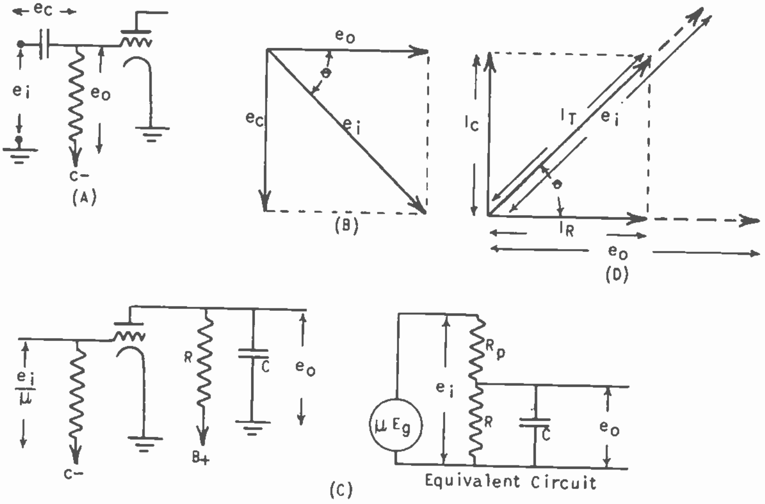


Fig.8 Two sources of phase distortion.

Fig. 7C shows the type of waveform distortion caused by the phase shift introduced by the grid condenser-grid leak combination of an amplifier. In Fig. 8A, the voltage e_o developed across the grid leak will lead the input voltage e_i by an amount determined by the ratio of the reactance of the coupling condenser to the resistance of the grid leak. From the vector[1] diagram in Fig. 8B, we see

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[1] Vector diagrams will be explained in the appendix of this lesson.

that this lead increases for lower frequencies, as the reactance of the coupling condenser is greater. For the condition shown in Fig. 8B, the reactance of the coupling condenser and the resistance of the grid leak are equal, and the voltage e_o leads the voltage e_i by 45 degrees. Also, the voltage e_o will be down to .707 of e_i. If the voltage e_i has the waveform shown in Fig. 7C, and the frequency of the fundamental component is that which will make the coupling condenser reactance and the grid leak resistance equal, the voltage e_o will have the waveform shown in Fig. 7C. This waveform was drawn on the assumption that there is no frequency distortion caused by the grid condenser-grid leak combination. The waveform distortion caused by frequency distortion is small in comparison to that caused by phase distortion.

Fig. 8C shows the circuit conditious at high frequencies. The load at high frequencies consists of load resistance R, and the capacity C, in parallel. C includes the output capacity of the tube, the input capacity of the following tube, and the wiring capacity to ground. Fig. 8D shows a vector diagram of the circuit when the capacitive reactance and the resistance in the load are equal. For this condition, the gain at high frequencies is down to .707 of the gain at intermediate frequencies. The current I_R through the resistance part of the load is in phase with the output voltage e_o across the load. The current I_C through the capacitive part of the load leads the voltage e_o across the load by 90 degrees. The tube current I_T is the vector sum of I_R and I_C. Since in television amplifiers, R_p is many times larger than R, the input voltage e_i will be in phase with the tube current I_T. Since I_T leads I_R by the angle θ, the output voltage will lead the equivalent voltage e_i by the same angle. (Of course e_i is μ times as great as the input voltage to the tube, and is opposite in phase to the input voltage.) Therefore, at high frequencies, the higher frequency components lag the lower frequency components and, when the capacitive reactance in the load becomes equal to the resistance in the load, this lag amounts to 45 degrees. Then Fig. 7C can also represent the phase distortion produced at the high frequency end of an amplifier's response if the attenuation produced by the shunt capacity is neglected.

Another expression that is used to describe the condition for minimum phase distortion is that the amplifier has "constant time delay" for all frequencies. This means that the transmission velocities of all the frequencies are slowed up equally during transmission through the amplifier.

The characteristics of a picture element will be almost completely destroyed if its component frequencies are spread over a distance equivalent to two picture elements. Since the time of scanning one picture element is approximately .13 microseconds, the total variation in time delay for the range of frequencies required to transmit a 441-line picture should not exceed .13 microseconds. This includes the delay occurring in the whole amplifier chain from the camera to the cathode ray tube in the receiver.

There are other causes of phase distortion than that described in preceding paragraphs. In fact, the phase distortion caused by the grid condenser-grid leak combination is corrected by introducing a phase distortion of the opposite sense. Phase distortion and its correction will be described in more detail in a later part of this lesson.

Frequency distortion results when the gain of an amplifier is not uniform over the required frequency range. In general, when frequency distortion exists in an amplifier, the gain is less at the very low and at the very high frequencies of the required frequency range. Fig. 9 shows the distortion produced when frequency distortion is present. In this particular figure, the effects of poor low frequency response and poor high frequency response have been separated. Fig. 9A shows the input waveform to an amplifier.

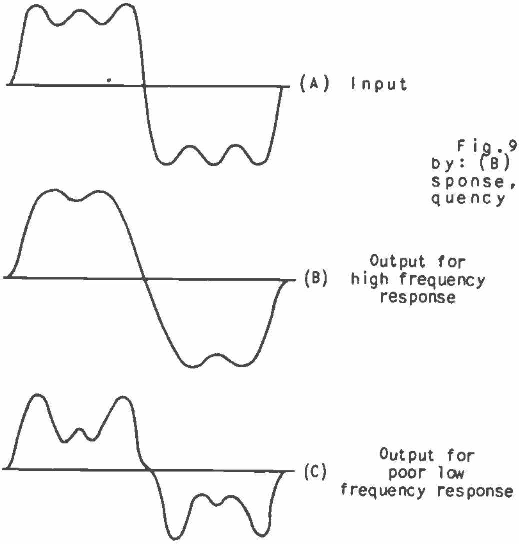


Fig.9 Distortion produced by: (B) poor high frequency response, and (C) poor low frequency response.

This waveform consists of the fundamental, and the third and fifth harmonic component sinusoidals of a square wave. Fig. 9B shows the waveform distortion produced when the fifth harmonic is not passed by the amplifier, and the gain for the third harmonic is down fifty per cent when compared to the fundamental. Fig. 9C shows the waveform distortion produced when the gain for the fundamental component is fifty per cent less than for the other component frequencies. When the high frequency response is poor, the steepness of the front and rear edges of the wave is reduced and the corners become more round. When the low frequency response is poor, the amplitude in the center of the wave pulse is reduced. In plotting these waveforms, only the effect of frequency distortion has been considered.

Amplitude distortion exists in an amplifier when the amplitude of the output is not proportional to the amplitude of the input. It is caused by the lack of linearity in the operating region of the tube characteristics or poor power supply regulation. Fig. 10 shows the effect of amplitude distortion on a sine wave. The shape of the wave is changed. If the effect is due entirely to amplitude distortion, the change in shape means that the amplifier has introduced new frequency components that were not present in the input waveform.

In discussing the various types of distortion encountered in amplifiers, we have considered each separately, Actually, the phase and frequency distortion occur together. When an amplifier has a flat high frequency response over the desired range, there is very little phase distortion occurring over that range. Conditions are somewhat different for the low frequency end of the response characteristic. Constant time delay, or minimum phase distortion does not occur simultaneously with uniform low frequency response. Since constant time delay is more important than uniform frequency response, the low frequency end of a video amplifier is designed to have constant time delay, or minimum phase distortion.

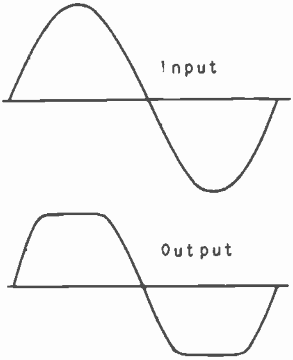


Fig.10 Distortion produced by amplitude distortion.

If we examine the waveform shown in Fig. 2, for one line of a television picture, we see that it is similar to a square or rectangular wave, as it has very rapid amplitude changes and has sections where the amplitude is constant for a considerable interval (relatively speaking). Therefore, an amplifier that will amplify square waves without distortion, will amplify television picture signals without distortion. The square waves used in testing amplifiers must have their fundamental frequencies and major components included in the frequency range produced in the scanning of a modern high-definition television picture.

The lowest frequency encountered in modern television systems is the field frequency, or 60 cycles. Therefore, an amplifier that will pass a 60-cycle square wave without distortion, will have satisfactory low frequency characteristics for the undistorted transmission of television signals. Fig. 11 shows the output waveforms produced for an input square wave when the amplifier has the low frequency characteristics with respect to gain and phase shown in the adjacent columns. The frequency of the square wave is designated by f_o. The phase characteristic is plotted in terms of the delay in microseconds versus frequency. The increase in delay is plotted along the vertical axis. If we examine Fig. 8B, it will

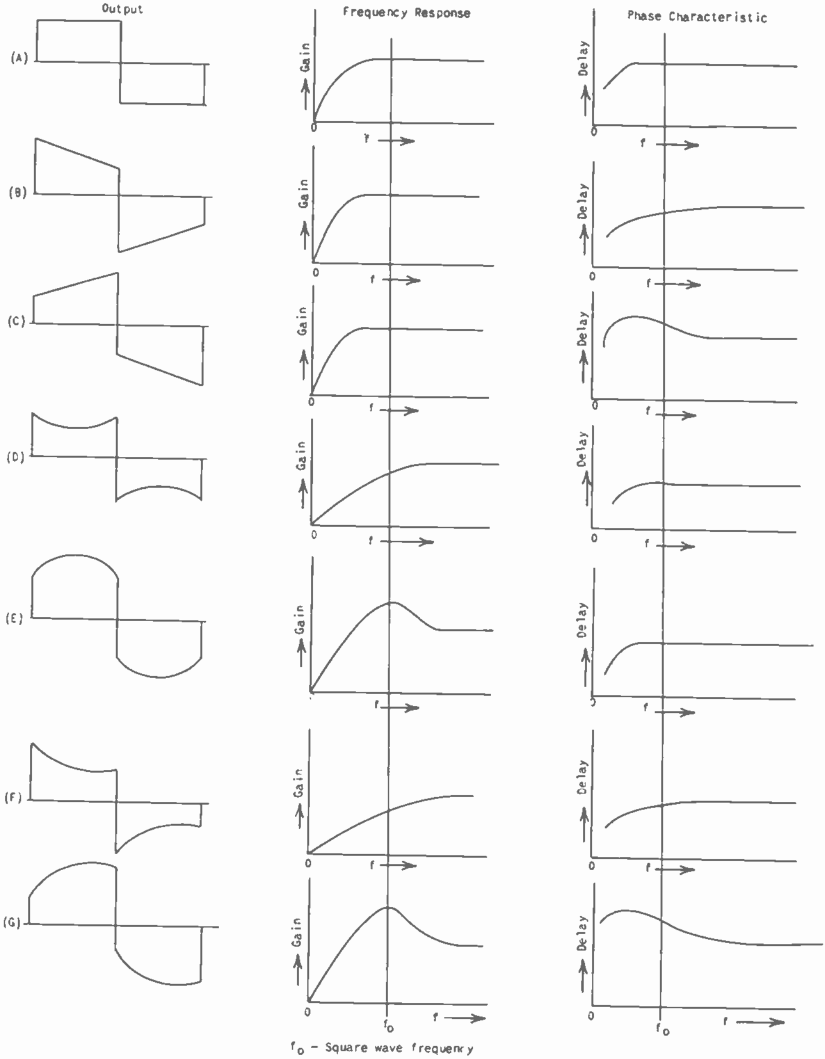


Fig.11 Output waveforms and the corresponding low frequency gain and phase characteristics.

be evident that an advance in phase will correspond to smaller delay in transmission through the amplifier.

The highest frequency encountered in modern television systems is approximately 4 megacycles. An amplifier that will pass a 100,000-cycle square wave without distortion, will have satisfactory high frequency gain and phase characteristics. The phase and gain characteristics in the intermediate range of frequencies from four or five hundred cycles to one or two hundred thousand cycles, do not usually require checking. The low end of this intermediate group is sufficiently high so that the reactance of the coupling condenser is negligible, and the highest frequency in this range is sufficiently low so that the reactance of the shunt capacity is many times larger than the load resistance.

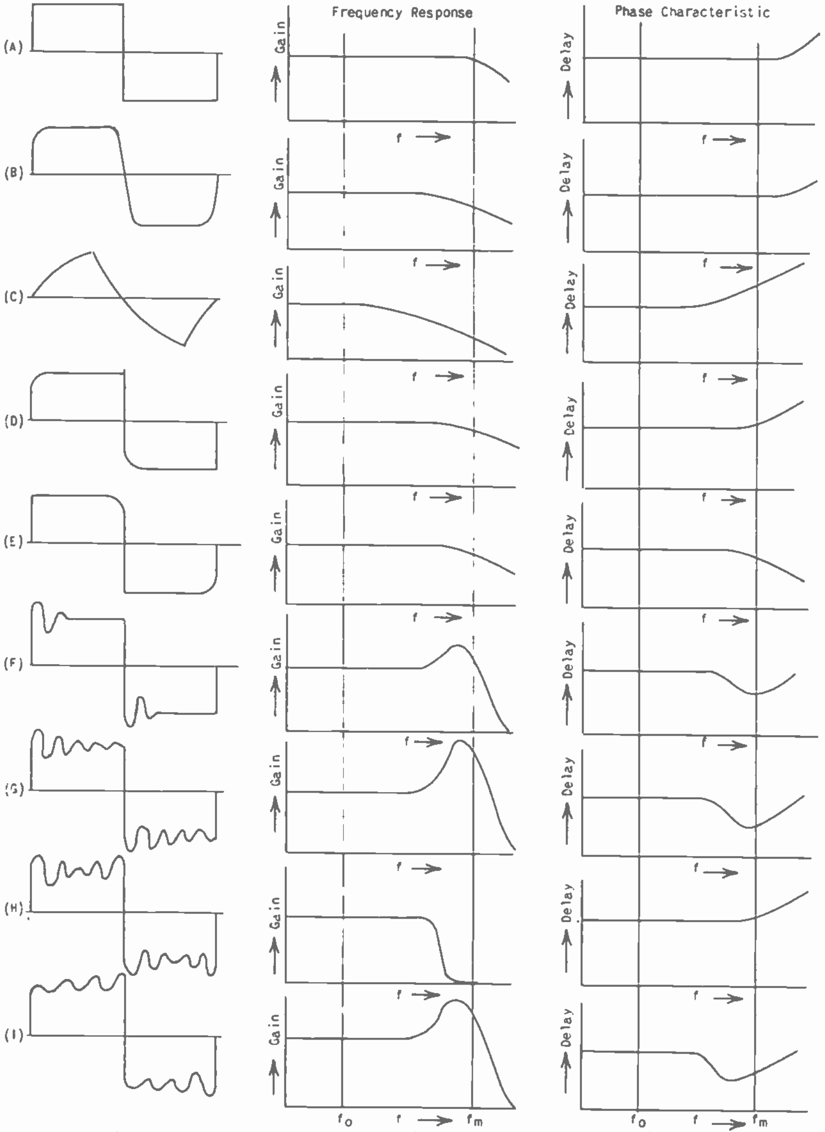
In a previous lesson, you learned that the loss in gain at the high frequencies, caused by the shunting effect of the reactance of the tube and wiring capacities, could be overcome by inserting inductance in series with the plate load resistance. You also learned that, for a given load resistance, the maximum frequency passed by a compensated amplifier stage without loss in gain, was that frequency which made the shunt capacitive reactance equal to the load resistance. For this condition, the inductive reactance of the compensating coil at the maximum frequency, was equal to one-half of the capacitive reactance.

The plate load of a simple compensated amplifier stage is a parallel tuned circuit with considerable resistance in the inductive arm. You are familiar with the fact that a damped oscillation is produced in a tuned circuit when a sharp voltage pulse is applied to it. The rate that the amplitude of the damped oscillation is attenuated depends upon the amount of resistance in series or shunt with the inductance. By increasing the magnitude of the series resistance sufficiently, the oscillation is suppressed. The circuit is said to be critically damped when the resistance has been increased to the point that the oscillation is just suppressed.

In a correctly compensated amplifier stage of the type that you have studied, the parallel tuned circuit formed by the shunt capacity and the compensated inductance, is critically damped by the load resistance. If the load resistance is reduced, and the compensating inductance is unchanged, the circuit is no longer critically damped, and the parallel tuned circuit will produce a damped oscillation whenever a voltage having a steep slope is impressed on the amplifier stage. The frequency of the oscillation is determined by the LC product. Whenever the load resistance is reduced below critical damping, it becomes less than the impedance of the parallel tuned circuit at resonance. Therefore, the amplifier stage will have maximum gain at the resonant frequency of the load circuit.

Fig. 12 shows the distortion of a square wave produced by amplifiers with different phase and gain characteristics over the high frequency part of their response. In Figs. 12A, B, C, D, and E, the damping is critical, or more than critical. In Figs. 12F, G, H, and I, the damping is less than critical, and damped oscillations accompany rapid voltage changes. The appearance of Fig. 12H is due to the sharp cutoff in the frequency response. In Fig. 12I, positive feedback is present, and the oscillation builds up instead of decaying.

Most of these causes of high and low frequency distortion can be recognized by observing the reproduced picture. The effects of low frequency distortion will show up as uneven shading along the vertical axis of the picture. Also, the vertical return lines will not be completely blanked out. The most common type of low frequency distortion in the video amplifiers is that illustrated in Fig.



f_o - Square wave frequency f_m - Maximum frequency of television signal

Fig.12 Distortion resulting from different high frequency gain and phase characteristics.

11F. This type of distortion is caused by poor low frequency gain and less delay at the lower frequencies.

The waveform generated in scanning a solid white background, if the horizontal blanking and synchronizing impulses are omitted, will have the form shown in Fig. 13A. Fig. 13B shows this same waveform after transmission through an amplifier with the phase and gain characteristics shown in Fig. 11F. This means that the reproduced picture will decrease in brightness from top to bottom. The blanking impulse may not suppress the vertical return lines completely near the top of the picture. The other types of low frequency distortion can be recognized by comparing the shading of the picture to the waveforms shown in Fig. 11.

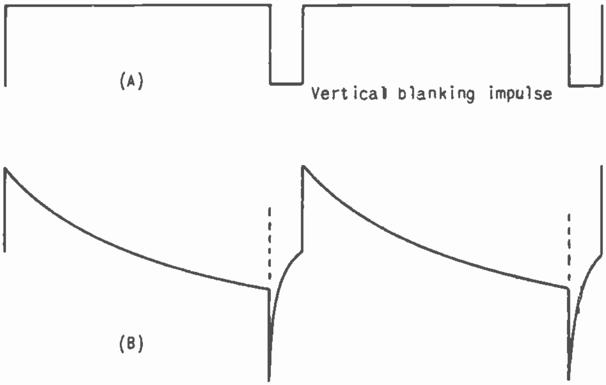


Fig.13 (A) Waveform generated in scanning a white background, and (B) the distortion of the waveform produced by an amplifier.

The effects of high frequency distortion will show up as modifications of the picture detail. Fig. 14A is the waveform obtained in scanning one line of a pattern consisting of a series of vertical black bars of various widths on a gray background. Fig. 14B shows the same waveform in the output of an amplifier that has the high frequency phase and gain characteristic shown in Fig. 12B. The amplifier has considerable attenuation at the higher frequencies. The boundaries between gray and black in the reproduced picture are not sharp, as the gray and black gradually merge. The very narrow black bar does not appear in the reproduced picture.

Fig. 14C shows the same wave after transmission through an amplifier that has the high frequency gain and phase characteristics similar to those illustrated in Fig. 12F. Each change between gray and black is followed by a highly damped oscillation. The oscillation is not very noticeable when the signal changes from gray to black, as the peak swing of the oscillation is blacker than black. However, the oscillation shows up as a line whiter than the gray background when the signal changes from black to gray. The narrowest bar appears in the reproduced picture as a black bar followed by a white bar of equal width.

In a similar manner, the other forms of distortion caused by unsatisfactory gain and phase characteristics at high frequencies can be recognized by careful observation of the picture.

When there is amplitude distortion present in an amplifier chain, it shows up as a non-linear reproduction of the shades between black and white in the picture. In other words, amplitude distortion affects the contrast of the picture. Fig. 15A shows the waveform generated in scanning one line of a pattern consisting of vertical bars. The shades of the bars change in equal steps from black to white. Fig. 15B is the same waveform after transmission through an amplifier that has amplitude distortion. In the reproduced picture, the shades of the bars do not change in equal steps

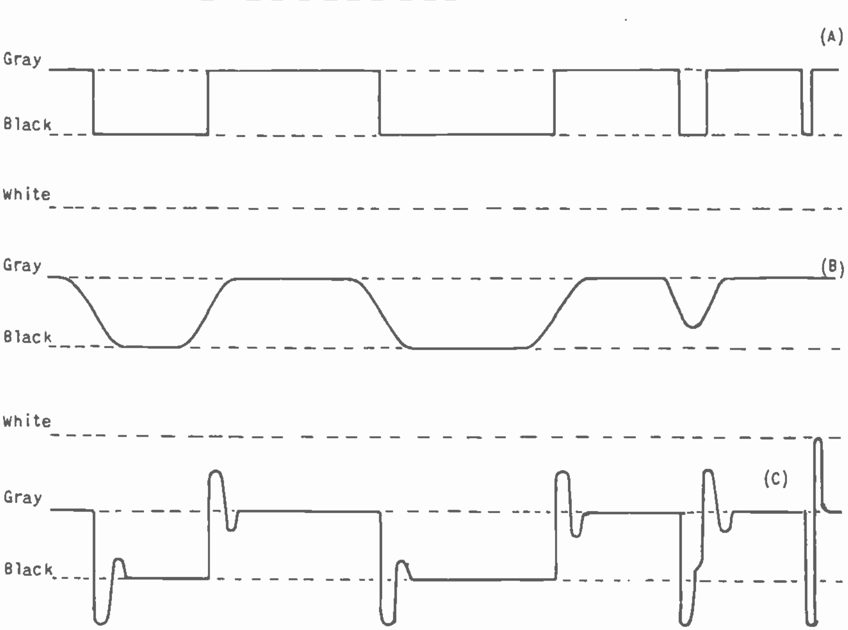


Fig.14 (A) Waveform generated in scanning vertical black bars on a gray background, (B) distortion of the waveform caused by low high-frequency gain, and (C) the distortion caused by excessive high-frequency gain.

from black to white. The contrast for the whiter shades is much less than that for the blacker shades. Amplitude distortion can cause a reduction in contrast in the blacks as well as for the whites. It is possible to have the contrast in the reproduced picture reduced in the extreme blacks and whites, while the intermediate shades are reproduced correctly. As stated previously, amplitude distortion is caused by curvature in the operating range of the tube characteristics in one or more stages. Fig. 15C is the overall characteristic for the amplifier which produces the amplitude distortion illustrated in Fig. 15B. The change in output voltage is plotted against the input voltage.

The presence of the amplifier distortion can be detected very readily be means of a test pattern as described in the lesson on television cameras. The student should study the diagram of the test pattern given in the lesson and determine for himself how the various parts of the pattern can be used to detect the different types of amplifier distortion.

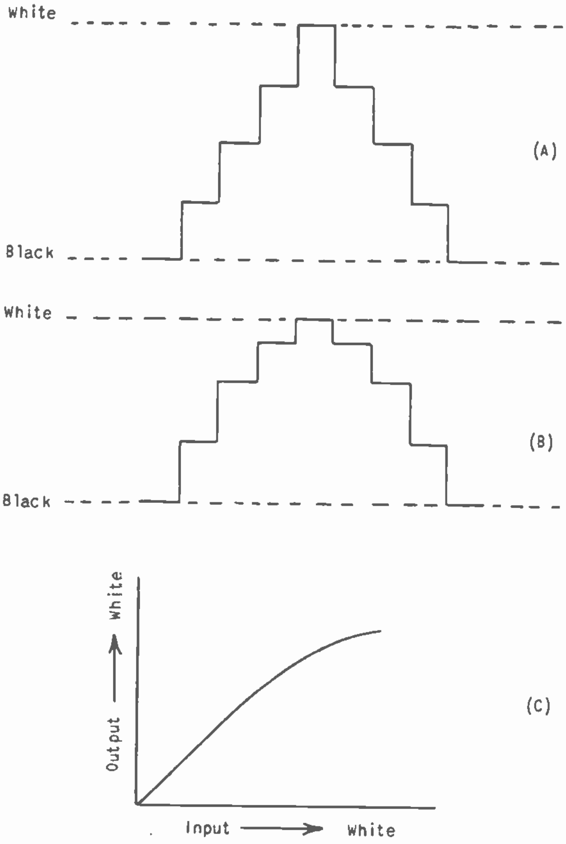


Fig.15 Effect of amplitude distortion.

4. HIGH FREQUENCY COMPENSATION OF AMPLIFIERS. The next few paragraphs will be devoted to a study of methods suitable for obtaining the correct phase and gain characteristics over the high frequency range of video amplifiers. As stated previously, the phase characteristics are usually satisfactory over the high frequency range when the gain is uniform over the desired range. Therefore, only the methods used in obtaining the proper gain characteristic will be discussed.

In a previous lesson you studied one method of compensating an amplifier to obtain a flat response over the high frequency range. We shall review this method briefly. Fig. 16A shows the coupling network between two tubes in a conventional resistance-coupled amplifier. C_1 and C_2 represent the input and output capacities of the tubes, and the wiring capacity to ground. C_2 includes the grid-to-plate capacity of the second tube, which is multiplied by a factor determined by the voltage gain of the second tube. You recall from a previous lesson that the input capacity of a tube was given by the expression $C_{gc} + C_{gp}(1 + A)$, where C_{gc} is the grid-to-cathode capacity, C_{gp} the grid-to-plate capacity, and A is the voltage gain of the second tube. Fig. 16B shows the equivalent circuit. The load impedance that the first tube works into consists of a resistance and capacity in parallel. Thus the load impedance, and therefore the gain, becomes less as the frequency is increased. To minimize the variation of the load impedance with frequency, the load resistance is made small. To overcome the shunting effect of the capacitive part of the load at higher frequencies, a small inductance is added in series with the load resistance (see Fig. 17). The highest frequency that can be passed by this type of compensated stage without loss in gain is the frequency at which the capacitive part of the load will have a reactance equal in magnitude to the resistive part of the load. For correct compensation, the inductive reactance of the compensating or peaking coil must be equal in magnitude to one-half of the capacitive reactance at the top frequency that can be passed without loss in gain.

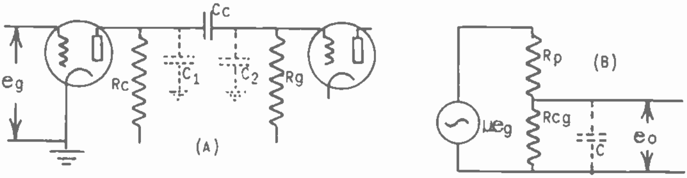


Fig.16 Circuit of an ordinary resistance-coupled amplifier at high frequencies.

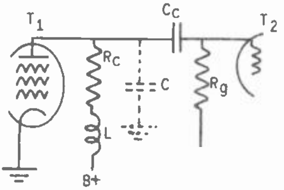


Fig.17 Shunt peaking circuit.

Therefore, the total capacity in shunt with the load resistance determines the top frequency that can be passed by an amplifier stage without loss in gain. If we wish to design an amplifier stage so that it will have uniform gain up to the frequency f_m, we select a load resistance R, and a peaking inductance L, so that the following relations hold:

$$R = X_c = \frac{1}{2\pi f_m C} \qquad \text{and} \qquad L = \frac{X_L}{2\pi f_m} = \frac{\frac{1}{2}X_c}{2\pi f_m}$$

and C is the total shunt capacity.

The factor which limits the gain obtainable from the stage of video amplification is the total shunt capacity of the load circuit. If C_1 and C_2 (see Fig. 16A) can be isolated, the capacitive part of the plate load will be reduced and a larger plate resistor can be used. This will increase the gain obtainable from the stage. In Fig. 18, C_1 and C_2 are separated by the inductance L. The voltage developed across R will be attenuated at the high video frequencies by the shunt capacity C_1. The inductance L and the capacity C_2 form a series resonant circuit across R. The voltage developed across C_2 is applied to the grid of the following tube. If the resonant frequency of the series circuit LC_2 is near the top frequency of the video band, the resonant rise of voltage across C_2 will compensate for the attenuation of the higher frequencies produced by C_1. This method of compensating an amplifier is known as "series peaking".

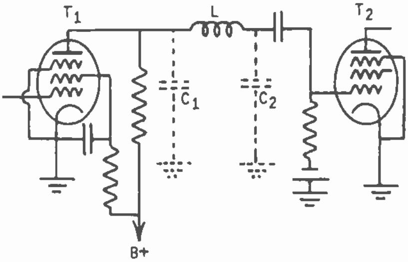


Fig.18 Series peaking circuit when C2/C1 = 2.

This circuit gives the most satisfactory results insofar as gain and a flat frequency response are concerned, when $C_2 \div C_1 = 2$. When this condition exists, the required load resistance R is equal to one-half of the reactance of C_1 at the top video frequency to be passed by the amplifier without loss in gain. Also, the size of L is determined by the condition that its reactance at the top video frequency should be equal to one-half the reactance of C_1.

Putting these statements in the form of equations, we have:

$$C_2 \div C_1 = 2$$

$$R = \tfrac{1}{2}\left[\frac{1}{2\pi f_m C_1}\right] \qquad 2\pi f_m L = \tfrac{1}{2}\left[\frac{1}{2\pi f_m C_1}\right]$$

$$\text{or,} \quad L = \tfrac{1}{2}\left[\frac{1}{(2\pi f_m)^2 C_1}\right]$$

where f_m is the maximum frequency to be passed by the video amplifier.

The condition that $C_2 \div C_1$ must equal two, is not difficult to realize in practice, as the input capacity of a tube is usually higher than the output capacity. Also, the coupling condenser can be placed on either side of the inductance L so that its capacity to ground can be used to obtain the correct ratio between C_1 and C_2. If necessary, additional capacity can be added to either side of L to produce the required ratio. This however, is not advisable, as an increase in capacity results in a reduction in gain.

The advantage of series peaking over shunt peaking can be best shown by means of an example. Let us calculate the size of the load resistor and peaking coil required for both the series and shunt peaking of a video amplifier stage for a maximum video frequency (f_m) of 3,000,000 cycles. For this calculation, we shall take:

$$C_2 \div C_1 = 2 \qquad C_1 = 10\ \mu\mu f. \qquad C_2 = 20\ \mu\mu f.$$

$$\text{and,} \qquad C_1 + C_2 = 30\ \mu\mu f.$$

In shunt peaking (See Fig. 17),

$$R = \frac{1}{2\pi f_m(C_1 + C_2)}$$

and therefore,

$$R = \frac{1}{2\pi \times 3 \times 10^6 \times 30 \times 10^{-12}} = 1770 \text{ ohms.}$$

also,

$$2\pi f_m L = \tfrac{1}{2}\left[\frac{1}{2\pi f_m(C_1 + C_2)}\right] \quad \text{or} \quad L = \tfrac{1}{2}\left[\frac{1}{(2\pi f_m)^2(C_1 + C_2)}\right]$$

and therefore,

$$L = \tfrac{1}{2}\left[\frac{1}{4\pi^2 \times 9 \times 30}\right] = 43 \times 10^{-6} \text{ henries}$$

or, L = 43 microhenries.

In series peaking (See Fig. 18),

$$R = \tfrac{1}{2}\left[\frac{1}{2\pi f_m C_1}\right]$$

and therefore,

$$R = \tfrac{1}{2}\left[\frac{1}{2\pi \times 3 \times 10^6 \times 10 \times 10^{-12}}\right] = 2665 \text{ ohms.}$$

also,

$$2\pi f_m L = \tfrac{1}{2}\left[\frac{1}{2\pi f_m C_1}\right] \quad \text{or} \quad L = \tfrac{1}{2}\left[\frac{1}{(2f_m)^2 C_1}\right]$$

then,

$$L = \tfrac{1}{2}\left[\frac{1}{4\pi^2 \times 9 \times 10^{12} \times 10 \times 10^{-12}}\right] = 129 \times 10^{-6} \text{ henries.}$$

or, L = 129 microhenries.

Since the gain of a video amplifier stage is $G_m R$, the voltage gains obtained by series and shunt peaking will be proportional to the ratio of their load resistances. Therefore, the gain obtained for series peaking will be 2665 ÷ 1770, or 1.5 times the gain obtained with shunt peaking. Thus, by the use of series peaking, the gain per stage can be increased by fifty per cent over that obtainable with shunt peaking.

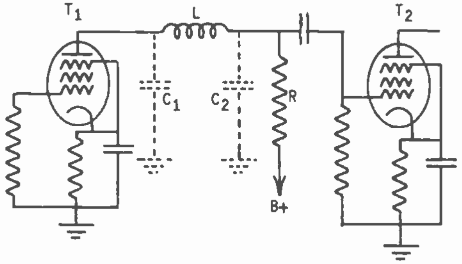


Fig.19 Series peaking circuit when C2/C1 = ½.

Series peaking can be used to advantage when the ratio C_2/C_1 has some other value than two. When the ratio is greater than two, the gain is reduced, but the frequency response remains reasonably flat to the highest desired video frequency. For ratio values between one and two, the frequency response peaks at the high end. A rising characteristic is not usually desirable. However, it can be used to compensate for the high frequency attenuation that occurs for transmission over a coaxial cable, and other causes.

Occasionally it is necessary to couple a tube with high output capacity to one of low input capacity. For this condition, the ratio of C_2/C_1 will be less than one. Maximum gain with a flat response to the highest video frequency is obtained when this ratio is one-half. The load resistor R is placed on the grid side of the peaking inductance instead of the plate side (see Fig. 19). Its magnitude, and the magnitude of L, will be identical to those required when the ratio of C_2/C_1 is two.

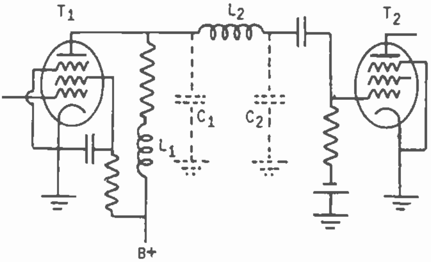


Fig.20 Circuit employing a combination of shunt and series peaking. $C_2/C_1 = 2$.

Series peaking gives 50 per cent more gain than shunt peaking. By using a combination of shunt and series peaking, additional gain can be obtained from a video stage. The circuit is shown in Fig. 20. As in series peaking, the best results are obtained when the ratio C_2/C_1 is two. The magnitudes of R, L_1, and L_2, are given by the following formulas:

$$R = \frac{1.8}{2\pi f_m(C_1 + C_2)}$$

$$L_1 = .12(C_1 + C_2)R^2 \quad \text{and} \quad L_2 = .52(C_1 + C_2)R^2$$

where: f_m is the highest frequency passed by the amplifier without loss in gain.

Let us calculate the magnitudes of R, L_1, and L_2 in the circuit shown in Fig. 20, when $f_m = 3{,}000{,}000$ cycles, $C_1 = 10$ μμf., and $C_2 = 20$ μμf.

$$R = \frac{1.8}{2\pi f_m(C_1 + C_2)} = \frac{1.8}{2\pi \times 3 \times 10^6 \times 30 \times 10^{-12}} = 3200 \text{ ohms.}$$

$$L_1 = .12(C_1 + C_2)R^2 = .12(30 \times 10^{-12})(3200)^2 = 36.8 \times 10^{-6} \text{ henries.}$$
$$= 36.8 \text{ microhenries.}$$

$$L_2 = .52(C_1 + C_2)R^2 = .52(30 \times 10^{-12})(3200)^2 = 159 \times 10^{-6} \text{ henries.}$$
$$= 159 \text{ microhenries.}$$

For the same values of C_1, C_2, and f_m, the maximum load resistance that could be used with shunt peaking was 1770. As the gain of a compensated video stage is proportional to the load resistance, the gain of the combination circuit will be 3200 + 1770, or 1.8 times the gain obtained with shunt peaking.

This circuit can also be used when the ratio of C_2/C_1 is less than two. It is often used in circuits where $C_1 = C_2$, and the load resistor R and L_1 are placed on the grid side of L_2. The gain is less than that obtained when $C_2/C_1 = 2$. Also, different design formulas must be used.

The phase characteristic of these three methods of compensating video stages are quite satisfactory. Of the three, series peaking has the smallest variation in time delay over the frequency band passed with uniform gain. When an amplifier consists of several stages, the total time delay is equal to the sum of the time delays of the individual stages. Sometimes, in order to keep the total variation in time delay of the entire amplifier within limits, it is necessary to modify slightly the values of load resistances and compensating inductances from those given by the formulas. A complete discussion of the methods required in calculating the total time delay in amplifier circuits is beyond the scope of this lesson.

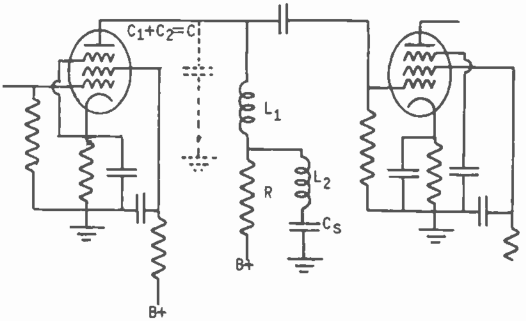


Fig.21 A circuit which gives higher gain without requiring the separation of the total shunt capacity into two capacities with a ratio of 2 to 1.

Fig. 21 shows another method that can be used to obtain high frequency compensation of video amplifiers. This particular circuit will give approximately the same gain as the circuit using a combination of shunt and series peaking. However, it resembles the shunt peaking method in that C_1 and C_2 are not separated. This circuit differs from the others in that the gain is reduced very rapidly as the frequency is increased beyond the highest frequency in the video band. In other words, the amplifier stage has a sharp cutoff. This sharp cutoff characteristic can be used advantageously, as we shall see later in the discussion on head amplifiers.

The design formulas for this particular method of high frequency compensation are given below:

$$R = \frac{1.8}{2\ f_m C}$$

$$L_1 = .8CR^2 \qquad L_2 = .534CR^2$$

$$C_s = .3C$$

where C is the total shunt capacity ($C_1 + C_2$), and f_m is the highest frequency to be passed by the amplifier without loss in gain.

Let us calculate the magnitudes of R, L_1, L_2, and C_s, for a total shunt capacity of 30 mmfds., and a top video frequency of 3,000,000 cycles.

$$R = \frac{1.8}{2\pi f C} = \frac{1.8}{2 \times 3.14 \times 3 \times 10^6 \times 30 \times 10^{-12}} = 3200 \text{ ohms}$$

$$L = .8CR^2 = .8 \times 30 \times 10^{-12} \times (3200)^2 = 246 \times 10^{-6} \text{ henries}$$

$$= 246 \text{ microhenries}$$

$$L = .534CR^2 = .534 \times 30 \times 10^{-12} \times (3200)^2 = 164 \times 10^{-6} \text{ henries}$$

$$= 164 \text{ microhenries}$$

$$C = .3C = .3(30 \times 10^{-12}) = 9 \times 10^{-12} \text{ farads}$$

$$= 9 \text{ microfarads}$$

The value of the load resistance is the same as that for the circuit using a combination of shunt and series peaking. Therefore, the gain for this type of high frequency compensation is 1.8 times that obtained for simple shunt peaking.

5. LOW FREQUENCY COMPENSATION. There are two factors which can cause a loss in gain and incorrect phase relation at the very low frequencies (below 200 cycles). These are: (a) the low frequency attenuation produced by the grid leak and grid condenser combination, and (b) the low frequency degeneration produced by inadequate by-passing of the cathode resistor.

The low frequency phase and frequency distortion caused by the grid leak and grid condenser combination can be minimized by making the time constant of the combination sufficiently large. In order to pass a square wave without excessive distortion, this time constant must be at least ten times the period of the square wave frequency. Since a video amplifier should be able to pass a 60-cycle square wave without distortion, the minimum required time constant is 10 times 1/60 second, or 1/6 second.

Often it is not practical to make the grid condenser and grid leak combination as large or larger than 1/6 second. When fixed bias is used, the maximum size of the grid leak is limited by grid emission and by gas in the tube. The emission of electrons from the grid is usually caused by a very small amount of the cathode emitter deposited on the grid during manufacturing and aging. Since the grid is close to the cathode, its temperature may be raised to the point where electron emission starts. The resultant grid current through the grid leak produces a positive bias on the grid. Often a tube contains sufficient gas so that some ionization takes place. The negative grid collects the positive ions and the resultant grid current through the grid leak produces a positive bias on the tube. Therefore, the size of the grid leak must be sufficiently small so that the positive bias resulting from grid emission or gas current will not upset the operating conditions of the tube.

When cathode bias is used, the effects of grid emission and gas current are reduced. The positive bias due to grid emission and gas current is partially neutralized by the increased cathode bias resulting from the higher plate current. Therefore, a larger grid leak can be used when cathode bias is employed.

The maximum size of the coupling condenser that can be used, is limited by its capacity to ground. Since the maximum gain of the video stage over the band of frequencies present in a television signal varies inversely with the shunt capacity, it is imperative to keep this capacity as low as possible.

Therefore, when fixed bias is used, the time constant of the grid circuit is limited by the maximum size of the grid leak and grid condenser than can be used. However, we learned in a previous lesson that the gain and phase distortion due to an inadequate grid circuit time constant could be corrected by means of a filter (R_FC_F) inserted in the plate circuit as in Fig. 22A. We also learned that, if the time constant C_gR_g was equal to the time constant RC_F, the phase and gain characteristic was satisfactory over the range of low frequencies for which the resistance R_F was ten times the reactance of C_F.

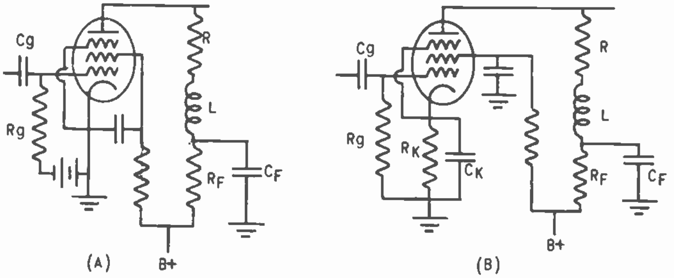


Fig.22 Circuits for low frequency compensation.

When self bias is employed, it is practical to use a grid condenser and grid leak combination that has the required time constant to maintain the desired low frequency phase and gain characteristics. However, if the cathode resistor is inadequately by-passed, low frequency degeneration produces an unsatisfactory gain and phase characteristic. When cathode resistors of two or three thousand ohms can be used, it is economical to use a cathode by-pass condenser large enough to minimize low frequency degeneration. When high transconductance, high plate current tubes are used, the cathode bias resistor is between one and two hundred ohms, and a cathode by-pass condenser of two hundred or more microfarads is needed to prevent low frequency degeneration. This will be true, as it is customary to keep the capacitive reactance of the cathode by-pass condenser at the lowest required frequency below ten per cent of the magnitude of the bias resistor. It is usually uneconomical as far as space and cost are concerned, to use by-pass condensers of this size.

Unsatisfactory low frequency gain and phase characteristics caused by cathode degeneration can also be corrected by use of a filter in the plate circuit (See Fig. 22B). Complete compensation down to zero frequency will exist when the following relations are true:

$$C_K R_K = C_F R_F$$

$$\frac{R_F}{R_K} = G_m R$$

and $$\frac{C_K}{C_F} = G_m R$$

You will note that in this case it is the time constant $C_F R_F$ in the plate circuit that is made equal to the time constant of the circuit ($R_K C_K$) that is causing the unsatisfactory low frequency phase and gain characteristic. If we compare the time constant relations for Fig. 22A and Fig. 22B, we see that in Fig. 22A, correct compensation is attained when the time constant of the series circuit $C_F R$ is made equal to the time constant of the series circuit $C_g R_g$; while in Fig. 22B, correct compensation is attained when the time constant of the parallel circuit $C_F R_F$ is made equal to the time constant of the parallel circuit $C_K R_K$.

For complete compensation of the effects of cathode degeneration, the ratio of the resistance R_f to the cathode resistance R_K is made equal to $G_m R$, which is the gain of the stage at intermediate frequencies. Also, the ratio of C_K to C_F is made equal to the gain of the stage. Therefore, the ratio of the impedance of the parallel circuit $C_F R_F$ to the impedance of the parallel circuit $C_K R_K$ will also be equal to the gain of the stage. At intermediate and high frequencies these two impedances are equal to zero. For low frequencies, the impedance of the circuit $R_K C_K$ is not equal to zero, and the out-of-phase voltage applied to the grid is equal to the product of the AC plate current and the impedance of $C_K R_K$. Therefore, to develop the same output voltage as in the intermediate frequency range, it is necessary to increase the load impedance of the extent that the voltage developed across the additional load impedance ($C_F R_F$) must be equal to $G_m R$ times the out-of-phase voltage developed across the impedance $C_K R_K$. Since the same AC current is flowing through the parallel circuits $C_K R_K$ and $C_F R_F$, the impedance of $C_F R_F$ must be $G_m R$ times as great as the impedance $C_K R_K$.

A loss in gain at low frequencies can also result from inadequate screen by-passing when the screen voltage is supplied through a series dropping resistor. However, since this resistor is usually greater than 50,000 ohms, a by-pass condenser of one or more microfarads is sufficient to prevent the screen voltage from varying with low frequency changes in the screen current.

When electrolytic condensers are used for by-passing in video amplifiers, it is necessary to connect a small paper condenser in parallel with the electrolytic. Electrolytic condensers are not effective as by-passing agents for the intermediate and high video frequencies. At high frequencies, the reactance due to the inductance of the leads and foil, predominates.

The most common methods of compensating video amplifiers in both the extreme high and low frequency ranges have been discussed in the past few paragraphs. In the next few paragraphs, will be a discussion of some of the problems involved in connecting together the units of a television system.

6. TRANSMISSION LINES. All the various parts of a television system, such as the camera and control room, or control room and transmitter, must be coupled together with lines that will pass the wide bands of frequencies involved, without excessive frequency and phase distortion. It would be a waste of time to use wide band amplifiers with little or no frequency and phase distortion over the video band if the lines connecting the various amplifier units had unsatisfactory frequency and phase characteristics.

The type of line used almost exclusively in television systems, is the coaxial line. The coaxial line, as the student knows, consists of one conductor enclosed within another cylindrical conductor. The coaxial line is a self-shielded line.

The characteristic impedance of a coaxial line is determined by the ratio of the inside diameter of the outer conductor to the outside diameter of the inner conductor. Minimum losses occur when this ratio is 3.6 for a given size inner conductor. The characteristic impedance of a line having this ratio is approximately 77 ohms. Therefore, 75 to 80 ohm lines are quite common in television systems.

The characteristic impedance (Z_o) of a coaxial line is found by either of the following equations:

$$(1) \qquad Z_o = 138 \log_{10} \frac{b}{a}$$

where b is the inner diameter of the outer conductor, and a is the outer diameter of the inner conductor; or:

$$(2) \qquad Z_o = \sqrt{L \div C}$$

where L and C are the inductance and capacity per unit length in henries and farads respectively.

The time (T) required for a signal to travel over a coaxial line of length (l) is given by the equations:

$$T = Z_o \times l \times C$$

$$\text{and,} \qquad T = \frac{l \times L}{Z_o}$$

where L and C are the inductance and capacity per unit length.

The phase shift (θ) in degrees between the output and input voltage for a coaxial cable of length (l) for a frequency (f) is:

$$\theta = T \times f \times 360$$

where T is the time of transmission over the cable of length (l).

The attenuation is approximately proportional to the square root of the frequency. The increase in attenuation is due to the increase in the resistance and dielectric losses with frequency.

A transmission line must be terminated by its characteristic impedance in order to prevent reflections from the termination. If the line is properly terminated, all the energy reaching the receiving end is absorbed by the terminating impedance or load. If it is not correctly terminated, part of the energy is reflected back through the line to the sending end or input end. Since it is impractical from the viewpoint of efficiency to make the output impedance of the circuit feeding the line equal to the characteristic impedance of the line, the energy reflected from the receiving end will again be partly reflected and sent back through the line. At the receiving end the signal will again be partly reflected. This will continue until the signal has been completely attenuated by absorption at the sending and receiving ends and by the losses in the line.

If a single pulse is fed into a line that is incorrectly terminated, several pulses with decreasing amplitudes will be received at the other end in succession. The time between two successive pulses will be equal to the time it takes for a pulse to make a round trip through the line. The pulses following the initial pulse are known as echoes. If a television signal is sent through a line that is incorrectly terminated, the reflections from the ends result in multiple images at the receiving end. Each image is attenuated with respect to the preceding image, and is displaced to the right of the preceding image. If the incorrect line termination has a reactive component, there will be a phase shift at the point of reflection and some of the reflected images may arrive at the receiving point completely reversed in phase. Such images would have their black and white sections reversed.

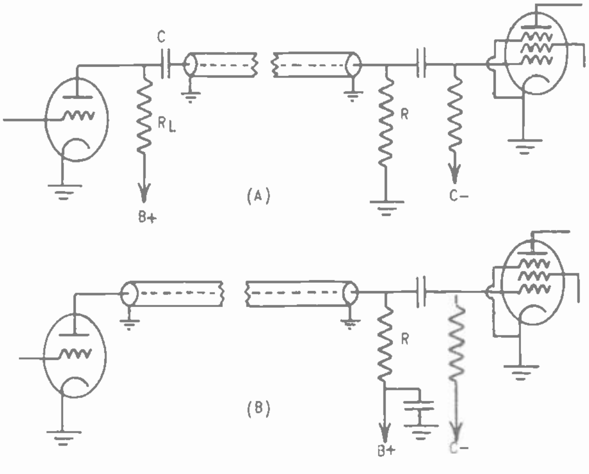


Fig.23 Two methods of feeding a coaxial line from the plate circuit of a vacuum tube.

Coaxial lines can be fed from either the plate or cathode circuits of vacuum tubes. Fig. 23 shows two ways of feeding a coaxial line from the plate circuit. In Fig. 23A, the center wire of the cable is isolated from the DC plate voltage by the blocking condenser C. The resistor R_1 supplies the plate voltage for the tube, and is very large in comparison to the load presented to the tube by the correctly terminated line. The line presents a very low impedance to the tube so that the input to the line will be uniform over the video band. The condenser C must be large so that the product of C and the resistance of the line and its termination will be ten or more times the period of the lowest frequency. If this time constant is less than this, low frequency compensation must be used in the plate circuit of the first tube following the termination of the line. The blocking condenser C and its accompanying low frequency phase and gain distortion can be eliminated by the method used in Fig. 23B. This is impractical if the coaxial line is very long. If the power supply for the amplifier feeding the line is not located near the amplifier, the inductance in the long plate supply lead may upset the termination of the line.

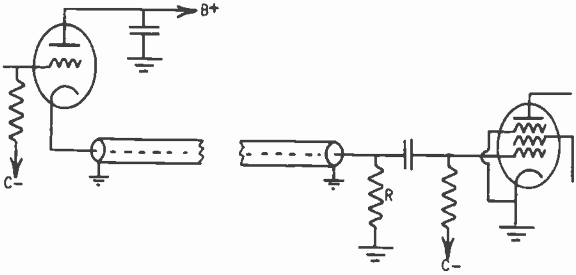


Fig.24 Feeding a coaxial line from the cathode circuit of a vacuum tube.

Fig. 24 shows a method for feeding a line from the cathode circuit. The terminating resistor R carries the DC plate current of the tube. Since the circuit feeding the line is degenerative, the gain of the driving stage will be less than one.

In both circuits, the load presented to the tube feeding the line is very low. In order to obtain sufficient output, it is necessary to use several tubes in parallel. When cathode coupling is employed, the phase of the signal going into the line is the same as that applied to the grid of the driving tube.

Earlier in this section we learned that:

$$Z_o = \sqrt{L \div C} \qquad T = \frac{L \times l}{Z_o}$$

and,

$$\theta = T \times f \times 360$$

where Z_o is the characteristic impedance of the line, L and C are the inductance and capacity per unit length, T is the time of transmission for the length l, and θ is the phase shift between the input and output for a line of length l. If we substitute $(L \times l) \div Z_o$ for T in the expression for phase shift, we have:

$$\theta = \frac{L \times l}{Z_o} \times f \times 360$$

Now if Z_o does not change with frequency, the phase shift of the line is directly proportional to the frequency. This is the condition for no phase distortion. For frequencies above 10 kc. or so, the equation $Z_o = \sqrt{L \div C}$ is a close approximation, and Z_o will be constant for frequencies above 10 kc. However, at very low frequencies, Z_o is determined by the resistance and leakage of the line. This means that at low frequencies, the line will be mis-terminated, and will have phase distortion if the line is terminated by an impedance equal to $\sqrt{L \div C}$. You recall that it is important to terminate a line correctly when its electrical length is greater or of the same order of magnitude as the wavelength of the frequencies applied to the line. Since one wavelength at 60 cycles is equal to 5,000,000 meters, mis-terminations of the relatively short lines used in connecting units in television systems will result in negligible phase distortion and reflections.

The frequency distortion caused by coaxial lines can usually be corrected by using one or two video stages after the line, which have a greater gain at the high frequency end of the video band.

6. THE CONTROL ROOM AMPLIFIER CHAIN. In the first part of this lesson we discussed some of the methods of designing and testing video amplifiers. In the remaining part of the lesson, we shall follow the picture signal from the signal plate of the Iconoscope to the line feeding the video signal to the transmitter.

The control room amplifier chain that will now be described is not the only system that can be used. However, the operations performed on the picture signal are fundamental, and will be the same in all television stations. The circuits may vary, and some of the operations may be combined, but the video signal going to the transmitter will, in all systems, conform to the established standards.

We shall first consider the control room amplifier chain as a whole, then we shall study each component unit. Fig. 25 shows a block diagram of the amplifier chain between the Iconoscope and the line to the transmitter. The waveform of the video output of each unit is shown. It is the waveform generated in scanning the last six lines of the pattern consisting of horizontal white bars on a gray background. Each bar is two lines wide, and the scanning aperture is moving so that the bars are being completely resolved. Perfect vertical resolution has been assumed merely for convenience. Part of the waveform corresponding to the vertical return is also shown. In all these waveforms, white was made positive for simplicity. Actually, the output may be positive or negative, depending on the number of video stages between that point and the Iconoscope.

The first unit in the chain is the head amplifier or pre-amplifier. This may be a three or four stage, AC-coupled video amplifier incorporated in the camera. Its purpose is to build up the video signal so that it can be transmitted over the cable connecting the camera to the control room. The waveform coming out of the head amplifier is the same as that coming from the Iconoscope except that its amplitude is greater. It contains not only the picture signal, but also the spurious signals generated in the Iconoscope. The major part of these spurious signals consist of 13,230-cycle and 60-cycle sawtooth and semi-sine wave components. There is also present in the output of the Iconoscope and head amplifier, the high amplitude transients generated during the vertical and horizontal return lines of the scanning beam.

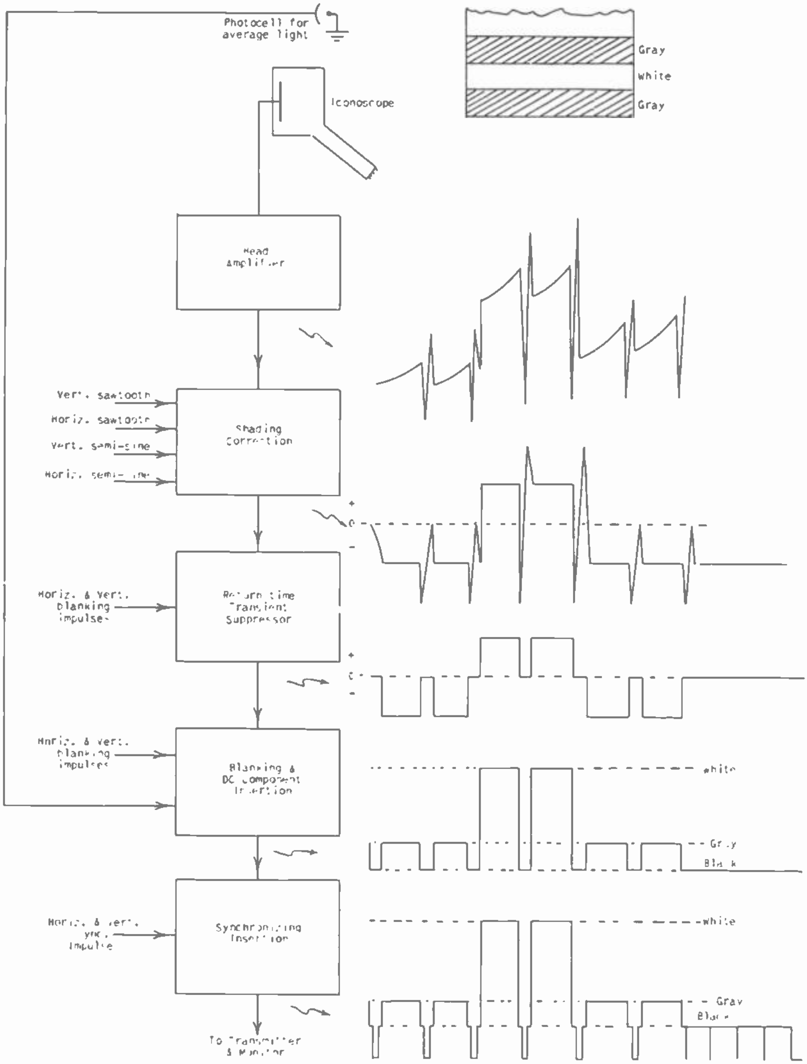


Fig.25 Block diagram of a video control room amplifier chain and the output waveforms for each unit.

The next unit in the chain is the shading amplifier. It is also AC-coupled. Here, sawtooth, semi-sine, and other types of waves (if necessary) are introduced with the proper amplitude and phase to neutralize the spurious signals generated in the Iconoscope. The output of the shading amplifier consists of the picture signal and the return time transients. This waveform is beginning to resemble the waveform we expect from scanning the pattern.

The return time transients are suppressed in the next unit. Sometimes this operation is combined with the insertion of the blanking impulses which occur in the succeeding unit in Fig. 25. The transients are suppressed through the introduction of the blanking impulses twice, but the two blanking impulse inputs are out of phase. This will be described fully later in the lesson. The output of this unit is completely free of the transients. Also, the signal has zero level during the vertical and horizontal return time of the scanning beam in the Iconoscope.

In the following unit, the vertical and horizontal blanking impulses are inserted into the signal. The blanking impulses bias the grid of the picture cathode ray tube to cutoff during the vertical and horizontal return times. The average amplitude of the blanking determines the average light level in the reproduced picture. The amplitude of the blanking is set either manually or automatically, according to the average light level of the picture being televised. In Fig. 25, the output of a photocell placed near the camera determines, automatically, the amplitude of the blanking inserted into the signal. The output of the photocell is a DC voltage which varies slowly with changes in the average amount of light reflected from the televised scene. For a completely black scene, the output of the photocell will be zero, and therefore, the amplitude of the blanking will also be zero. When the output of the photocell, the AC picture signal from the preceding unit, and the blanking, are mixed together, the photocell output serves as the zero axis for the picture signal; and, since the blanking impulse amplitude is proportional to the photocell output, the negative peaks of all the blanking impulses will occur at identically the same voltage level. This level represents absolute black in the picture. Therefore, the voltage difference between any part of the picture signal and the blanking level will be a measure of the absolute brightness of the corresponding part of the televised scene. All this will be evident from a careful examination of the output waveform for this unit. Examining this waveform and, knowing that the televised scene consists of gray and white of equal widths, we can see that the average light intensity is equal to approximately two-thirds of the maximum light intensity for white.

Since the video signal now contains a DC component, the succeeding stages in the amplifier chain must be either DC-coupled or have DC restoration in their grid circuits. (The methods of reinserting the DC component in AC-coupled amplifiers was described in a preceding lesson).

The horizontal and vertical synchronizing impulses (and the equalizing impulses) are inserted into the signal in the final unit of the chain. The output of this unit is a complete television signal. It contains all the components necessary to operate the video section of a television receiver. It is ready for the transmission to the modulators in the television transmitter. The quality of the signal is checked on monitoring cathode ray tube receivers in the control room, and its actual waveform is checked on cathode ray oscilloscopes.

The method of generating the vertical and horizontal blanking and synchronizing impulses will be described in the next lesson. The circuits involved are quite complex, and an entire lesson will be devoted to their study.

We shall now go into a more detailed study of the various units of the control room amplifier chain.

7. HEAD AMPLIFIER. The average signal developed by an Iconoscope across a ten thousand ohm load is approximately one millivolt. It is quite a problem to amplify such a small signal with a satisfactory signal-to-noise ratio. Therefore, the design of the head amplifier is directed toward obtaining the greatest signal-to-noise ratio. The two principal sources of noise in a vacuum tube amplifier are the noise generated by thermal agitation across the grid leak resistances of the first and second tubes, and the noise generated in the first tube by the shot effect. The amplitude of the noise generated is proportional to the square root of the range of frequencies passed by the amplifier.

There isn't much that can be done to reduce the thermal agitation noise generated across the grid leak of the first tube except to reduce the temperature of the resistor to near absolute zero (approximately -460° Fahrenheit, the temperature at which all electron drift and molecular motion ceases). This is impractical to do. The input resistor should be free of noise due to its mechanical construction. Ordinary carbon resistors are unsatisfactory for use as the input resistor because they are made of discrete particles of carbon. Variations in the contact pressure between the particles result in a slight change in the magnitude of the resistance. This causes a noise voltage to be developed across the resistor. The best type of resistor to be used in this position is one where the resistance element is a continuous metallic film, or its equivalent.

The designer can minimize the noise caused by the shot effect in the first tube by the proper selection of tubes and operating conditions. The magnitude of the noise due to shot effect depends on the type of tube. Also, tubes of the same type will vary as to the amount of noise that they generate. To secure a large signal-to-noise ratio for the voltage developed across the tube load resistance, it is essential that the tube have a high transconductance. Triodes generate less shot noise than pentodes. In a pentode, the shot effect will produce slight variations in the screen current. Thus, there are two sources of shot noise in a pentode. Operating a pentode or tetrode as a triode results in a considerable reduction in shot noise. As mentioned in a previous lesson, the shot effect can be reduced by having an adequate space charge present in the tube. Using a relatively low anode or screen voltage helps, as it increases the space charge density. The 1851 and 1852, when operated as triodes, have the highest signal-to-noise ratio of any commercially available type tubes. Even when operated as pentodes, they are superior to most types.

The signal-to-noise ratio can be improved by feeding as large a signal as possible into the amplifier. Since an Iconoscope has a high internal impedance, it should work into a relatively large load resistance in order to obtain a reasonable output. The Iconoscope gives most satisfactory results, as far as output is concerned, with load resistances of 30,000 to 300,000 ohms. However, with conventional coupling circuits, it is impractical to use a large load resistance for the Iconoscope, as the input capacity of the first tube in the video amplifier is in parallel with the load resistance, and the effective load impedance becomes very small at the high video frequencies.

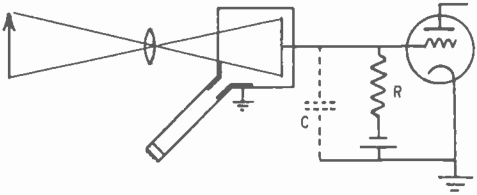


Fig. 26 Circuit for coupling Iconoscope to head amplifier.

Fig. 26 shows a conventional coupling circuit between the signal plate of an Iconoscope and the grid of the first video amplifier. In order to minimize the shunting effect of the input and wiring capacity (C) at high frequencies, the magnitude of R is usually less than 30,000 ohms. Ten thousand ohms is a quite common value.

The amplitude of the noise generated by thermal agitation and shot effect is proportional to the square root of the band width passed by the amplifier. Therefore, the signal-to-noise ratio can be improved if the gain of the head amplifier falls off very rapidly for frequencies above the highest frequency generated in scanning the televised picture.

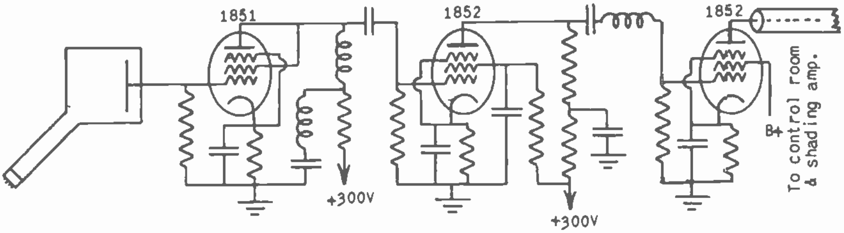


Fig. 27 Circuit of a camera amplifier.

A typical head amplifier circuit is shown in Fig. 27. The load resistance for the Iconoscope has a magnitude of 10,000 to 30,000 ohms. The first tube in the amplifier is an 1851 (high transconductance pentode) operated as a triode to reduce noise generated by the shot effect. In order to obtain a high signal-to-noise ratio from the first tube, the high frequency compensation circuit is of the type shown in Fig. 21. This circuit has a high uniform gain over the required frequency range, and a sharp cutoff beyond the maximum frequency.

The second stage uses an 1851, operated as a pentode. The high frequency compensation netowrk is a series-peaked circuit, in which the ratio of C_2/C_1 has been adjusted to be less than two. For this condition, the gain is greater at the high frequency end of the video band. A rising high frequency characteristic is desirable in this stage to compensate for the attenuation of the high frequency components caused by the shunt capacity across the Iconoscope load resistance. This stage will have less gain than the preceding stage, because of this rising characteristic.

The third tube is also an 1851 operated as a pentode. It feeds a low impedance coaxial cable connecting the camera to the control room and the shading amplifier. The load resistor for the final tube is placed at the end of the line. This is possible, as the power supply for the head amplifier is usually located in the control room.

In this particular circuit, self-bias was used, and the filters in the plate circuits were designed to compensate for low frequency degeneration resulting from inadequate cathode by-passing. If fixed bias had been used, the plate filters would have been designed to compensate for the low frequency phase and gain characteristic due to the grid condenser-grid leak combination.

The heaters of the tubes in the head amplifier are usually operated on DC. This is advisable in order to keep the hum level low in the amplifier chain. However, at Midland Television, the hum level has been satisfactory with AC operation of the heaters.

8. SHADING CORRECTION. The next step in the process of building up a complete television signal is the removal of the spurious signals generated in the Iconoscope. This is accomplished in the shading amplifier which is usually located in the control room.

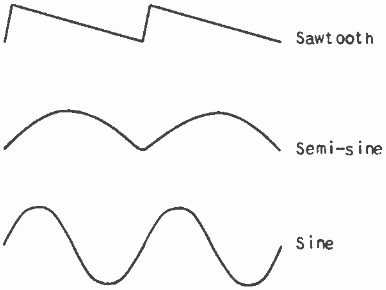


Fig.28 Waveforms used to remove the spurious signal generated in Iconoscope.

The shading unit provides means of introducing six voltage waveforms into the picture signal with the correct amplitude and phase to neutralize the spurious signals or incorrect shading. The waveforms required to produce normal vertical and horizontal shading are 60 and 13,230-cycle sawtooths, semi-sine, and sine waves. These forms are shown in Fig. 28. Normally, satisfactory shading correction can be obtained by the use of just the sawtooth and semi-sine forms. This is true for the output of the head amplifier in Fig. 25. However, under some conditions, a sine wave of variable phase is an aid in obtaining good shading. The 60-cycle sine wave can be used to neutralize hum in the system.

Fig. 29 is a complete diagram of a typical shading correction circuit. T_1 and T_2 are two more stages in the video chain. The input to T_1 is the output of the head amplifier previously described. The terminating resistance for the coaxial line to the camera is a potentiometer which is used as the master picture gain control. The line usually has a characteristic impedance of two or three hundred ohms. Simple shunt peaking is used in the plate circuits of T_1 and T_2. If more gain were required, it would be advisable to use other methods of peaking. If the coaxial line to the camera is long, the network in the plate circuit of T_1 must have a rising high frequency characteristic to compensate for the high frequency attenuation in the coaxial line.

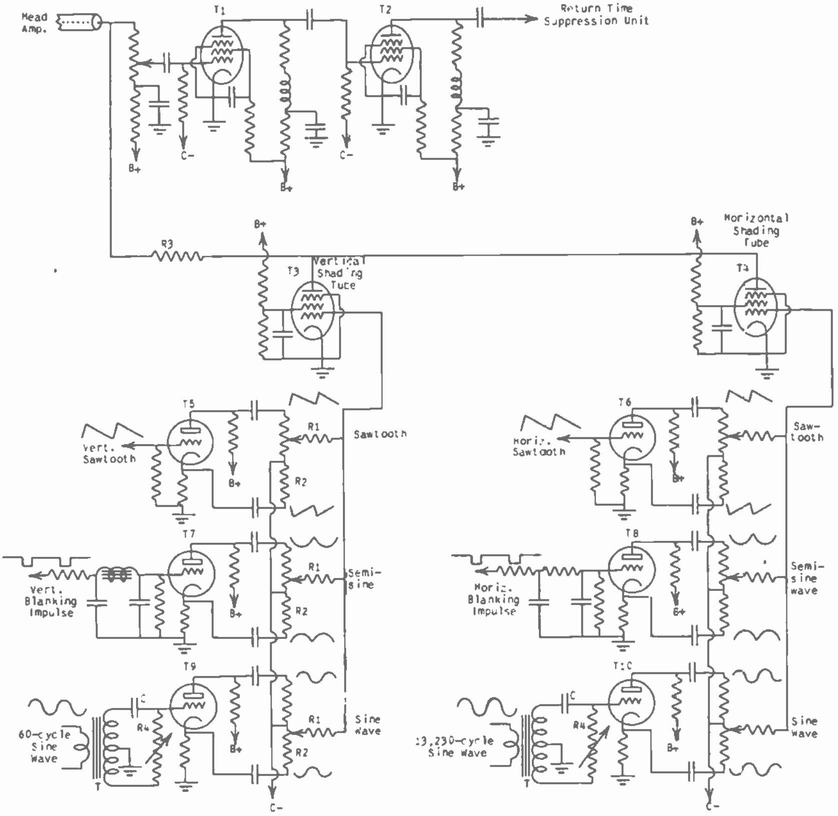


Fig.29 Circuit of a shading amplifier.

Before going farther into the discussion of shading correction it may be advisable to say a little more concerning the method used in Fig. 29 to control the gain. A potentiometer can be used in the grid circuit to control the gain, provided the resistance of the potentiometer is small in comparison to the reactance of the input capacity of the tube at the highest video frequency. If a high resistance potentiometer is used, the tube input capacity, which is in parallel with the part of the potentiometer between the arm and ground, will by-pass the higher frequencies to ground and produce high frequency phase distortion. Other methods used to control the gain in a video chain are to vary the screen voltage on a pentode or to use a variable self bias resistor on a remote cutoff pentode. The bias resistor is not by-passed.

Returning to Fig. 29, we see that the line-terminating resistance also serves as the load resistance for the vertical and horizontal shading tubes T_3 and T_4. The resistance R_3 reduces the effect of loading on the line termination by the output capacities of the shading tubes. The input for the shading tubes is obtained from three sources, each supplying one of the required shading correction waveforms. The tubes T_5 to T_{10} supplying the various waveforms are both cathode and plate loaded. Thus, two outputs having a 180° phase difference can be obtained from each tube. The amplitude and phase is controlled by the center-tapped potentiometers R_2. The series grid resistances R_1 prevent changes in one input from affecting the others. The load resistances for the horizontal tubes T_6, T_8, and T_{10} must be selected so that the tubes will have a flat gain and uniform delay over the band of frequencies included in the horizontal shading voltages. Also, R_1 and R_2 must be small in comparison to the input reactance of T_4 at the top frequency in this same band of frequencies.

The sawtooth waveforms which feed T_5 and T_6 are usually obtained from discharge tubes driven by the Iconoscope vertical and horizontal blocking tube oscillators. The semi-sine wave inputs for T_7 and T_9 can be obtained by passing the vertical and horizontal blanking impulses through a low-pass filter which will filter out the higher frequency components in the blanking impulses. These filters are shown in the grid circuits of T_7 and T_9. The sine wave inputs for T_9 and T_{10} can be obtained from the 60-cycle and 13,230-cycle sources in the synchronizing generator. The network, T, C, and R_4 provides a means of changing the phase of the sine waves over a wide range. Their operation was discussed in a previous lesson under phase control of gaseous triodes.

As stated previously, most of the shading troubles encountered can be corrected by the use of sawtooth and semi-sine waveforms. Also, in general, the phase of the two will be the phase required to correct the shading in the output of the head amplifier shown in Fig. 25. That is, the picture increases in brightness toward the left and bottom sides.

9. RETURN TIME TRANSIENT SUPPRESSION AMPLIFIER. In this lesson, the suppression of the transients generated during the horizontal and vertical return times in the Iconoscope are to be treated separately from the insertion of the blanking impulses. Often, both operations are carried on simultaneously.

The picture signal has a positive polarity at the output of the shading amplifier, that is, white corresponds to a positive voltage change. This is true, as the signal coming from the Iconoscope is negative, and there is an odd number of video stages between the signal plate of the Iconoscope and the output of the shading amplifier.

Fig. 30 is a diagram of a circuit suitable for suppressing the transients generated in the Iconoscope during the vertical and horizontal return times. Tubes T_1, T_5, and T_6 form a part of the control room video chain. The cathodes of T_1, T_2, and T_3 are connected together and have a common cathode resistor R_1. The plates of T_2 and T_5 are connected together and have a common plate load. The potentiometer R_2 controls the amount of bias applied to T_3 and thus it controls the cathode bias applied to T_1, T_2, and T_3. R_3 is a small variable resistor used to vary the gain of T_1 slightly.

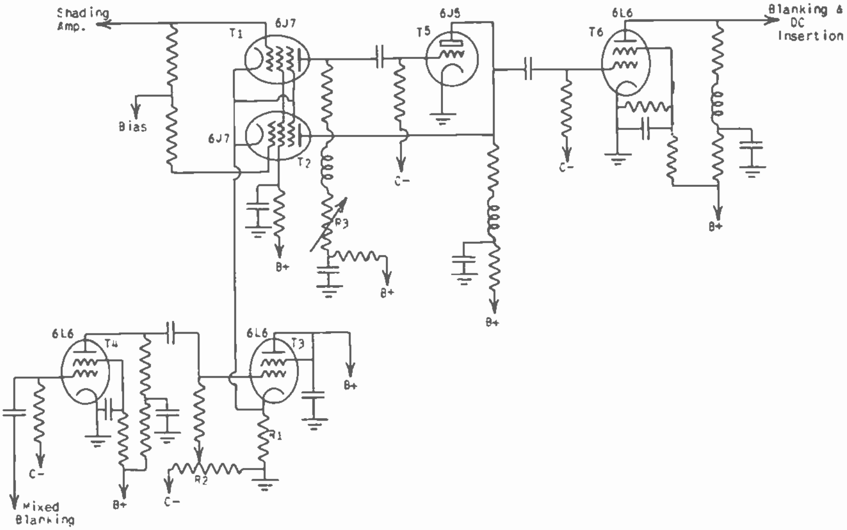


Fig. 30 Circuit for suppressing transients generated during horizontal and vertical return times.

The output of the shading amplifier, which has the waveform shown in Fig. 31A, is applied to the grid of T_1. This consists of the last six lines and a part of the vertical return for the pattern shown at the top of Fig. 25. Mixed blanking impulses from the synchronizing generator are applied to the grid of T_4. The waveform of this signal, which contains both the horizontal and vertical blanking impulses, is shown in Fig. 31B. After amplification by T_4, the blanking signal will appear on the grid of T_3, with a phase difference of exactly 180° to that shown in Fig. 31B. It will also be developed across the cathode resistance R_1 with the same phase. Since this resistance is common to T_1, T_2, and T_3, the blanking signal will also be applied to the grids of T_1 and T_2 with the phase shown in Fig. 31B. Thus, the voltage fed into the grid of T_1 consists of both the picture signal shown in Fig. 31A and the blanking signal shown in Fig. 31B. The combined signal will have the form shown in Fig. 31C. The fixed bias on T_1 is adjusted so that cutoff will appear at the level indicated in Fig. 31C. Therefore, the waveform appearing in the plate circuit of T_1 will contain only the part of waveform C that is above the cutoff bias for T_1. Of course, the phase will be reversed. This waveform, which contains the picture signal and the blanking signal will be free of the horizontal and vertical return time transients, as they are more negative than the cutoff bias for T_1.

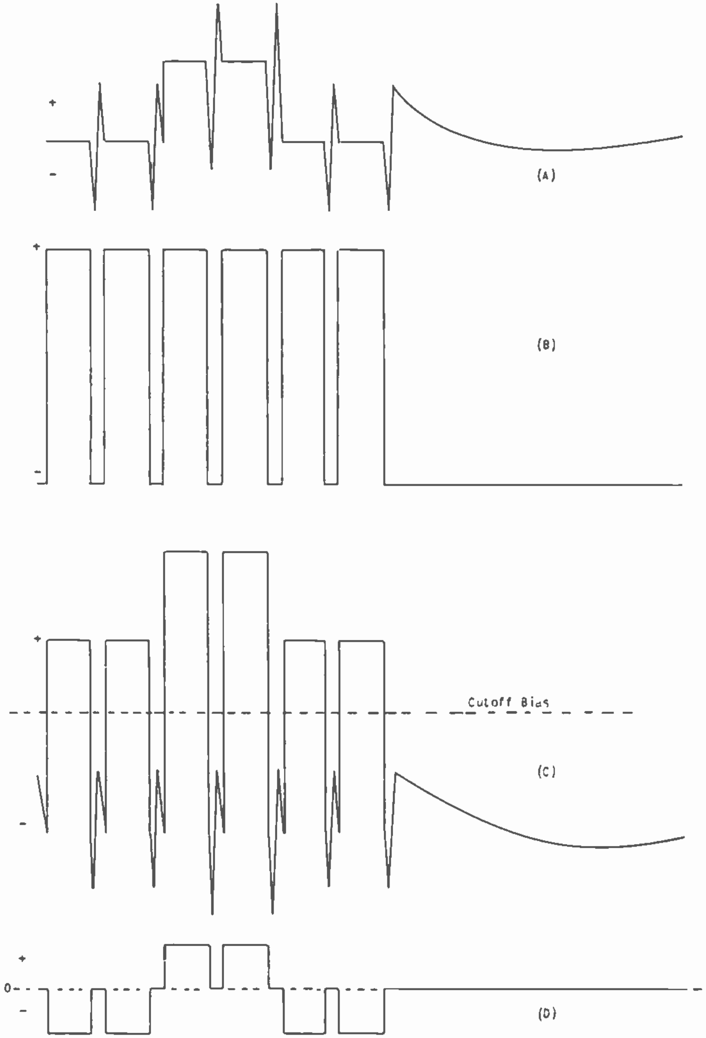


Fig.31 Process of transient suppression.

The signal in the plate circuit of T_1 contains the blanking impulses. In a complete video signal, the average height of the blanking impulses is a measure of the average light level of the televised scene. However, in this case, they were introduced without reference to the average light level of the picture. Therefore, they must be removed again so that they can be re-introduced with the correct level. This is done in the plate circuit of T_5. Tube T_5 reverses the phase of the output of T_1. The blanking was applied to the grids of both T_1 and T_2, and cutoff occurs at the same level insofar as the blanking is concerned. Therefore, the blanking signals developed across the common plate load by T_5 and T_2 will have a phase difference of 180°. Through the correct adjustment of R_3, the amplitude of these two blanking signals can be made equal, and therefore they will cancel. The resultant waveform to the grid of T_6 will have the form shown in Fig. 31D. Note carefully that the signal level is zero during the vertical and horizontal return times.

10. BLANKING AND DC INSERTION. The next step in preparing a complete video signal is to insert the blanking with an average level that is proportional to the average light level of the televised scene.

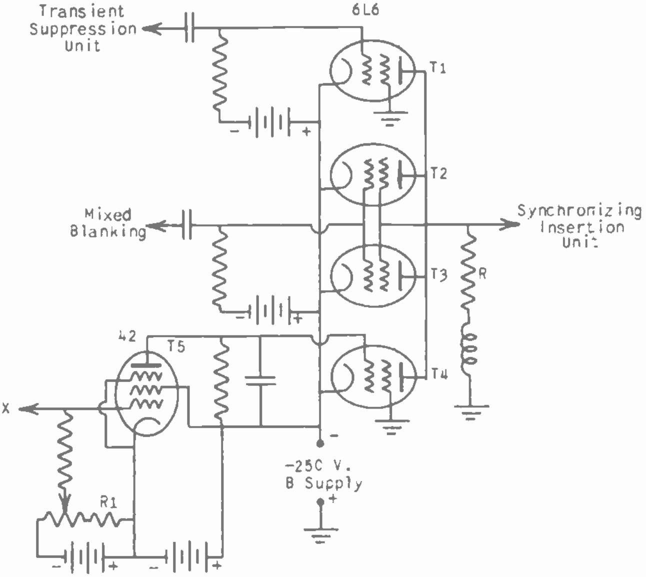


Fig.32 Circuit for inserting blanking and DC component.

Fig. 32 is a diagram of a typical circuit suitable for inserting the blanking with the correct level. Tubes T_1 through T_4 are beam power pentodes. However, T_2 and T_3 are operated as triodes; that is, their plates and screens are connected together. All the tubes are in parallel insofar as their plate and cathode circuits are concerned. The plus side of the plate supply for these four tubes is tied to ground. Therefore, the cathodes must be operated below ground, or negative with respect to ground. This is done so that this unit and the following unit can be direct-coupled conveniently. Tube T_5 is direct-coupled to the grid of T_4. The bias on T_5, and therefore, the bias on T_4 can be controlled by the potentiometer R_1 across the bias battery. This potentiometer is known as the manual background control.

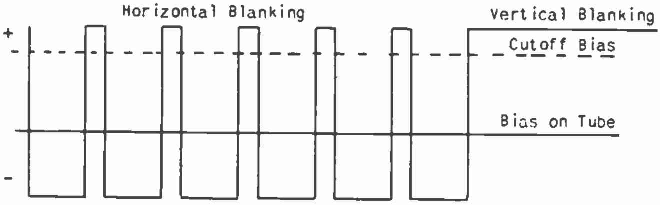


Fig.33 Waveform on grids of T2 and T3 in Fig. 32.

The output from the transient suppression unit (Fig. 31D, with the phase reversed) is applied to the grid of T_1. Mixed horizontal and vertical blanking, with the phase shown in Fig. 33, is applied to the grids of T_2 and T_3 in parallel. Two tubes are required to develop sufficient blanking output. When automatic background control is used, the amplified output of the photocell is applied to the grid of T_5. Of course, the photocell amplifier must be DC-coupled. The output of the amplifier must change the grid of T_5 positive for white. The necessity for DC-coupling is one reason why the use of automatic background control is not common.

Normal bias is used on T_1. T_2 and T_3 are biased to considerably beyond cutoff so that only the peaks of the blanking impulses will swing the grids sufficiently positive for plate current to flow. Since T_2 and T_3 are triodes, the flow of plate current will also be controlled by the plate voltage. In other words, the cutoff bias is affected by the plate voltage. This means that the amplitude of the blanking developed across the plate load R will depend on the plate voltage. The actual plate voltage being applied to T_2 and T_3 is below ground by the drop across the resistance R. Therefore, the plate voltage will decrease with an increase in current through R. If we examine the waveform of the blanking applied to the grids of T_2 and T_3 (Fig. 33) and the waveform of the picture swing developed across R (Fig. 34A) we see that T_2 and T_3 draw plate current only during the vertical and horizontal return times when the amplitude of the picture signal is zero. Therefore, the plate voltage of T_2 and T_3 is independent of the picture signal during the interval when they can conduct. However, their plate voltage will be affected during the conduction period by the current taken by T_4. Therefore, the amplitude of the blanking developed across the plate load R is controlled by the plate current of T_4. When the bias on T_5 is made very negative, the high plate current of T_4 will lower the plate voltage on T_2 and T_3 to the extent that only a small part of the peaks of the blanking will be developed across R. When the bias on T_5 is shifted in the positive direction, the plate current of T_4 is reduced, and the amplitude of the blanking developed across R is greater because of the higher plate voltage on T_2 and T_3.

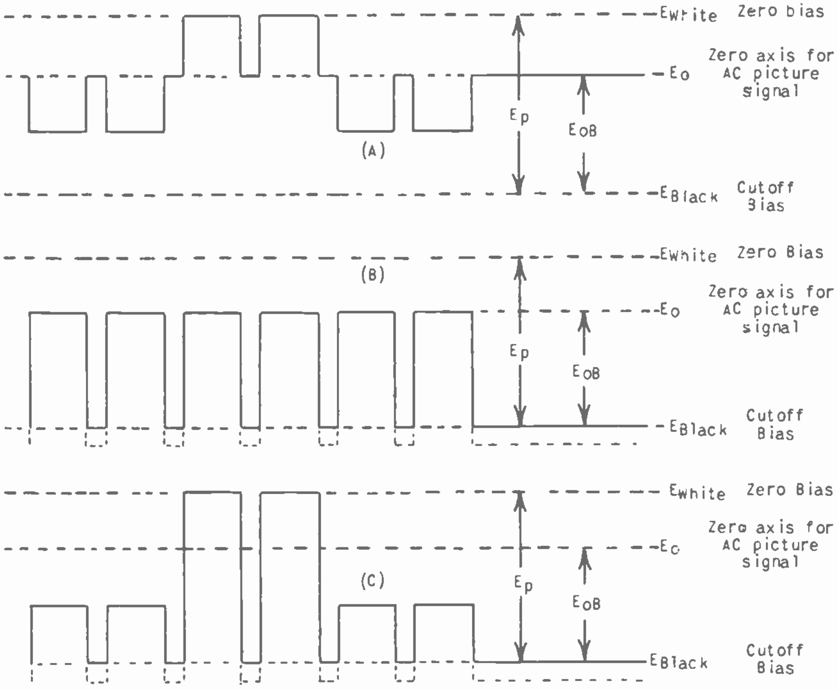


Fig.34 Illustrating insertion of blanking and DC component.

Changing the bias on T_5 raises or lowers the plate voltage on T_1, and thus raises or lowers the absolute value of the zero axis of the AC picture signal developed across R. Raising and lowering the axis of the picture signal is the same as inserting a DC component into the picture signal. If the axis is raised and lowered in accordance to the average light level of the televised picture, this DC component will be proportional to the average light level in the scene. Also, the amplitude of the blanking developed across R will be proportional to the average light level. Since T_2 and T_3 are at cutoff except during the vertical and horizontal return times, the positive peaks of the blanking developed across R will coincide with the zero axis for the AC picture signal. In other words, the blanking impulses extend from the zero axis downward (See Fig. 34B).

Although the signal developed across R contains information concerning the average light level of the picture, no absolute voltage levels corresponding to black or white have been established.

Since black is the absence of light, it is a definite energy level. The total brightness of a scene is not absolute, and it is always possible to increase the intensity of the illumination. Thus, we can establish a definite voltage level in a television system for black, while the maximum level corresponding to the brightest part of the scene will be controlled by the peak-to-peak amplitude of the signal that the system can handle without distortion.

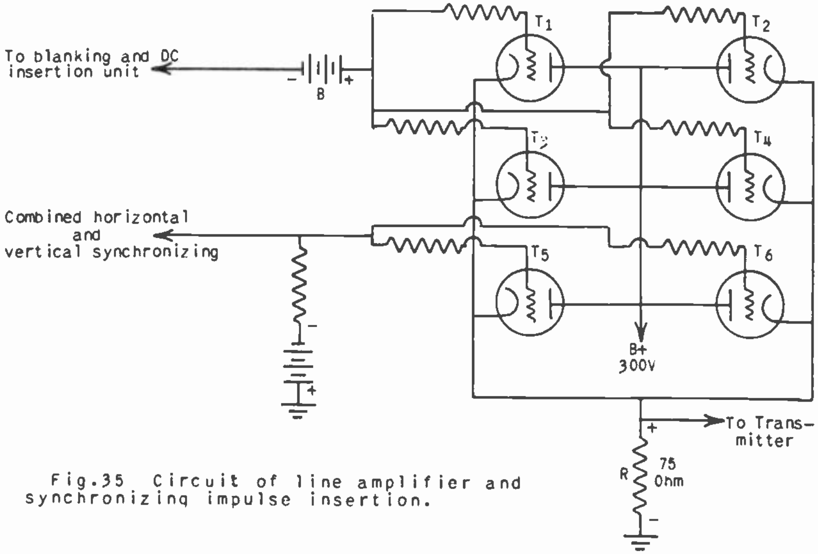


Fig.35 Circuit of line amplifier and synchronizing impulse insertion.

The absolute black level is established in the grid circuit of the final unit of the chain. The output of the blanking and DC insertion unit is direct-coupled to the unit shown in Fig. 35. Direct coupling is necessary, as the signal now contains a DC component. The synchronizing is also inserted into the composite video signal in this circuit. A discussion of synchronizing insertion will be given a little later in the lesson.

Fig. 35 consists of six tubes. All the tubes use a common 75 ohm cathode load. All the plates are operated at ground potential as far as the video signal is concerned. The grids of the upper four tubes are connected together and the grids of the lower two are also connected together. Small isolating resistors of one or two hundred ohms are inserted in each grid lead to prevent self oscillation. The output of the blanking and DC insertion unit is applied to the grids of the upper four tubes and the synchronizing is applied to the grids of the lower two tubes.

The absolute black level is established by driving the grids of the upper four tubes or the picture tubes to cutoff by the blanking. Fig. 34 shows the waveform of the signal, and its components, that is applied to the grids of the upper four tubes in Fig. 35. Fig. 35A shows the AC picture signal varying about its zero axis. Fig.

35B shows the blanking. The positive peaks of the blanking have the same level as zero axis for the AC picture signal. Fig. 35C shows the combined signal containing the blanking and the AC picture signal.

The amplitude of the blanking and the level of the zero axis is controlled by the bias on T5 in Fig. 32. The maximum peak-to-peak voltage swing on the grids of the picture tubes will be from cutoff to zero bias. Therefore, the maximum peak-to-peak value of the AC picture signal, which is determined by black and the brightest part of the picture, must be limited to this range. The voltage developed across the cathode load in Fig. 35 will range from zero at cutoff to the maximum positive value obtained with zero bias as far as the picture tubes are concerned. Due to the degeneration, this peak-to-peak voltage change will be less than the allowable grid swing. Since the voltage developed across the cathode load cannot be less than zero as far as the upper four tubes are concerned, it makes an excellent absolute value for black. Therefore, in order to have zero output always occur for black in the picture, it is necessary to shift the zero axis of the picture signal so that the negative peaks occurring for black will just drive the grids of the picture tubes to cutoff. The amplitude of the blanking, which increases and decreases as the zero axis is raised and lowered, is always sufficient so that the negative peaks of the blanking are greater than the cutoff bias on the picture input tubes in Fig. 35. Therefore, the negative peaks of the blanking developed across the cathode load will have the same absolute value as black.

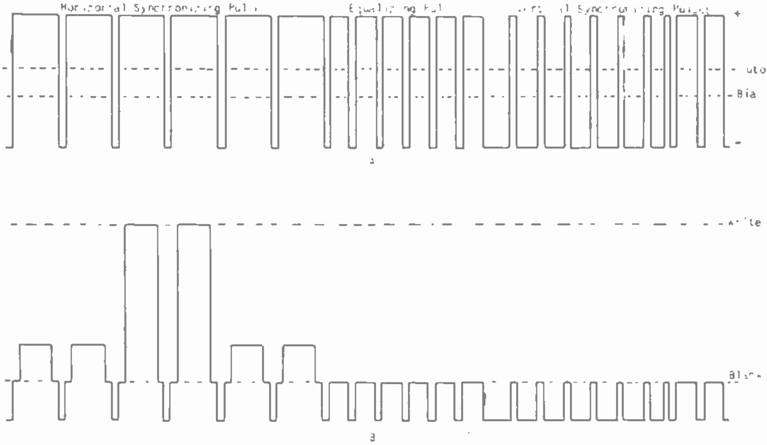


Fig.36 Waveforms showing process of inserting synchronizing impulses.

When a solid shade is being televised, such as black or white, there will be no AC picture signal applied to the grids of the picture input tubes in Fig. 35. The voltage applied to these grids will be DC, and is determined by the bias on T5 and the plate current of T1 in Fig. 32, and by the bias battery B, in Fig. 35. The battery bias for the picture input in Fig. 35 is adjusted so that the DC applied to the picture input grids can be varied from cutoff to zero by varying the bias on T5 in Fig. 32 over its range. When televising solid black, the bias on T5 in Fig. 32 is set at maximum, and the DC applied to the grids of the picture input in Fig. 35 is the cutoff value. When televising white, the bias on T5 in Fig. 32 is set at minimum, and the DC applied to the grids of the picture input in Fig. 35 is zero. When televising intermediate shades, the bias on T5 is adjusted so that the DC applied to the grids of the picture input will correspond to the shade. When televising ordinary scenes, the bias on T5 is adjusted so that the DC applied to the grids of the picture input is proportional to the average light level or shade of the scene. This DC level is also the zero axis for the AC picture signal. A similar set of conditions exists for

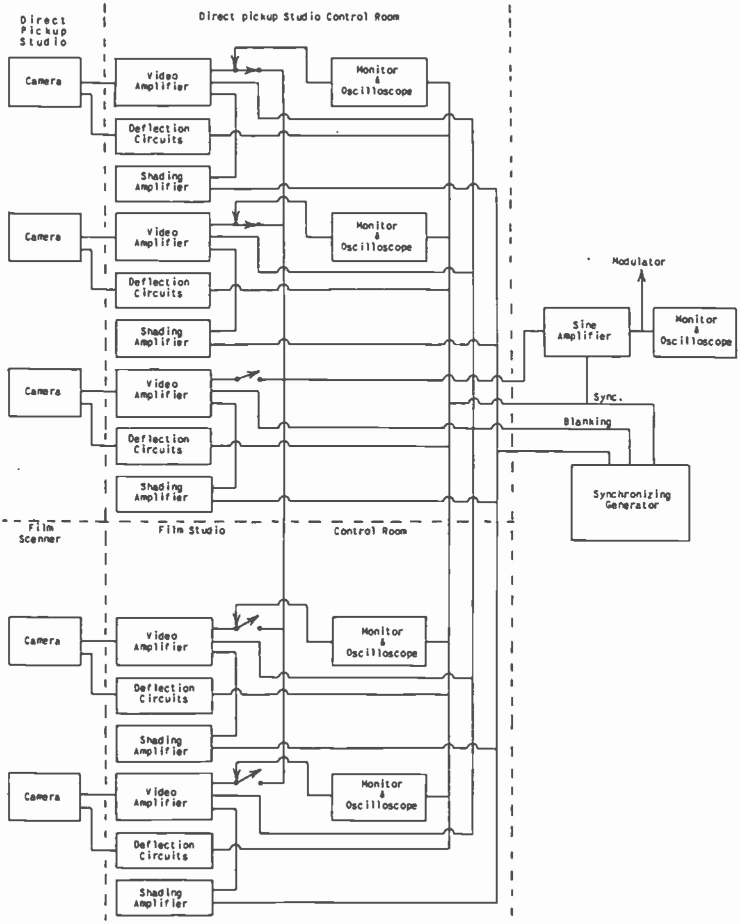


Fig.37 Block diagram of a television system with five cameras.

the signal developed across the cathode load for the picture tubes in Fig. 35. The DC component will vary from zero to the maximum positive value occurring for zero bias on the picture input grids as far as the upper four tubes are concerned. It also will be proportional to the average shade or average light level of the picture.

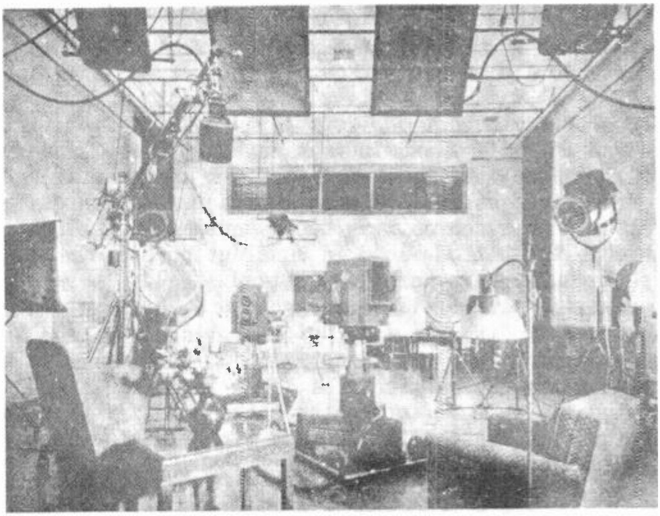


Fig.3B NBC television studio.

The waveform used for explaining the operation of the various units in a television control room chain resulted from scanning white bars on a gray background. The bars were equal in width. If we assume the gray is a shade one-third of the way up on the scale from black to white, the average light level or background level will be a shade of gray two-thirds of the way up on the scale from black to white. Then, in order to reproduce this pattern with its correct shading and background level, the control room operator will adjust the bias on T_5 in Fig. 32 so that the zero axis for the picture signal or the DC component applied to the grids of the picture tubes in Fig. 35 will be one-third of the cutoff bias. In Fig. 34B and 34C, E_{Black} is the cutoff bias for the picture tubes in Fig. 35, while E_{White} is zero bias. The bias on T_5 in Fig. 32 is adjusted so that the zero axis of the picture signal is two-thirds of the way between E_{Black} and E_{White}, or the distance E_{OB} in Fig. 34B. Fig. 34A shows the AC picture signal alone. The voltage is changing above and below the zero axis so that the areas above and below the zero axis are equal. Fig. 34B shows the blanking alone. The blanking swings the grid voltage on the picture input tubes in Fig. 35 from E_0 to beyond cutoff. Fig. 34C shows the AC picture signal and the blanking combined. In this waveform, instantaneous voltages correspond to a definite shade between black and white.

11. SYNCHRONIZING IMPULSE INSERTION. The next and final step in building up the complete video signal is to insert the horizontal

and vertical synchronizing impulses. This operation is carried out in tubes T_5 and T_6 in Fig. 35. The mixed horizontal and vertical synchronizing impulses with the phase shown in Fig. 36A is applied to the grids of T_5 and T_6. The amplitude of the mixed synchronizing is sufficient so that the negative peaks drive the grids of T_5 and T_6 beyond cutoff.

The 75 ohm cathode load resistance in Fig. 35 carries both the plate current for the picture tubes and the synchronizing tubes. The picture tubes are biased to cutoff for black. Therefore, the actual black level will be positive with respect to ground by the voltage drop across the cathode resistor, caused by the plate current of T_5 and T_6. If we examine the waveform of the synchronizing in Fig. 36A, we see that the amplitude of the positive peaks is constant during the interval between the actual synchronizing impulses. Since the synchronizing occurs during the blanking, the current through the cathode load is constant during the time that the picture signal appears across the cathode load. Therefore, this constant positive voltage will not upset the light levels that have been established by the previous unit. During the time that the actual synchronizing impulses are applied to the grids of T_5 and T_6, these two tubes are biased to cutoff. Therefore, the synchronizing impulses swing the voltage across the cathode to zero for the negative peaks of the synchronizing. The synchronizing impulses add to the blanking, as shown in Fig. 36B.

Fig.39 NBC television control room.

In the usual video signal, the synchronizing takes twenty per cent of the peak-to-peak amplitude of the signal. In order to obtain this ratio, it is necessary to adjust the bias on T_5 and T_6 and the amplitude of the synchronizing so that the constant positive voltage developed across the cathode load is one-fourth of the peak-

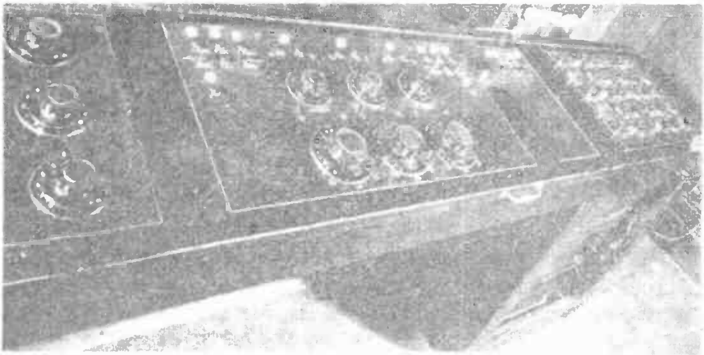


Fig.40 Video control console.

to-peak voltage developed across the same load by the picture tubes.

The signal developed across R is a complete video signal. It is ready for transmission over a coaxial line to the transmitter and to the control room monitors. A regular cathode ray tube receiver is used in the control room to check the quality of the picture signal going out on the coaxial line. A cathode ray oscilloscope is used to check the actual form of the video signal and to set the correct percentage of synchronizing to picture. It is also used to see that the proper amount of DC has been inserted into the signal.

Fig.41 NBC control room.

12. POWER SUPPLY REQUIREMENTS. The plate power supplies used in the television control room chain must have a much lower ripple output and far better voltage regulation than can be supplied by conventional full-wave rectifier and low-pass filter units. In many installations, the plate power is obtained from storage batteries. This is both inconvenient and expensive. AC operated power supplies with the regulation and ripple output required in television installations will be described in a following lesson. These power supplies have automatic vacuum tube regulators.

13. A MODERN TELEVISION SYSTEM. The simpler modern television installations are designed for the operation of two to three cameras. One or two cameras are used in the studio for televising live talent. The other is used in conjunction with a motion picture projector for televising film. At least two cameras are required in the studio, one for televising the complete set, and the other for close-ups. Also, if the program involves the use of two or more sets, a much smoother transition from one set to the other can be made when two cameras are available. In more elaborate installations there will be more than one studio, and each studio will be equipped with at least two cameras, and probably three. Also, two projectors are usually used for televising film.

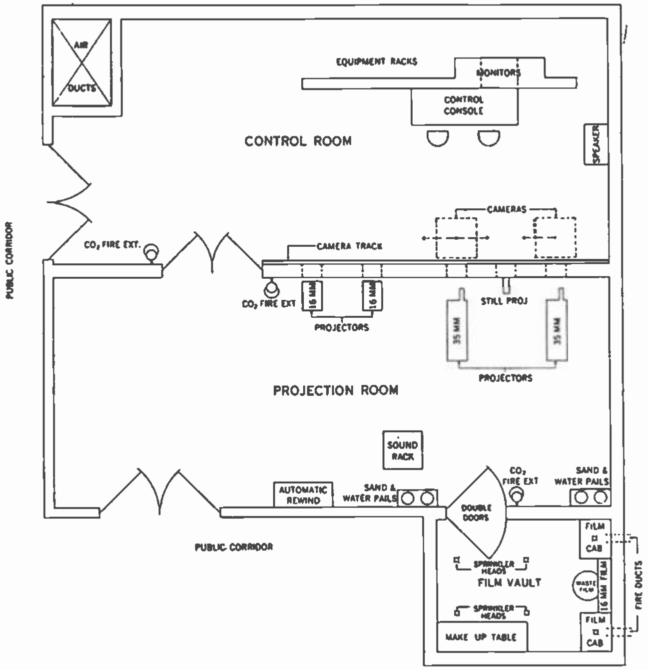


Fig.42 Block diagram of NBC film studio.

Each camera has a separate amplifier chain for the shading correction, suppression of return time transients, and for the insertion of the blanking and the DC component. This is necessary, as the outputs of the various cameras will be different as far as shading correction is concerned. Also, the background level of the televised scene will vary for each camera. The background level for a close-up will differ from that for the entire set. There will also be a separate set of deflecting circuits, keystone correction circuits, and high voltage supply for each Iconoscope. Of course, each camera will have its own head amplifier. The only unit in the amplifier chain that all the cameras use in common is the line amplifier where the vertical and horizontal synchronizing impulses are inserted into the video signal.

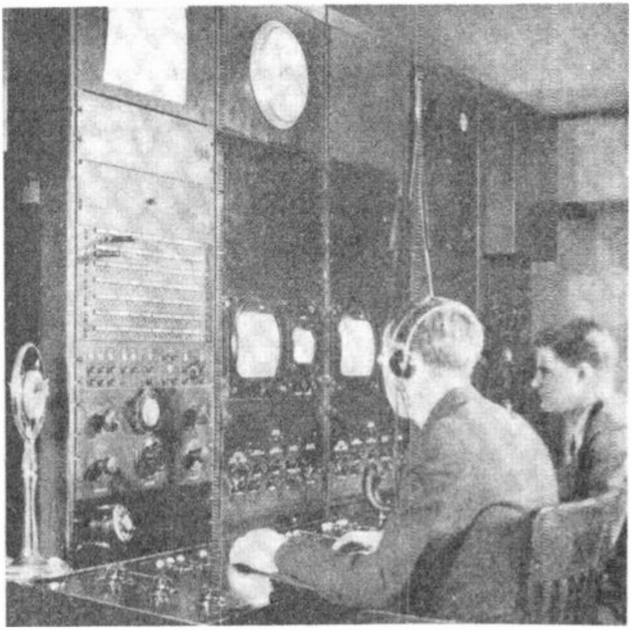


Fig.43 NBC control console for film cameras.

Fig. 37 is a block diagram for a five camera system. Three cameras are for direct pickup and the other two are for televising film. In order to simplify the diagram, the shading circuits, return time transient suppression, and the blanking and DC insertion are included in a single unit. This is labeled "video amplifier" in the diagram. Two monitors are provided for observing the outputs of the three studio cameras. Either monitor can be connected to any of the studio cameras. Similarly, there are two monitors to check the output of the film cameras. Since these monitors are connected in the amplifier chain before the insertion of synchronizing, it is necessary to synchronize their deflecting circuits by the vertical and horizontal synchronizing impulses taken directly from the synchronizing generator. Each monitor is equipped with an oscilloscope for monitoring the amplitude of the signal, and for adjusting the background level. There is also a monitor at the output of the line amplifier or synchronizing insertion unit for an overall check of the picture. There is also an oscilloscope for observing the level of the synchronizing and for a final check of the video signal.

In small television installations, the amplifier chains for each camera, the line amplifier, and the synchronizing generator are located in a single control room next to the studio. Usually, a separate studio is provided for the film cameras. In larger installations, where there are several studios, the camera amplifier chains for the cameras in each studio are located in the control room for that studio. The line amplifier and synchronizing generator are located in a room centrally located with respect to all the studios.

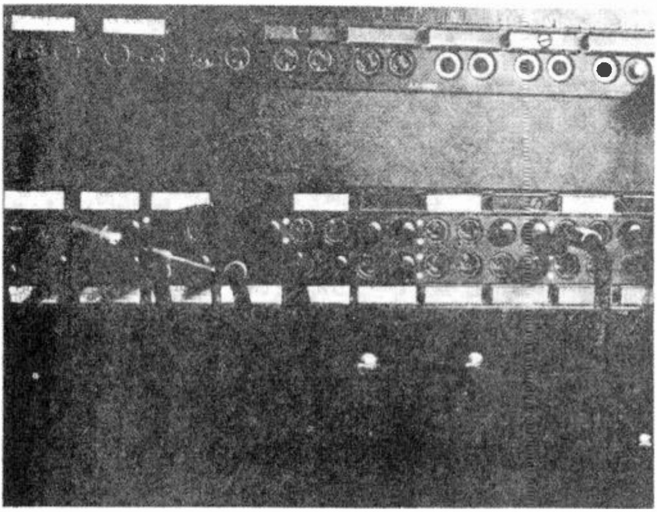


Fig.44 Coaxial cable jack panel.

Fig. 38 is a picture of an NBC television studio. The windows in the rear wall are to the studio control room. Fig. 39 shows the interior of the control room looking toward the windows to the same studio. The cathode ray tube picture monitors and oscilloscopes are located above the window. The six controls directly in front of the operator are the contrast and brightness controls for the three studio cameras. Directly above these knobs are switches for operating relays for switching the cameras and picture monitors. The panel to the right of the operator contains the shading controls. The three large knobs to the left of the operator control the focus for the three Iconoscopes. Fig. 40 shows a better view of the video control desk used at NBC. Fig. 41 is a more complete view of an NBC television control room. The man in the foreground is the audio engineer. The desk contains the fading and gain controls for monitoring the sound part of the broadcast. The next man is the production man. The third individual is the video engineer. Part of the control room amplifiers can be seen back of the men. The video controls are controlled remotely from the video engineer's desk.

Fig. 42 shows the floor plan of an NBC film studio. There are four projectors and two cameras. The cameras can be moved along a track so that any projector may be used. A complete discussion of televising film will be given in a later lesson. Fig. 43 is a picture of control equipment for the film studio. Controls are provided here for switching both the audio and video circuits between studios.

All the video units in a television installation are connected together by low impedance concentric lines. Of course, all connecting lines must be properly terminated, in order to prevent reflections. Switching of video circuits is done by specially designed relays. These relays must have very low capacity so that the high frequencies will not be excessively attenuated, and so there will be very little capacity connected across the terminating resistances for the concentric lines. The terminations should be pure resistances. The input and output of the relays are electrostatically shielded to prevent the higher frequencies from feeding through even when the relays are open. Also some switching is done by the use of coaxial patch cords. Fig. 44 shows a coaxial cable jack panel and patch cords.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. Which type of amplifier distortion must be held to a minimum in a video amplifier? Why?

2. What is a quick and simple way of testing the transmission characteristics of a video amplifier?

3. A video amplifier has lower gain and less delay at the low frequency end of the video band than it has at the intermediate frequencies. Diagram the output waveform if the input is a 60-cycle square wave.

4. How will the type of distortion described in the previous question show up in the reproduced picture?

5. Illustrate by diagrams, three methods of compensating video amplifiers in order to obtain satisfactory high frequency response.

6. Why is it necessary that a coaxial transmission line be terminated in its characteristic impedance?

7. Draw a block diagram of control room amplifier chain, and state briefly the function of each part.

8. How are the spurious signals removed from the output of the Iconoscope?

9. What information is conveyed by the DC component in a video signal?

10. How much of the control room amplifier chain diagrammed in question 7 must be duplicated when more than one camera is used?

APPENDIX

(VECTORS)

Vectors provide a very simple method for studying the amplitudes and phases of voltages, currents, and impedances in AC circuits. Fig. 1 shows several vectors. Vectors have two important properties: length and direction. In alternating current electricity, we can represent the magnitude of a voltage, current, or impedance, by a vector of proportionate length. We can represent the phase of the voltage, current, or impedance, by the direction we give the vector.

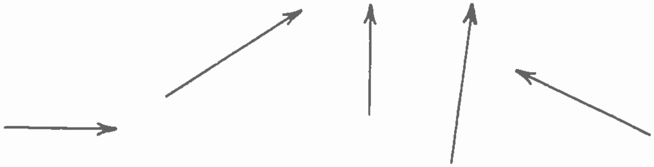
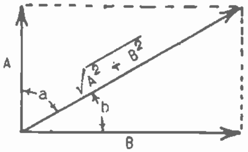


Fig. 1 Vectors.

If we wish to add two voltages together by vectors that have a phase difference of 90 degrees, we draw two vectors perpendicular to each other, as A and B in Fig. 2. The lengths of the vectors are drawn proportional to the magnitudes of the voltages. The next step is to complete the rectangle formed by the two perpendicular vectors. The diagonal of this rectangle drawn from the intersection of the two vectors is a vector which has the same magnitude as the sum of the two voltages A and B, This is true, as we know that the hypotenuse of a right triangle is equal to the square root of the sum of the squares of the sides. The angle between the diagonal vector and either of the sides is equal to the phase angle between the total voltage and the voltages A and B respectively. In using a vector diagram, we measure the leading angles in the counter-clockwise direction, and the lagging angles in the clockwise direction. Therefore, the total voltage is leading the B component by b degrees, and is lagging the A component by a degrees.

Fig.2 Adding vectors.



Now let us find, by means of vectors, the potential drop across a series circuit containing an inductance and resistance. We know that the potential across the resistance will be IR, and the potential across the inductance will be $2\pi fLI$. We also know that the voltage and current through a resistance are in phase, and that the voltage across an inductance leads the current by 90 degrees. Then let us draw a vector IR horizontally (Fig. 3). Since the current through the inductance and resistance are in phase, the voltage across the inductance will lead the voltage across the resistance by 90 degrees. Therefore, the vector $2\pi fLI$ is drawn perpendicularly to the vector IR. It will point upward because the voltage across the inductance leads the voltage across the resistance by 90 degrees.

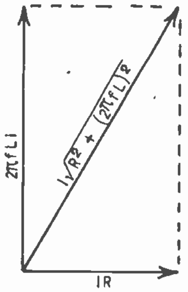


Fig.3 Vector diagram for inductance and resistance in series.

If we complete the rectangle and draw in the diagonal, the length of the diagonal will be equal to the voltage drop across the series circuit, if the other two vectors have been drawn to scale. The number of degrees in the angle between the diagonal and the vector IR is the number of degrees that the voltage across the series circuit leads the current through the circuit.

The magnitude of diagonal from the geometry of the figure is seen to be:

$$I\sqrt{R^2 + (2\pi fL)^2}$$

Since the current occurs in the expression for the magnitude of each voltage, the magnitudes of all the vectors can be divided by I, and their relative magnitudes are unchanged.

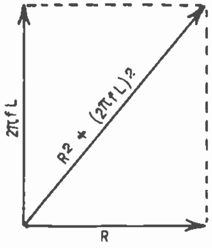


Fig.4 Vector diagram of impedances.

The vector diagram will now consist of impedances instead of voltages (Fig. 4). The horizontal vector is the resistive component, and the vertical is the inductive component. The diagonal is the total impedance of the combination. Thus, vectors can be used for adding impedances together.

Now let us consider a series circuit consisting of a condenser and resistance (See Fig. 5). The potential drop across the resistance is IR, and the potential drop across the condenser is $I \div 2\pi fC$. Since the voltage across a condenser lags the current through the condenser by 90 degrees, the vector representing the voltage across the condenser is drawn pointing downward. The diagonal of the rectangle using IR and $I \div 2\pi fC$ as sides will represent the magnitude and phase of the voltage across the combination. In this case, the total voltage lags the current in the circuit. The amount of the lag in degrees, is equal to the size of the angle, in degrees, between the diagonal and the vector IR. As in the case of the previous example, the total impedance can be found vectorially.

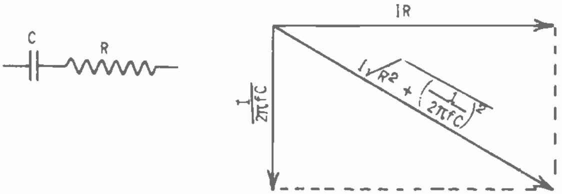


Fig.5 Vector diagram for resistance and capacity in series.

Now let us consider a series circuit containing resistance, inductance, and capacity. The vector voltage diagram will have the form shown in Fig. 6B. The vectors representing the voltages across the inductance and capacity, point in opposite directions, as the voltages across the two are 180 degrees different in phase. Therefore, the sum of the two will be the difference in length between their respective vectors. The direction or phase of the sum will be the direction and phase of the larger vector. After adding, the voltages across the reactive part of the vector diagram take the form shown in Fig. 6C. In this case, the voltage across the inductance is greater, and the vector representing the sum of the two,

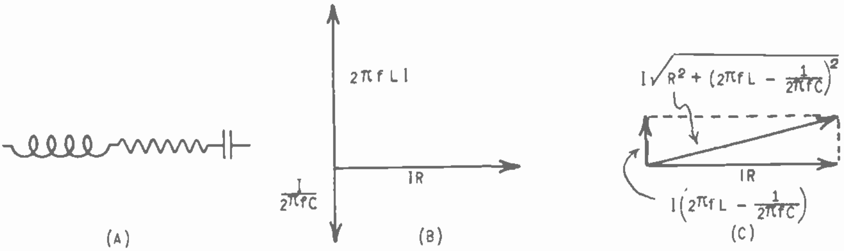


Fig.6 Vector diagram for resistance, inductance, and capacity in series.

points upward. The total voltage drop is found vectorially, as before. The voltage across the combination leads the current through the combination by the angle between the diagonal and the vector IR. By removing the factor I from each vector, the vector diagram in Fig. 6 can be used for finding the total impedance of the circuit. For the resonant condition, the vectors representing the inductive and capacitive reactances will be equal and opposite, and will cancel. The total impedance of the circuit will be equal to the resistive component.

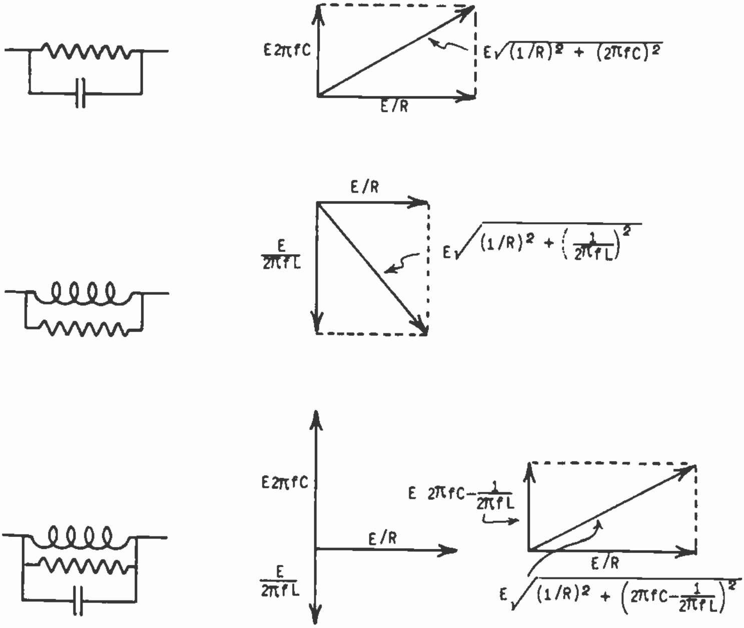


Fig.7 Vector diagrams for parallel circuits.

Vector diagrams can be used for parallel circuits, as well as series. In the case of parallel circuits, we represent the currents through the various branches by vectors instead of the voltages, since in a parallel circuit the current in each branch differs, while the voltage across all the branches is the same. The vector sum of the branch currents will be the line current. Fig. 7 shows the vector current diagrams for several parallel combinations. In these diagrams, the vector representing the current through the capacity is drawn up, and the vector representing the current through the inductance is drawn down, as the current through the condenser leads the current through the resistance, while the current through the inductance lags the current through the resistance. If the vector E is removed from all the components, the resultant vector diagram represents the admittances of the components and the combination.

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MIDLAND TELEVISION INC.

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT
NO.
7

TELEVISION SYNCHRONIZING GENERATORS

LESSON
NO.
3

SYNCHRONIZE

.....to happen simultaneously.

One of the definitions of the word "synchronize" is "to happen simultaneously", which in plain every-day English means that several things happen at the same time.

In television, synchronization is extremely important because neither the transmission nor reception of television signals will produce a high definition picture unless the various units of the equipment involved, function with split-hair precision....and at the same time.

In radio, the half-trained serviceman can "get away" with poor work at times because some of his customers may not be critical of moderate distortion or other defects in his workmanship. However, the inefficient technician will be just plain out of luck in television, for distortion will mean distorted pictures; perhaps a grotesque face, a warped nose, or arms that do not match. While such reception might be considered comical for a time, the eventual reaction would be disgust, and lost sales for the store where the set was purchased.

To assure the success of television, broadcasters will only employ operators capable of operating their equipment properly, and radio stores retailing television receivers will only employ technicians capable of keeping receivers in perfect working order. The men so employed will require a high degree of training...the kind of training that you are receiving from Midland. There will be no place for men with limited knowledge of television problems.

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Lesson Three

TELEVISION SYNCHRONIZING GENERATORS

"The synchronizing generator is unquestionably one of the most important pieces of equipment in the television transmission chain. Without synchronizing impulses it would be impossible to hold a receiver in step with the transmitter, and no picture could be received.

"This lesson may at first appear to be a little complicated; but if you will keep in mind the purpose of the generator, I am sure you will have less trouble in comprehending how it operates."

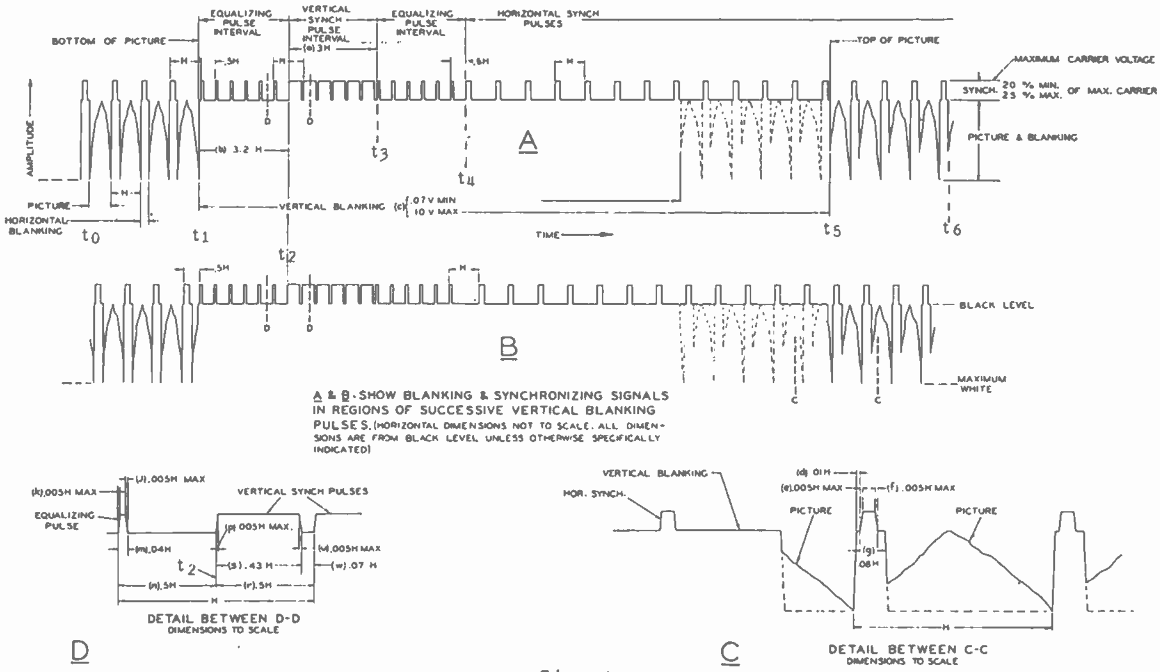
1. THE PROBLEM OF SYNCHRONIZING. Before discussing the transmitting circuits for generating the RMA standard synchronizing impulses, we will consider in detail the reason for choosing this relatively complex standard. The problem of synchronizing has already been discussed in a previous lesson and the fundamental principles discussed in that lesson will now be enlarged upon.

The student has already learned that in order to have interlacing, an odd number of lines must be employed. Why do we choose 441? Would not 431 or 451 be equally satisfactory? To insure a steady picture with good interlacing, it is necessary that some definite odd number of lines be employed for each frame. In order to maintain a definite number of lines, it is necessary that the frame frequency and line frequency be related by simple numbers so that stable harmonic generators can be used to maintain the definite numerical relation between frame frequency and line frequency. The electrical means for accomplishing this will be discussed later in the lesson.

If the frame frequency and line frequency are related by the number 441, then it is possible to use frequency multipliers or frequency dividers in cascade which bear a frequency ratio between input and output of 7, 7, 3, 3, ($7 \times 7 \times 3 \times 3 = 441$), to obtain the line frequency from the frame, or the converse. If we wish to have a picture of approximately 400 lines, no other odd number except 441 can be chosen which may be formed by multiplying small numbers, such as 3, 7, or 9.

RMA STANDARD T-111
TELEVISION SIGNAL

441 LINES, 30 FRAMES PER SEC., 60 FIELDS PER SEC., INTERLACED



A & B - SHOW BLANKING & SYNCHRONIZING SIGNALS IN REGIONS OF SUCCESSIVE VERTICAL BLANKING PULSES. (HORIZONTAL DIMENSIONS NOT TO SCALE. ALL DIMENSIONS ARE FROM BLACK LEVEL UNLESS OTHERWISE SPECIFICALLY INDICATED)

Fig. 1

The special synchronizing waveform shown in Fig. 1 was devised so that the vertical and horizontal synchronizing impulses could be separated while leaving the horizontal impulses continuous even during the time of vertical synchronization. Furthermore, the vertical synchronizing impulses are identical for every field, regardless of the presence of the horizontal impulses. The waveform of Fig. 1 will be discussed part by part throughout the lesson, but first we will consider the figure as a whole. The time represented as t_2 may be considered as the true reference of the figure since this represents the beginning of a frame. The vertical synchronizing pulse which begins at t_2 and ends at t_3 lasts for a time equivalent to three horizontal lines in the picture. During this time, six notches are cut in the vertical synchronizing pulse. These notches or serrations occur, therefore, twice per line, and the first is displaced from the beginning of the synchronizing pulse by a time equal to half a line of the picture.

The equalizing pulse intervals which occur just before (t_1-t_2) and just after (t_3-t_4) the vertical synchronizing, also last for a time equal to three lines of the picture. During these intervals, synchronizing pulses occur twice per line, instead of just once.

The intervals t_0-t_1, t_4-t_6, represent normal horizontal synchronizing pulses. Of course, horizontal synchronizing pulses also occur before t_0 and after t_6. In fact, Figs. 1A or 1B would have to show approximately 200 horizontal pulses to be complete.

The interval t_1-t_5 represents the vertical blanking pulse and lasts from 16 to 22 lines of the picture.

The horizontal blanking pulse lasts approximately 10% of the time for one horizontal line. The horizontal synchronizing pulse lasts approximately 8% of the time for one horizontal line as may be seen in Fig. 1C.

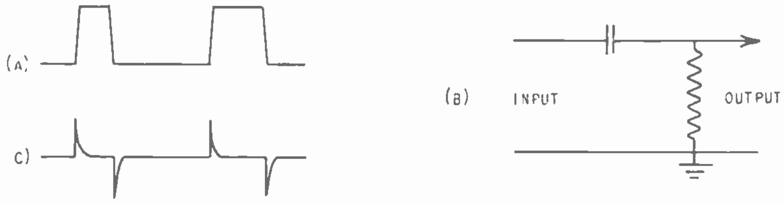


Fig.2 Transmission of rectangular pulses through a high-pass filter.

Horizontal and vertical synchronizing impulses are separated from each other by reason of their different time durations. Thus a low-pass filter is used to extract the vertical impulses and a high-pass filter is used to extract the horizontal impulses. If a square wave is passed through a high-pass filter, the higher frequencies which constitute the steep slopes on the leading and trailing edges of the square wave are passed by the filter. The low frequency components which determine the time duration of the square wave are filtered out. Thus, on the leading and trailing edges of the square wave, a sharp pulse is generated. Consider Fig. 2A. In this figure, two successive square waves are shown which have unequal durations. If the slopes of the leading and trailing edges are identical in the two pulses, then the output of the high-pass filter shown in Fig. 2B will give a series of sharp transients, whose magnitude and shape are independent of the width of the square wave. The output of the circuit shown in Fig. 2B is illustrated in Fig. 2C directly under the square wave from which it was derived. The high-pass filter shown in this circuit is essentially the same network used for separating the horizontal synchronizing impulses in a television receiver.

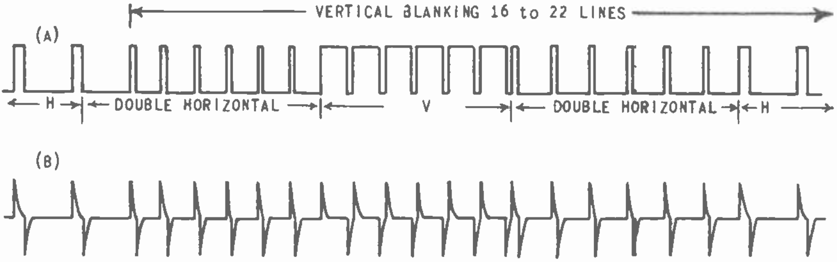


Fig.3 Transmission of a standard synchronizing signal through a high-pass filter.

Let us apply the standard synchronizing waveform to the circuit of Fig. 2B and observe the output. A portion of the standard wave is shown in Fig. 3A and the output from the horizontal separator circuit is shown in Fig. 3B. It may be seen that during a part of the vertical blanking out time, double horizontal impulses are inserted as well as a serrated[1] vertical impulse in addition to the normal horizontal pulses. Each one of the double horizontals delivers a transient through the high-pass filter exactly the same as that delivered by the ordinary horizontal impulse. Consequently, a synchronizing pulse occurs not only at the beginning of a line, but also at the center. Furthermore, the notches or serrations in the vertical synchronizing impulse also deliver transients through the high-pass filter which serve as horizontal synchronizing impulses, even during the time of the vertical synchronizing impulse. Thus the receiver is supplied with horizontal pulses during the time of the vertical blanking out; the horizontal oscillator is running in step; and the picture does not tear out at the top as it would if the serrated vertical were not employed. Twice as many horizontal impulses are supplied as necessary during part of the time, but every other one is of no consequence since the repetition rate is so far removed from the normal frequency of the horizontal oscillator. Every other pulse, and *only* every other one, will fire the horizontal oscillator in the receiver.

Since the horizontal pickoff circuit depends not on the duration of the square waves but on the steepness of the wave front, it should be noted that the normal horizontal synchronizing pulses,

---

[1] Serrated: toothed or notched, like a saw.

the double horizontal synchronizing pulses, the serrations in the vertical pulse, and the leading edge of the vertical pulse must all have the same slope. Otherwise, there will be an irregularity in the resultant transient through the horizontal separation circuit and the pulses shown in Fig. 3B will not all be of the same height. The details of the shape of the various pulses are shown in Figs. 1C and 1D and, as just mentioned, all of the pulses have the same slope on the leading edge.

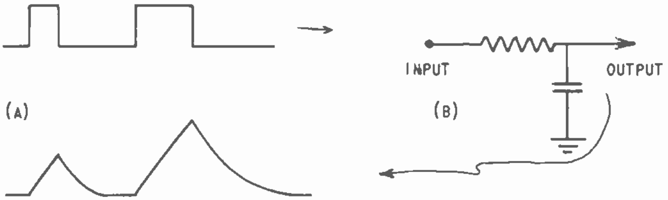


Fig. 4 Transmission of rectangular pulses through a low-pass filter.

Next we shall consider the operation of the vertical separation circuit. This circuit is essentially a low-pass filter as shown in Fig. 4B. If the waveform of Fig. 4A is impressed upon this low-pass filter, the rectangular pulses of unequal time duration yield unequal output, even though the slopes of the wave fronts are the same. This is due to the longer charging time for the wider pulse. The result is the exact opposite of the high-pass filter which was illustrated in Fig. 2. The low-pass filter stores up the energy in the square wave and thus the longer the duration of the pulse, the greater will be the output amplitude of the vertical separation circuit.

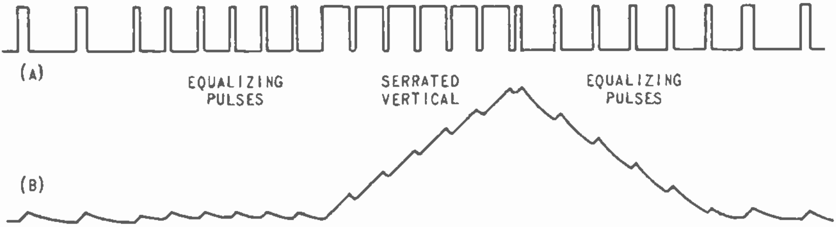


Fig. 5 Transmission of a standard synchronizing signal through a low-pass filter.

The action of a low-pass filter when fed with a standard synchronizing wave is shown in Fig. 5. During the time of the normal horizontal synchronizing impulses, the low-pass filter charges slightly and discharges for each line of the picture. Because of the exceedingly short time duration of the horizontal synchronizing pulse, only a small charge is accumulated by the condenser in the vertical separation circuit. During the time of the double horizontal synchronizing pulses, this condenser would normally charge up to a greater value because of the larger number of pulses. However, the pulses are made much narrower, in fact only half as wide as the normal horizontal pulses. Although there are twice as many pulses, each pulse charges the condenser only half as much as previously, and the net charge remains the same. The vertical synchronizing pulse is very long. It takes up a time equal to three complete horizontal lines of the picture. Therefore, the condenser in the separation circuit charges up to quite a high value and, at the end of the vertical synchronizing pulse, it slowly discharges to the value it had before the vertical synchronizing impulse was inserted. The serrations in the vertical synchronizing pulse allow the condenser to discharge slightly, but since these serrations are small compared to the space between them, the condenser does not discharge more than a fraction of the energy it had accumulated during the charging time. The condenser continues to build up as long as the vertical synchronizing pulse is present as shown in Fig. 5B.

The gentle rise of the waveform of Fig. 5B is a disadvantage in controlling the precise firing time of the vertical oscillator in the receiver. If the firing time of the vertical oscillator differs by more than the time required to scan a fraction of a horizontal line, pairing will result. For this reason, receivers are often arranged with circuits which clip the wave of Fig. 5B, square it up, and by this means increase the slope of the vertical wave front. Although this is a receiver problem, the circuits for accomplishing the result are identical with ones to be described later in this lesson.

The double horizontal pulses to which we have referred are known as equalizing pulses. Six of these pulses occur before and six after the insertion of the vertical synchronizing pulse. As already mentioned, these pulses occur with twice the frequency and half the width of the standard horizontal pulses.

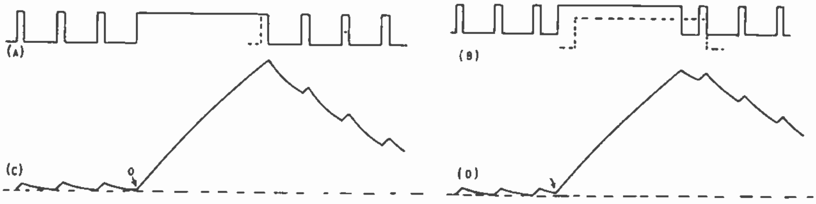


Fig.6 Transmission of simple synchronizing pulses through a low-pass filter.

The best way to describe the purpose of the equalizing pulses is to consider what would happen if these pulses were not present. In order to secure satisfactory interlacing by the odd-line method, the vertical synchronizing pulse is inserted in such a way as to trip the vertical oscillator at the end of a horizontal line during one field and to trip the vertical oscillator in the middle of the horizontal line during the following field. Thus, a simple synchronizing pulse without the refinement of the serrated vertical and the equalizing pulses would appear as shown in Fig. 6A and in Fig.

6B for alternate fields of one frame. The dotted pulse in Fig. 6B shows the relative position that this pulse had in the preceding field with respect to the horizontal pulses. Two things are apparent: first, the leading edge of the vertical pulse is closer to a horizontal pulse in Fig. 6B than in Fig. 6A. Second, the vertical pulse of Fig. 6A joins a horizontal pulse, and thus appears as if the vertical pulse is wider in Fig. 6A than in Fig. 6B. These differences between alternate fields affect the time of firing of the vertical oscillator in the receiver and also the magnitude of the vertical deflection. The effects are rather small but sufficient to upset the delicate process of interlacing.

Consider the effect on the pulses of Fig. 6A and Fig. 6B in passing through the low-pass filter which constitutes the vertical separation circuit. When the vertical pulse lies close to a horizontal pulse, the condenser in the low-pass filter has not completely discharged at the beginning of the vertical synchronizing pulse. Thus, the leading edge of the vertical pulse of Fig. 6D begins at a higher voltage than in the case of Fig. 6C. Since the slope of the condenser charging current is the same in either case, the voltage necessary to trip the vertical oscillator in the receiver will occur sooner for the field represented by the Figs. 6B-D than for the field represented by Figs. 6A-C, due to the presence of the adjacent horizontal synchronizing pulse. Thus, the vertical oscillator in the receiver will fire sooner than it should every other field and cause pairing or imperfect interlacing in the receiver.

As already mentioned, the vertical pulse in Fig. 6A is of longer time duration than that in Fig. 6B, due to the extra horizontal pulse which adjoins the trailing edge. This means that the condenser in the vertical separation circuit will charge longer for the case of Fig. 6A than for the succeeding field, and the magnitude for the synchronizing pulse leaving the separation circuit will be greater for the first field than for the second. The energy of the synchronizing pulse unavoidably adds to the energy of the discharge circuit in the receiver, hence the inequality of the successive vertical synchronizing pulses cause the vertical oscillator in the receiver to produce unequal magnitudes of deflection. If alternate deflection waves differ by as much as .1%, poor interlacing will result over part of the picture.

In the preceding paragraphs we have considered in detail the effect of simple synchronizing pulses in disturbing the quality of odd line interlacing. The standard synchronizing pulses, shown in Fig. 1, are arranged to overcome this difficulty. By doubling the number of horizontal synchronizing pulses just before and just after the vertical synchronizing pulse, the output of the vertical separation circuit is made to appear identical for every vertical synchronizing pulse. This action was shown for one field in Fig. 5, and would be exactly the same for every field. As may be seen from Fig. 1, the difference between the fields occurs not at the beginning of the vertical synchronizing pulse, but at the beginning of the equalizing pulses. The slight differences at the beginning of the equalizing pulses for alternate fields is of no consequence since the vertical oscillator will not fire at this time. The equalizing pulse interval is made sufficiently long to restore equilibrium before the vertical synchronizing impulse occurs.

2. ELECTRONIC GENERATOR. The standard synchronizing impulses which have just been described may be generated either by electromechanical means or by an electronic generator. Each system has its own advantages and disadvantages. It is difficult to secure sufficient precision by mechanical means in order to make successive synchronizing impulses identical in shape and timing. On the other hand, the electronic generator is relatively complex and the large number of tubes and circuits involved tend to make a unit which is not entirely reliable unless all components are operated very conservatively. Even then it is desirable to have a duplicate unit.

The circuit of the electronic generator is quite complex and each manufacturer differs to a considerable extent in most of the detailed circuits. However, the basic principles underlying all of them are essentially the same. For this reason, the basic circuit will be discussed in block diagram form and then each portion of the circuit analyzed and consideration given to the many possible variations. Fundamentally, the synchronizing generator consists of three parts as shown in Fig. 7. The first is a timing unit which generates the various pulses at the correct frequency, next the pulse shaping unit which takes these pulses and forms them into the shapes illustrated in Fig. 1, and third the output unit which supplies pulses of various shapes in duplicate to several low impedance outputs.

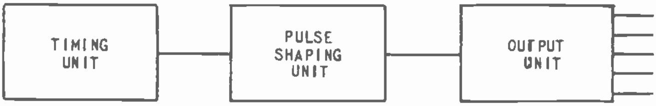


Fig. 7 Fundamental units of an electronic synchronizing generator.

The basic functions of each of the three units comprising the synchronizing generator are shown in Fig. 8. Consider one unit at a time. The timing unit consists of a master oscillator and a chain of multivibrators operating at successively lower frequencies. The master oscillator may be any conventional type of sine wave oscillator and may oscillate on either 13.23 or 26.46 kc. In Fig. 8 it is shown as operating on 26.46 kc. This oscillator has synchronized with it a multivibrator on the same frequency. This multivibrator supplies narrow pulses which eventually are used as equalizing or double horizontal pulses preceding and following the vertical synchronizing pulse. The 26.46 multivibrator drives a 13.23 kc. multivibrator which generates the horizontal blanking impulses. The 26.46 multivibrator also drives a chain of four successive multivibrators which divide the frequency by 1/7, 1/7, 1/3, and 1/3. The individual frequencies corresponding to these divisions are 3780, 540, 180, and 60. The 60 cycle rectangular pulse which is derived from this chain of multivibrators is used as a vertical blanking

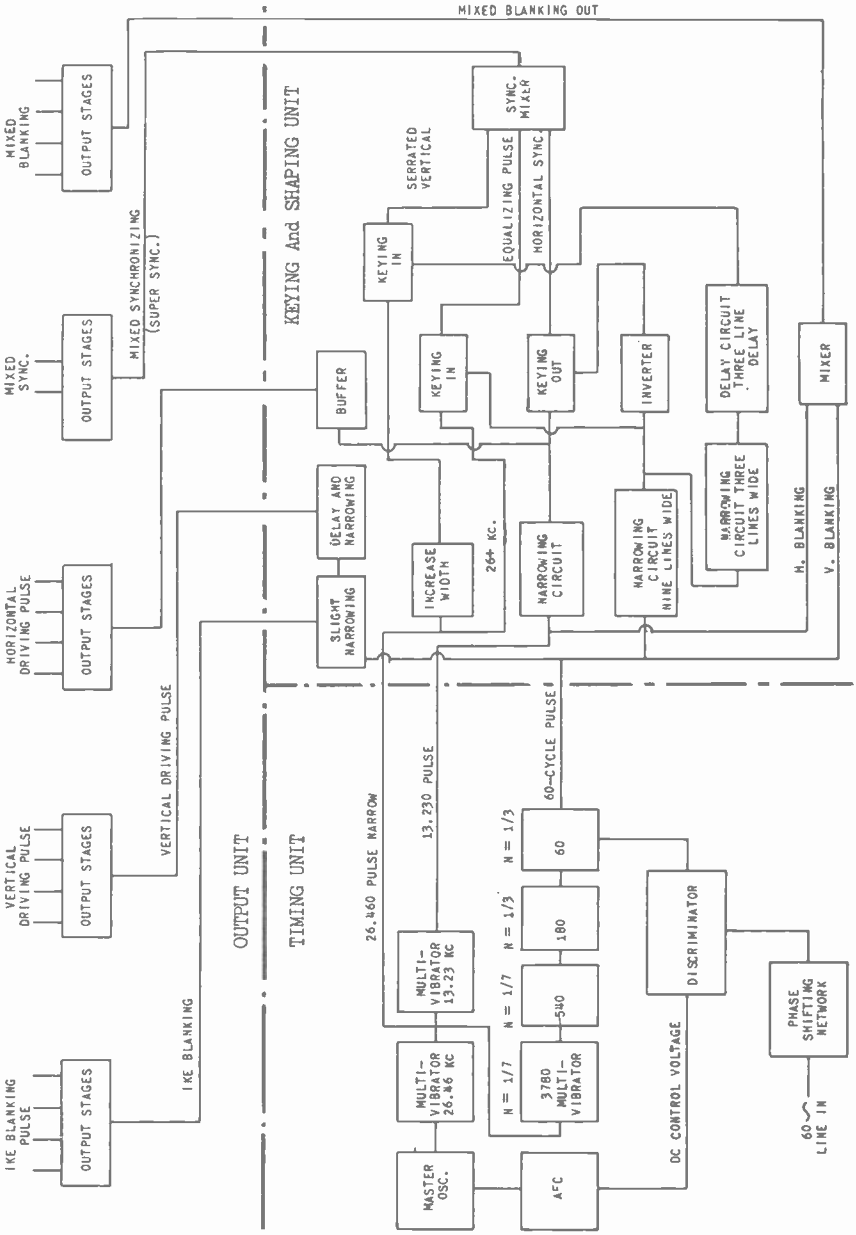


Fig.8 Block diagram of an electronic generator.

pulse and for other purposes which will be considered when we study the shaping unit.

The chain of oscillators which has just been described insures that a definite frequency relation is maintained between the line frequency (13,230 c.p.s.) and the frame frequency (30 c.p.s.). This ratio will be 441 to 1. However, it is necessary to insure that the 60 cycles generated by this chain of oscillators maintains a consistent phase relationship to the 60 cycles supplied by the power source. This relation is secured by a discriminator which compares the 60 cycle square pulse to the line voltage. Any difference in phase develops a DC voltage which is used to control the frequency of the master oscillator running at 26.46 kc. This will be considered in detail a little later on. A phase shifting network is also incorporated in order to lock the 60 cycle square wave at any phase to the line voltage.

The output of the timing unit consists of three sets of pulses whose fundamental frequencies are 26,460, 13,230, and 60 cycles. These outputs are fed to the keying and shaping unit.

The keying and shaping unit must take the fundamental pulses derived in the timing unit and form them into the complicated waveform which was shown in Fig. 1. This wave shape is seen to consist of a combination of five elementary waveforms:

1. Horizontal blanking
2. Vertical blanking
3. Horizontal synchronizing
4. Equalizing pulses
5. Serrated vertical pulse

Besides these five pulses, two others must be formed which serve as keying impulses to turn on and off the others at the correct time. The keying pulses are formed from the vertical blanking pulse, one beginning at the same time as the vertical blanking, but continuing for a period of only nine horizontal lines, instead of 21 horizontal lines. The other keying pulse lasts for only three lines and is delayed by a time equal to three horizontal lines from the beginning of the vertical blanking pulse.

The connection for the various shaping circuits may be seen in Fig. 8, but the mode of operation is more clearly indicated in Fig. 9, which shows how the various pulses are keyed so that they occur at the proper time and are then mixed in a combining stage. Occasionally, the slightly more complicated keying circuit of Fig. 10 is used and, in this case, the equalizing pulses are keyed out during the time that the serrated vertical pulse is keyed in. This refinement is not generally necessary since (as shown in Fig. 11), the trailing edge of the serrations in the vertical impulse occur at exactly the same time as the leading edge of an equalizing pulse. The coincidence is essential, because it is the upswing of the waveform that delivers the horizontal synchronizing pulse. The presence of the equalizing pulses in no way deters the operation of the circuit; however, some means must be used to limit the peak output of the stage. Fig. 11 shows the relative position of the equalizing pulses and the serrated vertical pulse. Fig. 11B shows the output secured from the circuit of Fig. 9 if the peaks are limited, and Fig. 11C shows what would occur if the two sets of impulses were simply added together by paralleling the plate circuits of two tubes.

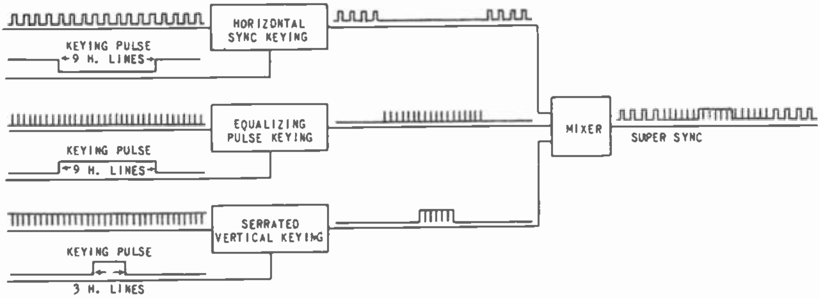


Fig. 9 Block diagram of keying circuit.

In this case, the signal would have to be trimmed along a line such as X-X to create the proper wave shape.

After the synchronizing pulses have been mixed as shown in Fig. 9 or Fig. 10, it is often referred to as "supersync". The supersync is not mixed with the blanking pulses in the synchronizing generator itself. As you learned in the preceding lesson, the mixing operation of the picture signal, the blanking signal, and the synchronizing signal takes place directly in the studio amplifier, and the circuits for accomplishing this have already been studied.

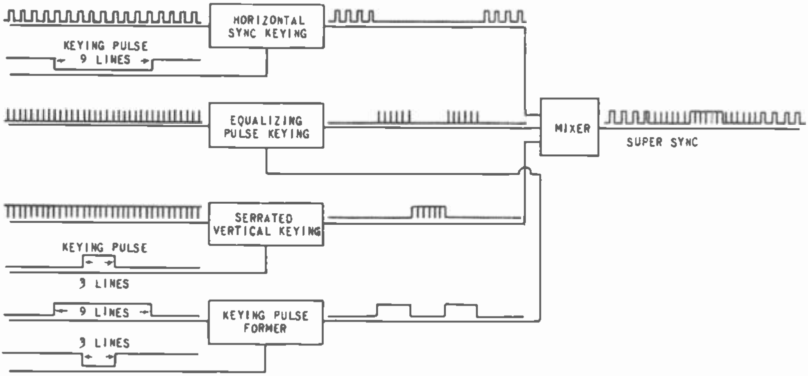


Fig. 10 Keying circuit for forming standard synchronizing signals.

Horizontal synchronizing pulses are formed in the keying and shaping unit from the horizontal blanking pulses simply by making the blanking pulses narrower. The detailed circuit for doing this will be shown later in the lesson.

The serrated vertical pulses are formed from the combination of a keying pulse and the equalizing pulses which have been either widened, or inverted and shifted in time. Notice in Figs. 5 and 9 that when the equalizing pulses are inverted, the appearance of these pulses is quite like the serrated vertical.

The keying pulses are formed from the vertical blanking by the process of narrowing the blanking pulse. The nine line wide vertical pulse is used to key in the equalizing pulses. A three line pulse which has not only been narrowed, but delayed, is used to key in the serrated wave. In addition to the supersync circuits, a circuit is provided for mixing the horizontal and vertical blanking.

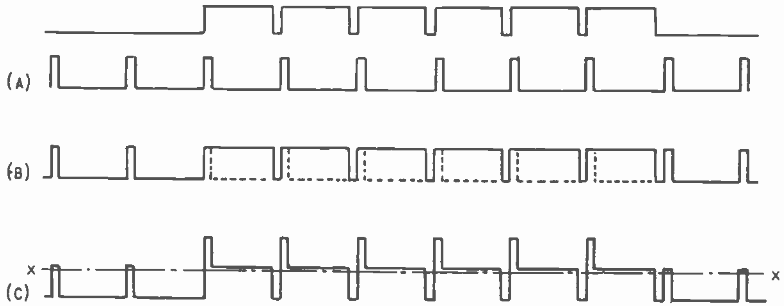


Fig. 11 Relative timing of equalizing and serrated vertical pulses.

In addition to its primary function of a synchronizing and blanking unit, the electronic generator usually furnishes several other special pulses. Vertical and horizontal pulses are formed which serve to drive the deflection circuits for the Iconoscope. The circuit required to generate a sawtooth wave from a narrow pulse is the familiar discharge circuit which you studied in a previous lesson. In this way, the blocking tube oscillator can be omitted from the Iconoscope deflection circuits and thus the possibility of the Iconoscope deflection failing to synchronize is eliminated. The driving pulses must have a width somewhat greater than the synchronizing pulses and considerably less than the blanking pulses. If the width of the driving pulse is too small, the return time of the deflection circuit will be so short that extremely high voltages will be developed in the deflection yoke and output circuit of the deflection amplifier. If the driving pulses are too wide, then the return time may not occur entirely within the interval of blanking, and thus transients will be observed on the edges of the picture. As shown in Fig. 8, the vertical driving pulses are formed from the vertical blanking pulses by narrowing these pulses and introducing a slight delay so that the return of the Iconoscope deflection cannot begin until after the leading edge of the vertical blanking-out. The horizontal driving pulse is generally identical with the horizontal synchronizing pulse and is extracted from the horizontal synchronizing circuit through a buffer.

The synchronizing generator should also furnish a pulse to blank the Iconoscope during the return time and thus remove the beam during the flyback in the Iconoscope. The Iconoscope blanking pulse is slightly less wide than the normal vertical blanking, but somewhat wider than the vertical driving pulse. Thus, there are at least three special pulses besides those used in forming the synchronizing and blanking, or a total of 11 pulses of various shapes and widths *formed* by the synchronizing generator.

The actual number of *output* wave shapes are only five, since the blanking pulses are mixed previous to the output stages as are the synchronizing pulses. However, for each of the five different types of waves, a multiplicity of output connections should be provided. The various waves should be supplied in either positive or negative polarity to facilitate changes in the rest of the studio equipment or in adapting the synchronizing generator to studio equipment supplied by various manufacturers. A multiplicity of outputs is necessary because it is general practice to incorporate at least two and preferably three or four cameras in the studio in addition to two or three movie machines (for flexible program facilities and continuity of programming). The outputs should preferably be supplied at low impedance, such as 70 ohms, in order that they may be carried over coaxial cables to various parts of the studio and control room. The block diagram of Fig. 8 shows four outputs for each type of pulse except the mixed synchronizing where generally only one is needed. The four output circuits would supply two studio cameras and two movie cameras, assuming that the polarity could be reversed at will within the output stage itself.

Fig. 12 shows the front view of the RCA synchronizing generator, complete except for power supplies which are housed in a separate rack. The block diagram of Fig. 8 does not refer to this particular unit, but was described to show the equipment generally included in any synchronizing generator. There are apt to be variations from this. For instance, some units include the vertical and horizontal deflection circuits in their entirety, while other units, such as the Dumont synchronizing generator, shown in Fig. 13, also include an oscilloscope for adjusting and maintaining adjustments in the synchronizing generator.

3. TIMING CIRCUITS. The circuit of Fig. 8 need not be followed explicitly in the timing unit. In an elaborate unit, buffers would be employed between each stage and, as mentioned previously, the master oscillator might operate at 13.23 kc., rather than at 26.46 kc. One variation in block form is shown in Fig. 14. A simpler and more compact unit might omit the buffers and divide the frequency by 9, 7, and 7, instead of coming down 7-7-3-3. Such a simplified unit is shown in Fig. 15. Still another variation would be to run the master oscillator at 105 kc., instead of 26 kc., and the 105 kc. output could be inserted in the video chain instead of the picture signal. This would place eight vertical bars across the picture which could be used for checking the linearity of the horizontal deflecting circuits in the monitor or receiver by adjusting the deflection until the bars are uniformly spaced. It might occur to you that it would be simpler to operate a chain of frequency multipliers directly from the 60 cycle line frequency to

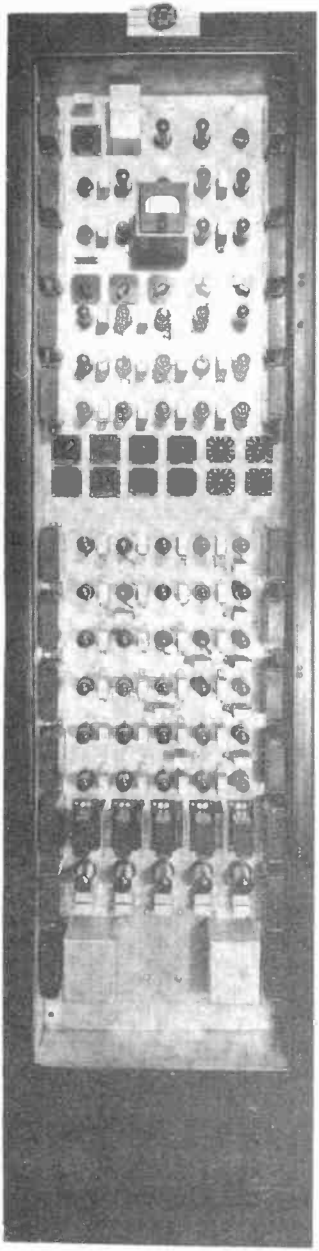


Fig.12 RCA synchronizing generator.

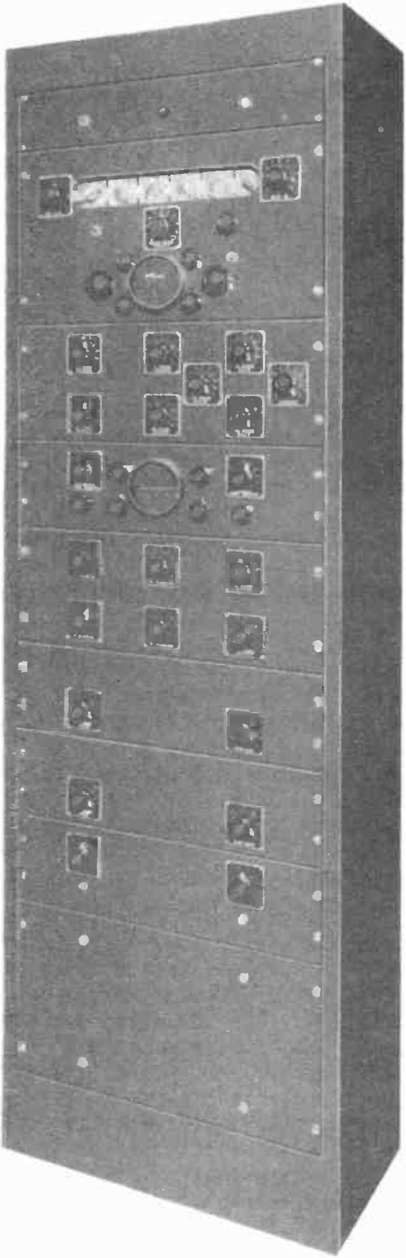


Fig.13 Du Mont synchronizing generator.

achieve the 441 relation between the horizontal and vertical impulses. Such a system is diagrammed in Fig. 16 and appears somewhat simpler than the system previously described which started with a high frequency oscillator and divided down to 60 cycles, comparing this latter frequency with the line frequency and controlling the master oscillator from the comparison circuit. However, the circuit of Fig. 16 will not generally give sufficient precision in the timing of the horizontal pulses.

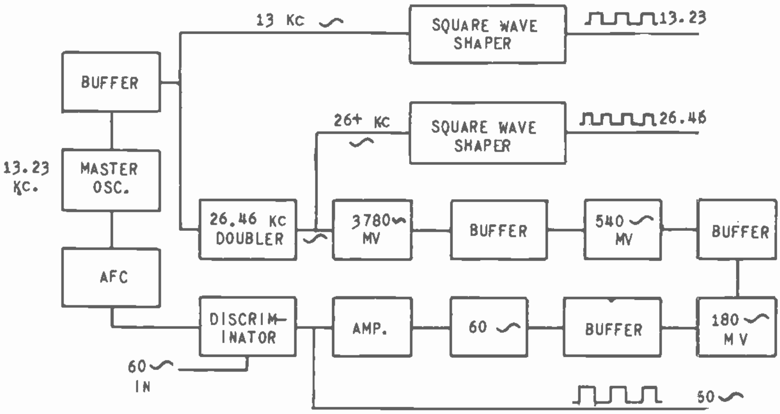


Fig.14 Block diagram of an electronic timing unit.

In order to secure a picture having high definition and freedom from ragged edges, the horizontal oscillator must fire at a predetermined period to within 1/5 microsecond. This means that the frequency multiplier which is driven by the 60 cycle line frequency

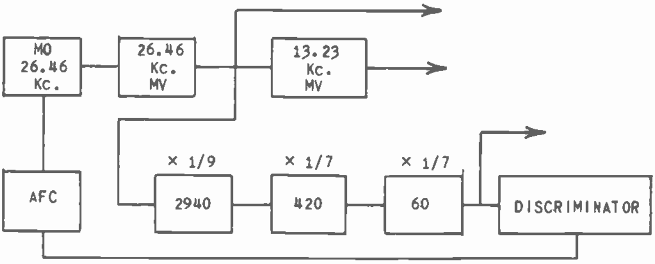


Fig.15 Simplified timing unit.

must not only be an exact multiple of 60 cycles, but must maintain the phase of this frequency to within .004 degree of a 60 cycle wave or to within .04 degree of the first multiplier of Fig. 16. Such extreme precision in the frequency multiplier is not ordinarily realizable and the frequency multipliers tend to "hunt" slightly in phase. In other words, the 441 relation between line frequency and frame frequency can easily be maintained to within one line, but it cannot readily be maintained to within a very small fraction of one line which is necessary to secure a perfect picture. Thus, the simpler circuit of Fig. 16 does not give as excellent results as the previously described type.

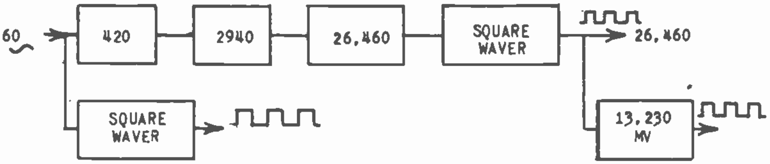


Fig.16 Timing unit utilizing frequency multipliers.

Another difficulty with the circuit of Fig. 16 is that slight irregularities in the line frequency such as a sudden surge, caused perhaps by turning on a switch, will be carried through the frequency multipliers and affect the line frequency pulses of the synchronizing generator. In the previously described circuit which included a master oscillator running at a high frequency, sufficient time delay is incorporated in the AFC circuit to eliminate the effect of instantaneous changes in the line frequency.

4. MULTIVIBRATORS. Multivibrators have been mentioned as the type of circuit used in dividing the frequency of the master oscillator to secure the vertical synchronizing impulses. In some cases, it is preferable to use blocking tube oscillators for this purpose since they may readily be synchronized on a subharmonic of a previous stage. Blocking tube oscillators are usually more stable than multivibrators, particularly at the lower frequencies, and for this reason they are sometimes incorporated in some of the frequency divider stages of the timing unit. Blocking tube oscillators have been thoroughly discussed in a previous lesson and will not be discussed here; however, there are types of multivibrators other than the one described in a previous lesson. Several types of multivibrators suitable for use as frequency dividing units will now be described.

A multivibrator is fundamentally a two stage resistance-coupled amplifier with feedback between the two stages. This feedback may be applied either from plate to grid or it may be achieved through common cathode coupling. Three variations of the plate to grid feedback method are shown in Figs. 17, 18, and 19, while circuits employing common cathode feedback are shown in Figs. 20 and 21.

The performance of the basic multivibrator circuit of Fig. 17A may be understood by following through the waveforms shown in Fig. 17B. These waveforms show the grid voltage and the plate current of the two triodes incorporated in the multivibrator circuit. Assume that the multivibrator is turned on and some slight disturbance causes a fluctuation in the plate voltage or plate current. If the voltage on the grid of T_1 changes ever so slightly in the negative direction, this change will be amplified by T_1 and T_2 and will be reimpressed upon the grid of T_1 as a still further negative change, finally biasing tube T_1 to cutoff. Under this condition, the plate of T_1 is highly positive (zero current) causing tube T_2 to draw considerable grid current. This grid current limits the voltage to

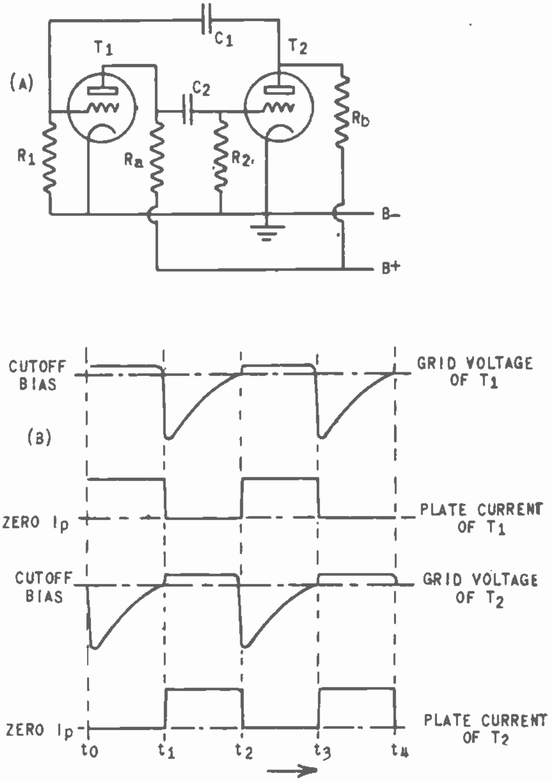


Fig.17 Basic multivibrator.

which the grid of T_2 can be driven as shown in the diagram of Fig. 17B. Thus, between the times shown at t_1 and t_2, the grid of tube T_2 maintains a constant positive voltage. The negative charge impressed upon the grid of T_1 by C_1 slowly leaks off via the grid resistor R_1 until finally (at time t_2), cutoff bias is reached and T_1 again draws current. The drop in plate voltage (current increase) of T_1 is amplified by T_2 and impressed as a positive voltage on the grid of T_1 (limited by grid current). This change takes place almost instantaneously (at time t_2) and the grid of tube T_2 is driven highly negative. During the next interval of time from t_2 to t_3, the condenser C_2, which was charged by grid current during the previous period, now discharges through resistance R_2 until T_2 once more draws some plate current. At this instant (t_3), the grid of T_1 will be driven negative once more, having completed an entire cycle. Thus the multivibrator oscillates, producing rectangular current pulses as shown in the diagram, first one tube drawing current, and then the other. The time to complete a cycle of two pulses is determined by the time constant C_1R_1 and C_2R_2. If these time constants are made shorter, the multivibrator will operate at a higher frequency.

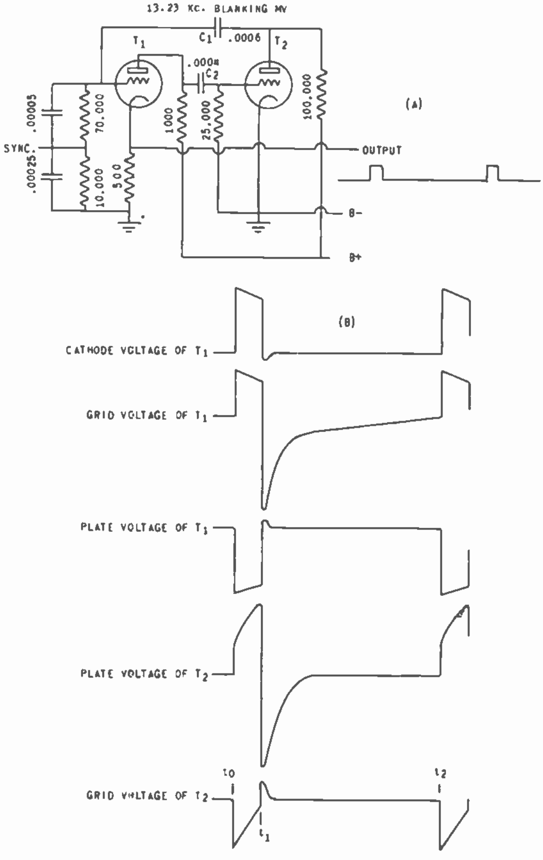


Fig.18 Typical multivibrator for synchronizing generator.

In Fig. 17, the current pulses of tube T_1 and tube T_2 were shown to have equal periods. This need not be the case. If the time constant R_2C_2 is made different than the time constant R_1C_1, then one of the rectangular pulses will be longer than the next. Carrying this condition to an extreme, it is possible to generate a narrow pulse suitable for synchronizing purposes.

If a small resistance is inserted in the cathode of one tube, the plate current pulse can be utilized as a voltage pulse by deriving the output voltage across the cathode resistor. This is done in Fig. 18. In this circuit, the plate resistors are no longer small compared to the grid resistors and must be figured in the time constants which relate to the charge and discharge of the coupling condensers. These are arranged to obtain a pulse width equal to 10% of the time for a horizontal line. The pulses derived from the cathode circuit are utilized as horizontal blanking pulses.

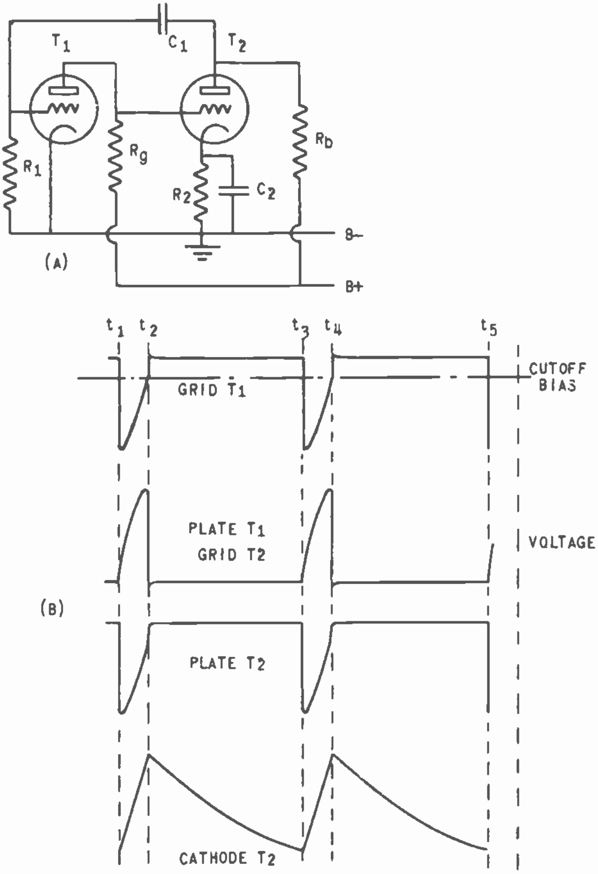


Fig.19 Bedford-Puckle multivibrator.

The waveforms of Fig. 18B show the manner in which the circuit operates. The firing time (t_1) of tube T_2 depends not only upon the rising grid voltage, but also upon the rising plate voltage since C_1 is charging and C_2 discharging at the same time. This condition is not true for tube T_1 whose load resistor is quite small.

The firing time of this tube (t_2) is determined almost entirely by the discharge time of C_1. The charging time of C_2 is immaterial because it becomes completely charged long before C_1 discharges.

The small condensers across the grid resistor of T_1 are for the purpose of properly timing the pulse which synchronizes this stage and will be discussed later in the lesson.

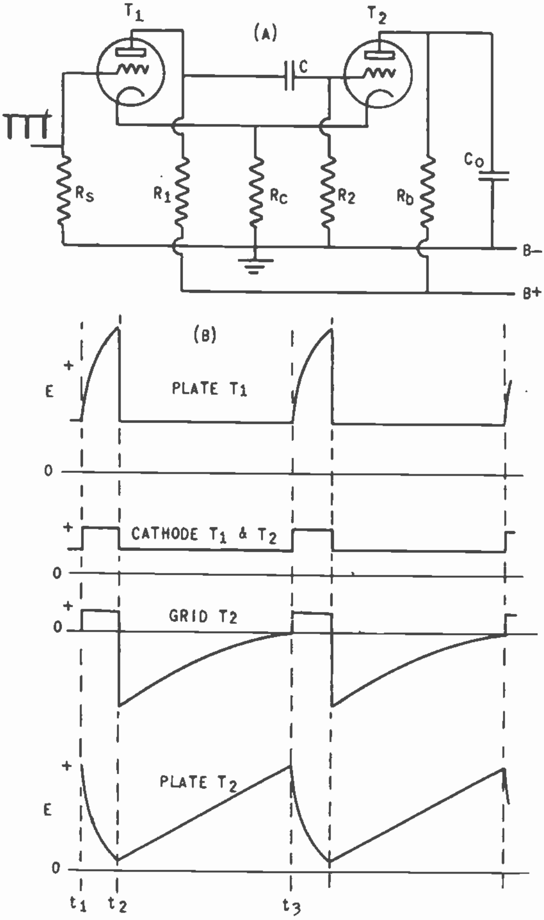


Fig.20 Potter multivibrator.

The Bedford-Puckle multivibrator shown in Fig. 19 was described in a previous lesson. The operation of this circuit will not be repeated here except that Fig. 19B shows the waveforms of various portions of the circuit.

Fig. 20 shows the Potter multivibrator which develops feedback between tubes T_2 and T_1 by means of a common cathode resistor.

This type of multivibrator has the advantage that the grid circuit of tube T_1 plays no part in the functioning of the oscillator and thus may be used more readily for synchronizing injection. The multivibrator shown in Fig. 18 makes use of a positive pulse for synchronizing while the circuit of Fig. 20 uses a negative pulse.

When plate voltage is applied to the circuit of Fig. 20, the condenser C charges through R_1 and the grid cathode resistance of tube T_2. The grid current of tube T_2 limits the voltage which can be had at that point as may be seen in one of the curves of Fig. 20B. The grid current flowing through tube T_2 places a positive voltage on the cathode, which biases the tube T_1 to cutoff. However, as condenser C charges, the plate voltage gradually rises on tube T_1 until it finally begins to draw some current. As soon as current starts to flow in tube T_1, the drop across resistor R_c is still further increased, which diminishes the grid current of tube T_2. This allows the plate voltage on T_1 to rise slightly and the cathode voltage to decrease, causing the tube to suddenly draw a considerable amount of plate current, driving tube T_2 to cutoff.

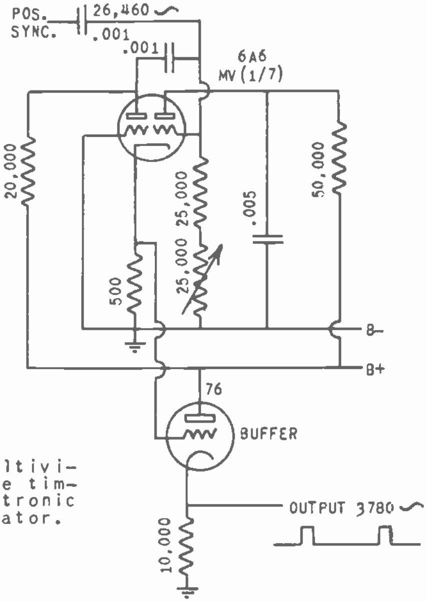


Fig.21 Potter multivibrator as used in the timing unit of an electronic synchronizing generator.

Thus, between the times t_2 and t_3, tube T_2 is cutoff. However, condenser C now discharges through resistor R_2 and eventually tube T_2 begins to draw grid current, thus initiating the cycle again. Condenser C_o plays no part in the functioning of the multivibrator, but causes the plate voltage of tube T_2 to vary in sawtooth waveform. If the condenser C_o is omitted, the plate voltage of tube T_2 is in the form of pulses of opposite polarity to those generated across the cathode resistor.

A Potter multivibrator circuit in the form usually used in synchronizing generators is shown in Fig. 21 in conjunction with its output buffer tube.

5. AUTOMATIC FREQUENCY CONTROL. In Figs. 14 and 15, the block diagrams of the timing unit indicated a discriminator which compared the phase of the 60 cycle output with the 60 cycle line frequency to control the frequency of the master oscillator. The details of this circuit may be seen in Fig. 22. Let us assume for the moment that the DC control voltage from the discriminator is zero when the 60 cycle line voltage is 90° out of phase with the output of the last multivibrator in the timing unit. Let us further assume that if the timing unit speeds up so that the phase of the last multivibrator is advanced with respect to the line voltage, the discriminator will generate a negative DC control voltage. If the output of the timing unit slows down, the discriminator will generate a positive voltage because the phase of the timing unit is retarded with respect to the comparison or line frequency. In a later paragraph, we will show how this condition is fulfilled, but we will first show the effects of the DC control voltage which we have assumed to exist.

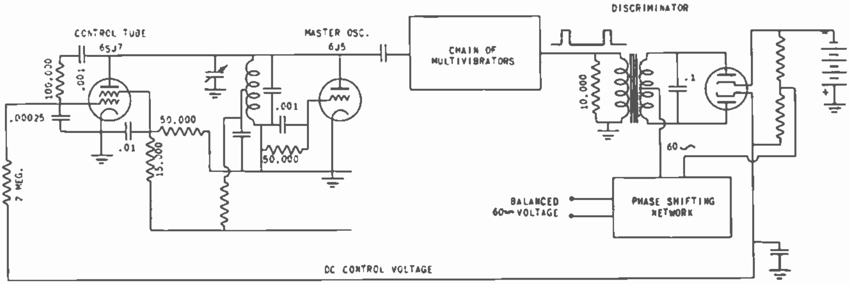


Fig.22 Automatic frequency control circuit of synchronizing generator.

In Fig. 22, the control tube has its plate circuit connected across the tank of a Hartley oscillator. Thus, an AC voltage is applied to the plate of the control tube. This voltage is also applied to the grid through a condenser (.001) and resistor in series. The condenser is merely a blocking condenser since its reactance is very low compared to the resistance between plate and grid (100,000 ohms). Thus, as far as AC is concerned, it is simply a resistor connected between the plate and grid of the control tube. A condenser appears between grid and ground of the control tube. The reactance of this condenser is also small, compared to the series resistor. The equivalent network is shown in Fig. 23A. Since the resistance is large, the voltage across the condenser lags the input voltage by 90°.

In a pentode, the plate resistance is so high that for reasonable values of load impedance, the plate current is always in phase with the grid voltage. Fig. 23B shows the voltage generated by a pentode applied in series with its plate resistance and the external load. Since R_p is large compared to the load Z, the current will always be in phase with the applied voltage. Furthermore, since the grid voltage is 90° out of phase with the plate voltage as shown in Fig. 23A, *then the plate current will also be 90° out of phase with the plate voltage*. Since the plate current lags the plate voltage by 90°, *the tube behaves as an inductance* as far as the tank circuit of Fig. 22 is concerned. Since the tube is placed across the tank circuit and acts like an inductance, it tends to increase the frequency of the master oscillator.

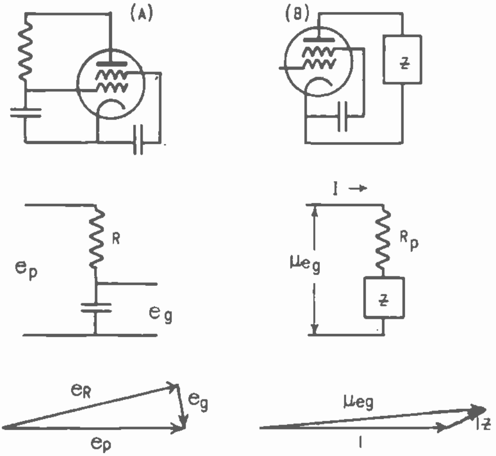


Fig.23 Operation considerations of AFC circuit.

The inductive reactance which the tube reflects across the tank circuit will depend upon the plate current. Since X = E/I (for 90° phase relationship), the higher the plate current, the lower will be the inductive reactance and the greater the frequency change. The plate current depends upon the grid bias of the control tube. Consequently, if the grid is made less negative, the plate current will increase and the master oscillator will be shifted to a higher frequency. If a negative voltage is applied from the output of the discriminator, the control tube will have less effect and the master oscillator will go lower in frequency. The student should note that these changes in frequency are exactly opposite to the changes needed to produce the control voltage. If the master oscillator tends to go higher in frequency, due to some drift on its own accord, the discriminator will immediately generate a negative control voltage which will act to lower the frequency of the master oscillator and consequently curb the tendency for the oscillator drift.

The functioning of the discriminator may be understood by the application of vectors which was thoroughly discussed in a preceding lesson. Fig. 24 shows the application of this method. Suppose that two voltages equal in frequency, but displaced 90° in phase are inserted at points X and Y. Both of the rectifier tubes are supplied with voltages from the two transformers connected in series. However, the voltage across OP is 180° out of phase with the voltage across OQ, because transformer X is centertapped. Fig. 24B indicates that the voltage OR is 90° out of phase with each of the other two voltages. The resultant voltage across tube T_1 indicated by RP in the diagram has exactly the same magnitude as the voltage fed to T_2 and indicated as RQ in the diagram. The outputs of the two rectifiers furnish equal currents to the load, but the currents from these two tubes flow in opposite directions in the load resistor; consequently, the voltage drop across R_1 bucks the voltage drop across R_2 and the resultant output voltage is zero.

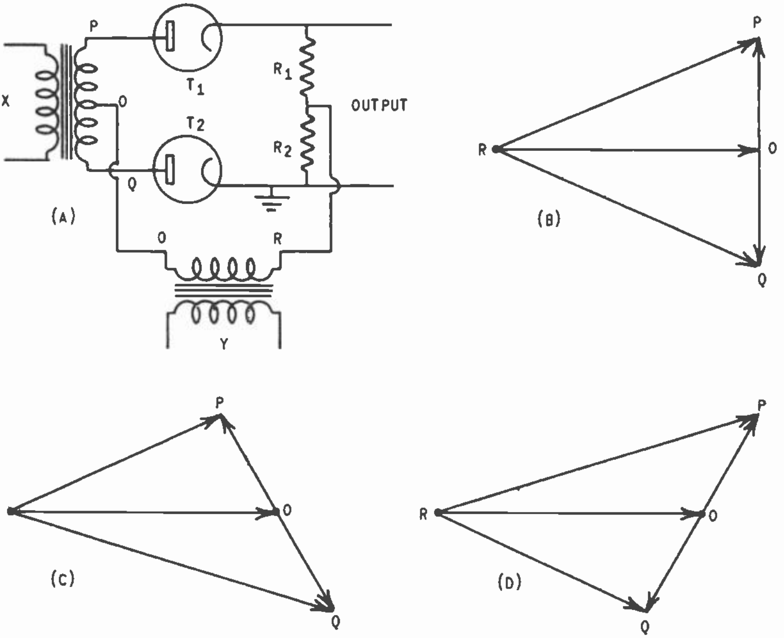


Fig. 24 Discriminator circuit.

Now suppose that for some reason the master oscillator was speeded up so that the voltage appearing at transformer X is advanced slightly in phase with respect to the control voltage at point Y. It may be seen from Fig. 24C that the voltage RP is now smaller than the voltage RQ. In other words, the output of rectifier T_1 has been reduced while that of T_2 is increased, and the drop across R_2 becomes greater than that across R_1. Because of the way that the rectifier tubes are connected, the drop across R_2 is negative and the drop across R_1 is positive. Since the negative voltage predominates, the output of the discriminator circuit will be negative and, as previously shown, a negative voltage operating the control tube will reduce the operating frequency and thus compensate for the drift which was originally assumed. If the oscillator tends to slow down, the vector diagram of Fig. 24D applies and the output of the discriminator becomes positive, increasing the frequency of the master oscillator.

The student should note that if the connections PQ are reversed, the discriminator functions in such a way that any deviation of the master oscillator is further increased and the frequency becomes very unstable.

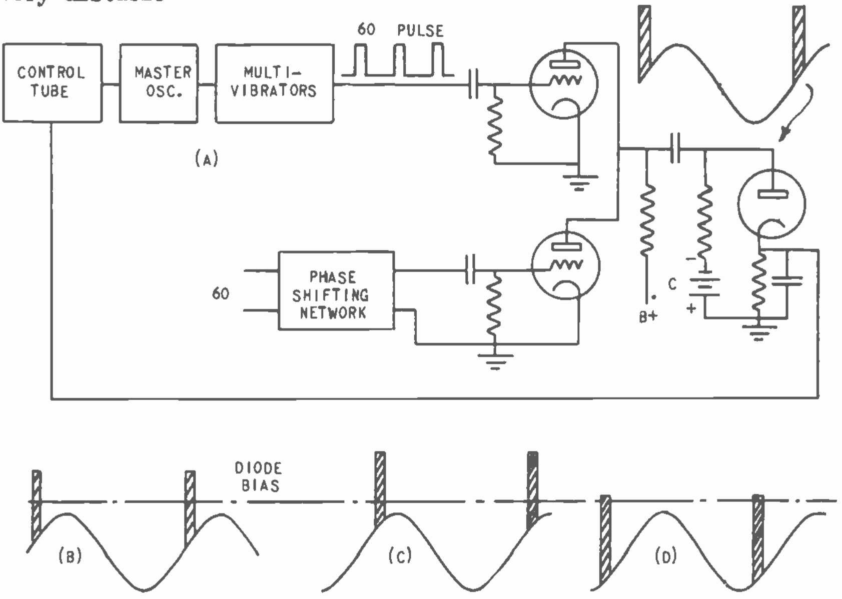


Fig.25 Circuit for obtaining control voltage to operate AFC circuit.

Another means of obtaining a DC control voltage from changes in oscillator frequency is shown in Fig. 25. In this case, the output of the multivibrator is applied to one tube and the 60 cycle comparison frequency applied to a second tube whose plate circuit is in parallel with the first. Thus, the pulse (from the multivibrator) is superimposed upon the sine wave and the resultant wave shown in Figs. 25B, C, and D is rectified in a diode, filtered by a condenser, and applied to the control tube. The diode is biased to such an extent that if the sine wave alone is impressed upon it, no current will flow and nothing will be supplied to the filter in its output. The combination of the pulse and sine wave will produce a wave of sufficient amplitude to overcome the delay bias and produce rectification as indicated by the broken bias line of Fig. 24B, C, D. For normal conditions, the pulse will maintain a position approximately as shown in Fig. 24B. If the master oscillator tends to slow down, the pulse is retarded in phase and rides higher on the sine wave as shown in Fig. 25C. Since the diode is biased to clip at a predetermined level, the output of the diode will be greater if the master oscillator is slowed down. By obtaining the control voltage from the cathode circuit of the diode, it will be positive when the master oscillator slows down, and as previously explained, a positive voltage will operate the control tube in such a manner as to increase the frequency of the master oscillator. The condition for the oscillator drifting to a higher frequency is shown in Fig. 25D. In this case, the output from the diode is much smaller and the master oscillator frequency is reduced.

Aphase shifting network was indicated in Figs. 22 and 25. The purpose of this network is to shift the phase of the comparison voltage so that the vertical blanking pulses will occur at any predetermined point on each cycle of the 60 cycle supply. You will learn in a subsequent lesson that with the intermittent type of motion picture projection for television, the picture is placed upon the Iconoscope for only a very small portion of the time. It is necessary to make this interval occur during the vertical blanking time and it is easier to shift the period of the blanking pulse than it is to shift the phase of the motion picture projector.

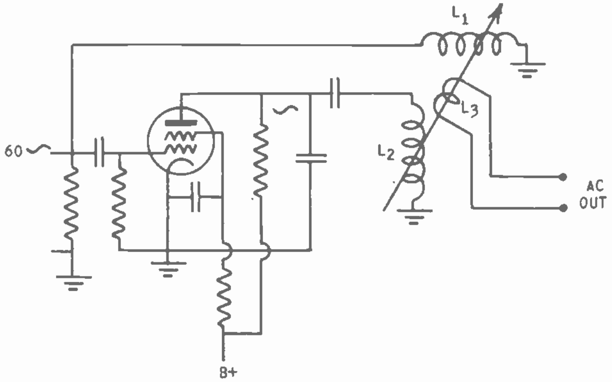


Fig.26 Phase shifting network.

One satisfactory means of accomplishing the phase shift is shown in Fig. 26. The inductances L_1 and L_2 are placed at right angles to each other and are wound to have a relatively high impedance at 60 cycles. The 60 cycle input is supplied to a pentode which has a capacity load in the plate circuit. Due to the high plate resistance of the pentode and the capacity load in its plate circuit, the output voltage is shifted 90° in phase from the input voltage. The output voltage is applied to one coil (L_2) and the input voltage is applied to L_1. Thus we have two coils at right angles to each other, carrying currents 90° out of phase. The student learned in a previous lesson that this condition produces a rotating field. By inserting an exploring coil L_3 into this rotating field, an AC voltage is produced. The phase of this voltage can be changed at will by varying the position of the exploring coil. Such a device is called a goniometer.

In the somewhat simpler and more popular circuit of Fig. 27, a centertapped transformer is used in conjunction with a capacity resistance phase splitting network. The voltage OX is 180° out of phase with the voltage OY. Also, the voltage drop across the condenser is 90° out of phase with the voltage drop across the resistor. The vector sum of the condenser and resistor drops equals the voltage XY. Since PX is displaced 90° from PY, and PX + PY = XY, it is possible to form a right angle triangle as shown in Fig. 27B by using vectors to represent the voltage drops across the various components. Fig. 27C shows that the vector OP may be inserted to represent the output voltage.

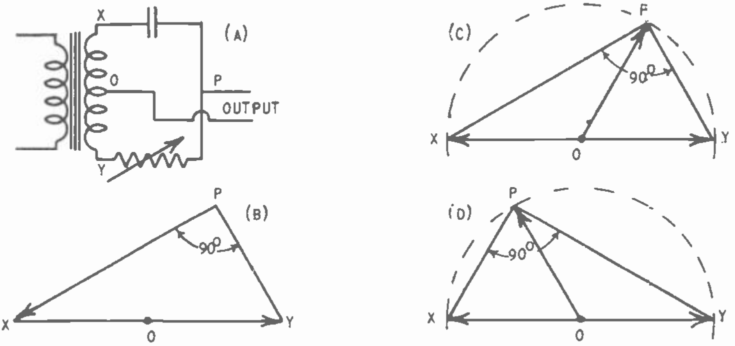


Fig.27 Resistance-capacity phase shifting network.

Since the voltage XY is constant, the sum of the voltage drops through the condenser and the resistor will also be constant. If the size of the resistor is changed, the only effect is to vary the *proportion* of the voltage drop across the condenser and the resistor. Furthermore, these two voltage drops are always 90° out of phase, regardless of their relative magnitudes. With various size resistors as shown in Figs. 27C and 27D, it will be found that point P always lies on the circumference of a circle whose diameter is XY. Since the radius of a circle is the same regardless of its direction, the output voltage, OP, will have a constant magnitude and a phase which is variable through 180° as the resistor is rotated from a short circuit to an open circuit.

6. SHAPING AND DELAYING CIRCUITS; SQUARING THE PULSE. The wave shaping unit of a synchronizing circuit is called upon to perform three major functions. First, it must be able to square a wave to have the desired wave front or slope as shown in Fig. 1. Second, it must vary the width of a pulse to the desired amount, either by making it wider or narrower. Third, it must be able to delay a pulse

by the proper time interval. It is not possible to speed up a pulse since, obviously, the pulse could not arrive at the end of a network before it started out.

The usual method of forming a square wave is to pass a pulse or wave through an amplifier having a limited available grid swing. By overswinging the tube in either direction, the wave is effectively chopped off. Two circuits are available for doing this, and both of these have been described before for a different purpose.

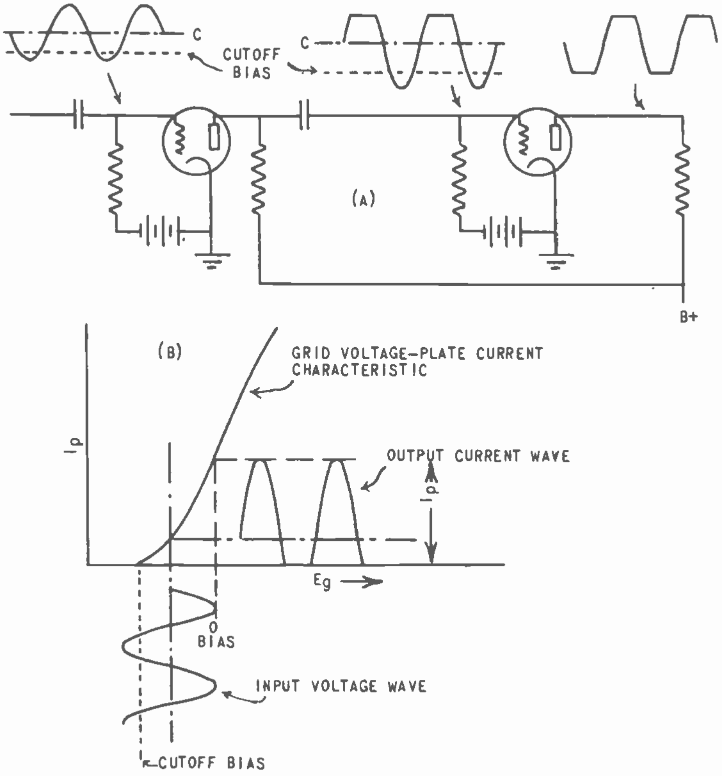


Fig.28 Formation of square wave by limited negative grid swing.

If a tube is biased as shown in Fig. 28, and is supplied with a large voltage swing; only a portion of this swing will appear in the plate circuit as shown. If the output of this stage is applied to a second stage, the reverse side of the wave will be clipped and a square wave having sloping sides will be created. Thus, it is possible to create a square wave from a sine wave in a double triode. In comparing Figs. 28A and 28B, the student should realize that the output wave of Fig. 28B represents the plate current pulses; while the waveform shown in Fig. 28A represents the plate voltage waveform. As in any resistance-coupled amplifier, the two are 180° out of phase.

The process shown in Fig. 28 need not be limited to one stage, and if several *clipper stages* are employed, the slope of the resultant square wave becomes steeper and steeper. This effect may be seen in Fig. 29. Figs. 29A through Fig. 29F represent the voltage waveforms on the grids of successive clipper tubes. It is obvious that each tube functions not only as a clipper, but also as an amplifier and, since the wave applied to the grids of successive stages becomes larger (until the grids are driven positive), only a small portion of the wave can be accepted by the stage. As the wave becomes steeper, the percentage of harmonics contained in it increases and the latter clipper stages must have low values of plate resistors in order to preserve a uniform response to the fundamental and all of the harmonics contained in the square wave. If an exceedingly steep waveform is desired, it may even be necessary to employ shunt peaking or some other form of high frequency correction as in video amplifier practice.

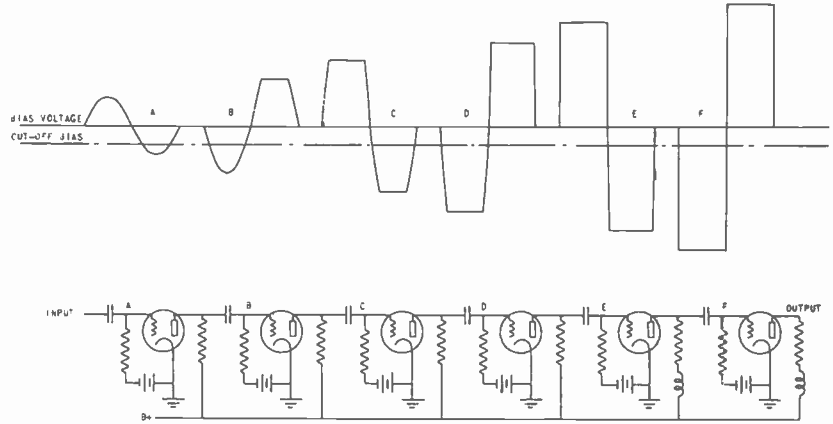


Fig.29 Formation of square wave by successive cutoff.

A large number of stages were shown in Fig. 29 to illustrate the principle of controlling the slope of the square wave. This method would not be economically practical, and more than two stages are seldom used. In this case, the desired steepness is obtained by starting out with a very large sine wave and restricting the swing of the amplifier tube by lowering the plate voltage. This causes the cutoff bias to occur at a much lower value and consequently only a smaller portion of the input wave is accepted by the amplifier.

Battery bias was shown in Fig. 29, but the common practice is to let the amplifier tube develop its own bias by the *process of leveling*. This process is similar to the DC restoration circuit studied in a previous lesson, except that in this case the grid of the amplifier tube itself, rather than a separate diode, performs the leveling action. If the amplifier tubes of Fig. 30 are operated initially without bias and the grid resistors are chosen to have a value between 100,000 ohms and several megohms, a bias charge will be built up across the coupling condenser C_1. Due to the high value of the grid resistor, the coupling condenser C_1 can have a low capacity without upsetting the low frequency response of the amplifier.

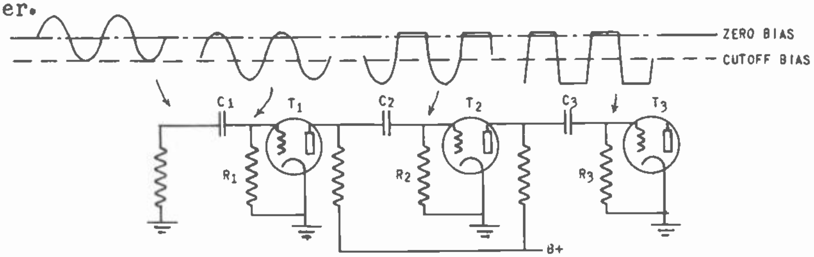


Fig.30 Application of leveling to wave shaping circuits.

During the positive half of the grid swing, the grid will draw current, which will cause it to have a low grid-cathode or grid to ground impedance. Since this impedance is low compared to the reactance of the coupling condenser, the grid cannot be swung very far in the positive direction and most of the drop will occur across the coupling condenser. On the negative half of the cycle, a low impedance no longer exists between grid and ground. Since the coupling condenser has a low impedance compared to the grid resistor R_1 in Fig. 30, practically all of the drop will occur across the grid resistor. Thus the coupling condenser has accumulated a charge from grid current flow which it maintains, diminished by only a small amount during the cycle, due to the energy leaked off by the grid resistor. A small sawtooth variation in the effective bias is generated, but this is of no consequence. Even if the reactance

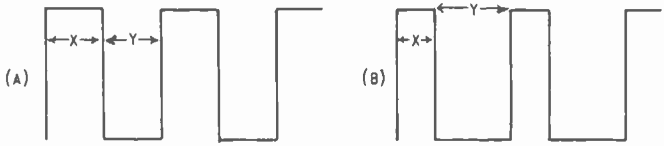


Fig.31 Symmetrical and unsymmetrical square waves.

of C_1 is quite low, it will still charge up to bias the tube T_1 in such a way that the peak value of the wave applied to its grid will just equal the zero bias. This condition will not take place during the first several cycles; but since the charge path for the condenser is always much lower resistance than the discharge path R_1 the condenser will eventually reach the correct charge to set the bias on tube T_1 at the necessary value.

Each successive stage limits the peak of the swing to zero bias and the axis of the wave is shifted in a negative direction so that most of the wave is depressed below the cutoff bias level. With this type of circuit, the gain per stage and the value of the initial input wave must be carefully chosen if the square wave is to be symmetrical. A symmetrical square wave is illustrated in Fig. 31A and is distinguished by the fact that the duration X and the duration Y are equal. An unsymmetrical square wave takes the form of a pulse as shown in Fig. 31B, having unequal values of X and Y. In synchronizing generators, a symmetrically square wave is never needed and consequently the components of the circuit shown in Fig. 30 need not be critical.

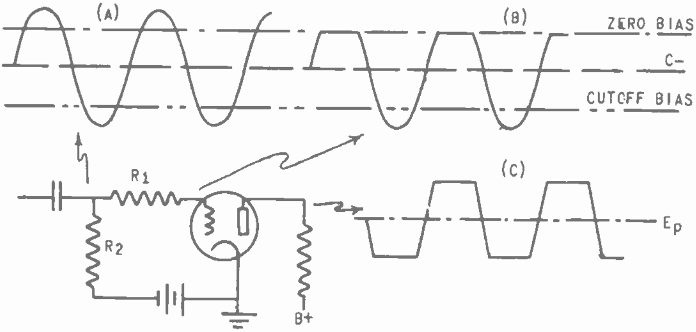


Fig.32 Formation of square wave by limited positive grid swing.

The second basic method of forming a square wave is shown in Fig. 32. In this case the positive half of the wave, rather than the negative half is trimmed. This is done by inserting a series resistor between the input terminal and the grid of the tube. This resistor has a high impedance compared to the grid-cathode impedance of the amplifier tube when the grid is positive, and most of the positive cycle will occur as a drop across R₁. Therefore, practically none of the positive portion of the cycle will be impressed on the grid of the amplifier tube. This condition is shown in Fig. 32B. If in addition the cutoff bias is properly chosen, the negative half of the cycle will also be trimmed in the amplifier tube and a square wave will be formed in the single tube as shown in Fig. 32C. By employing partial leveling, the bias battery may be omitted from this circuit.

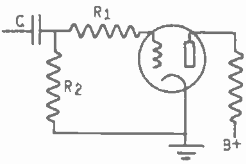


Fig.33 Wave shaping circuit employing leveling and double clipping.

The simplified circuit is shown in Fig. 33. If the time constant CR_2 is chosen so that an appreciable amount of the charge on the coupling condenser will be leaked off between cycles, then part of the positive swing will appear as a drop across R_1. Nevertheless, the charging path for the condenser is the combination R_1 and R_2, while the discharge path is simply the value R_2. This is due to the fact that the tube is drawing grid current on the charge portion of the cycle, but has an infinite grid impedance on the negative portion of the cycle. Since the charge and discharge paths for the coupling condenser are unequal, it will accumulate a bias which can be made the desired value for the amplifier tube by properly choosing the values C, R_1, and R_2. Since the values chosen depend upon the amplitude of the input signal, they are generally selected by a cut and try process.

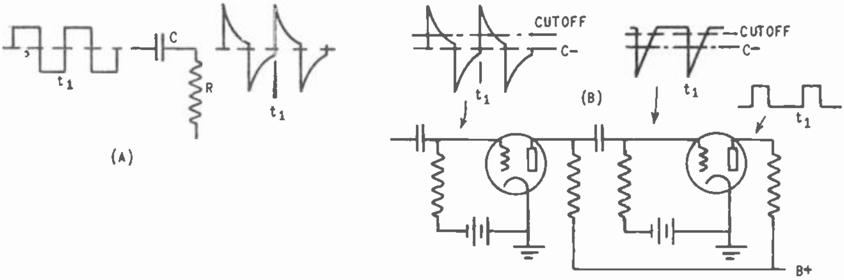


Fig.34 Decreasing pulse width by differentiation.

7. CONTROLLING THE PULSE WIDTH. In synchronizing generators, it is necessary not only to form square waves, but to vary the relative duration of alternate half-cycles as was illustrated in Fig.

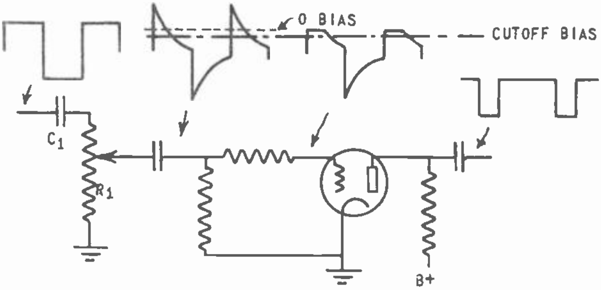


Fig.35 Simplified circuit for decreasing pulse width.

31. This is another way of saying that the duration of the pulses must be controlled. You have already learned in this and preceding lessons that the duration of a pulse from a blocking tube oscillator or a multivibrator may be controlled by suitable choice of the circuit components. In some synchronizing generators, multivibrators and blocking tube oscillators are used to generate pulses of varying widths by synchronizing them from either a symmetrical square wave, a narrow pulse, or occasionally from a sine wave. Some

engineers do not consider it good practice to generate a standard synchronizing wave from circuits which employ synchronizing in themselves, because this increases the possibilities of inaccuracy in timing or of an oscillator failing to synchronize altogether. Most synchronizing generators start out by generating a square wave with correct frequency in a timing unit and then developing the correct duration of the pulse in non-oscillating circuits in the shaping unit.

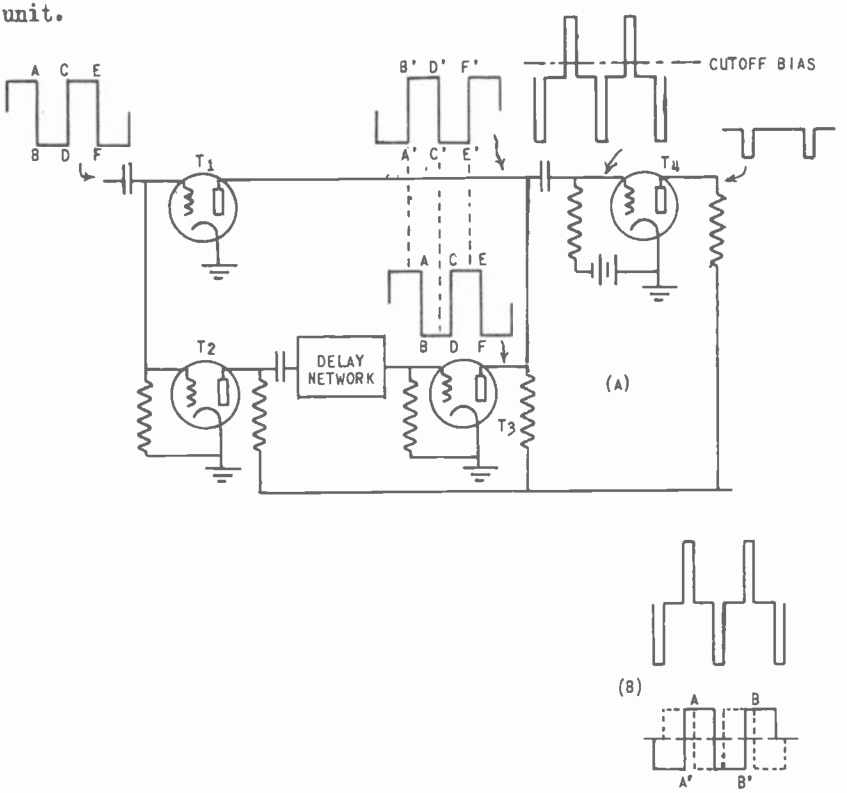


Fig.36 Controlling pulse width balancing positive and negative pulses.

It is more often necessary to narrow a pulse than to widen it. The best way of doing this is to pass the wave through a high-pass filter to remove the low frequency components, then pick off a portion of the wave, and square it up again. The first step in the process is shown in Fig. 34A, depicting the resultant waveform in passing a square wave through a capacity-resistance network having a small time constant. The second step in the process, shown in Fig. 34B, indicates how a double triode with both sections biased beyond cutoff may be used to pick out a small portion of the distorted wave and square it up, resulting in a pulse that is narrow, compared to the original pulse, but which has a leading edge beginning at exactly the same time as the original square wave.

It is possible to perform this entire function in one tube and its associated components as shown in Fig. 35. This is a combination of several circuits which we have just studied. The wave distortion or differentiation produced by C_1R_1 of Fig. 35 is the same process which was discussed under Fig. 34A. The method of squaring the pulse in the amplifier stage was discussed under Figs. 32 and 33.

Another method of forming a symmetrical square wave into a narrow pulse is to combine two identical square waves which are almost but not quite 180° out of phase. When this is done, the waves will cancel out during most of the time, but will add over a small portion of the cycle. This is best understood by referring to Fig. 36. The amplifier tube T_4 is fed by the combination of two waves in parallel. These two waves are derived from the square wave through two amplifiers, one having a single stage and the other having two stages. The extra stage in one of the chains serves to invert the polarity of the wave by 180°. Furthermore, the two stage circuit has a network which delays the square wave very slightly. How this is accomplished will be discussed in the following section. Fig. 36B shows the effect of combining the waves from the two amplifiers. The output of the two stage amplifier is shown by the dotted wave. The output of the other amplifier is shown by the solid curve. During most of the time these two waves are opposite in phase and cancel out. During a small interval, such as between A and A', the waves are in phase and add as shown. The amplifier tube T_4 serves to accept only the positive cycle, creating a narrow pulse which is controllable by the amount of delay introduced in the two stage amplifier.

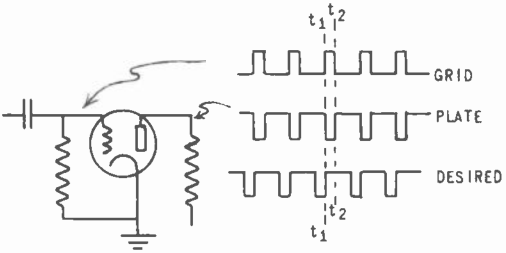


Fig.37 Simple circuit which is not suitable for forming serrated pulses from equalizing pulses.

For certain applications it is necessary to increase the width of a pulse. This may be done, for instance, to form a serrated vertical wave from a 26 kilocycle square wave pulse. If a narrow wave of positive polarity is passed through an amplifier stage, it is inverted, but it may also be considered as a positive wave with much longer duration. Unfortunately, the leading edge of the resulting wide pulse does not occur at the same time as the leading edge of the narrow pulse which fed the amplifier stage. This is illustrated in Fig. 37.

A wide pulse can be made from a narrow one without introducing a great deal of time shift, by incorporating a low-pass filter in the grid circuit of an amplifier which is adjusted to cutoff. Fig. 38 indicates a typical circuit applying this method. The low-pass filter formed by RC creates a distorted sawtooth from the narrow pulse which drives it. The first amplifier passes practically all of the sawtooth as shown by the broken line representing cutoff bias. The second stage, which receives the inverted and amplified wave, cuts off most of the sawtooth and passes a wave which is essentially square; beginning at almost the same time as the original pulse, but of longer duration.

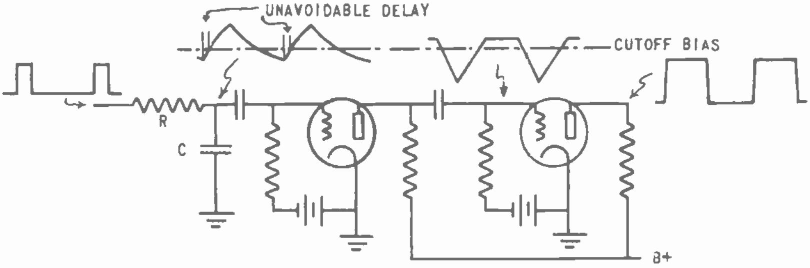


Fig.38 Circuit for increasing pulse width.

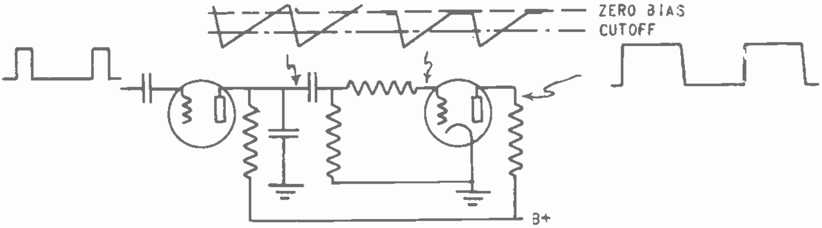


Fig.39 Microtome circuit for selecting proper pulse width.

Another method of widening a narrow pulse is shown in Fig. 39. In this case, the pulse is used to drive an ordinary discharge circuit, generating a sawtooth. A portion is sliced out of the sawtooth to form the square wave. This is done by a single tube using the two methods for squaring previously described.

8. **DELAYING THE PULSE.** Under the heading of Electronic Generators, it was seen that it is often necessary to delay a pulse either during or after its formation. The circuit shown in Fig. 38, with slight modifications, may be used to delay a square wave. This circuit is generally used for the double purpose of obtaining the right width of square wave and the correct amount of delay. All

that is necessary is to shift the bias of the first tube shown in Fig. 38 in order to control the amount of delay and the width of the square wave generated. Fig. 40 shows more clearly the operation of the circuit. If the bias on the first amplifier tube is increased the result is to increase the delay and reduce the width of the resulting pulse.

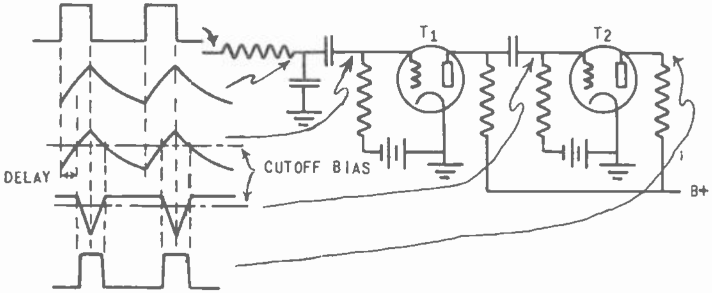


Fig.40 Circuit for delaying rectangular pulses.

In the type of synchronizing generators which form the pulses by triggering blocking tube oscillators or multivibrators, it is only necessary to delay the synchronizing pulse which is usually a form of a sharp transient. One method of securing a delayed transient to fire the oscillator is to excite a highly damped tuned circuit, which will generate a damped oscillation. By suitably arranging the polarity of the synchronizing wave, it is possible to pick off a portion of the transient which is delayed from the initial pulse that caused it. Fig. 41 shows how this may be done.

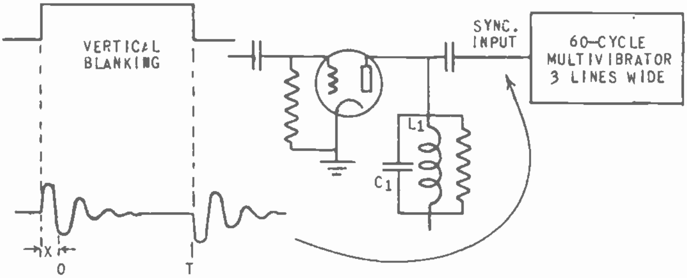


Fig.41 Circuit for controlling the firing time of a relaxation oscillator.

Suppose we wish to form the vertical synchronizing pulse which is delayed from the start of the vertical blanking pulse by three horizontal lines. The vertical blanking pulse is applied to the grid of the amplifier tube in Fig. 41. The tuned circuit connected in its plate circuit generates a damped oscillation, and a 60 cycle multivibrator which this oscillation synchronizes is arranged to fire only on a negative pulse. If we want a certain amount of delay, equal to X, then the tuned circuit L_1C_1 is chosen to resonate at such a frequency that the time X represents three-fourths of a complete cycle. The multivibrator will fire at time 0 on the graph of Fig. 41. It will not fire at time T, which is also a negative pulse, because this occurs soon after the multivibrator has just fired, and the conditions in its discharge circuit are such as to prevent its firing again unless a pulse of extremely large magnitude is applied.

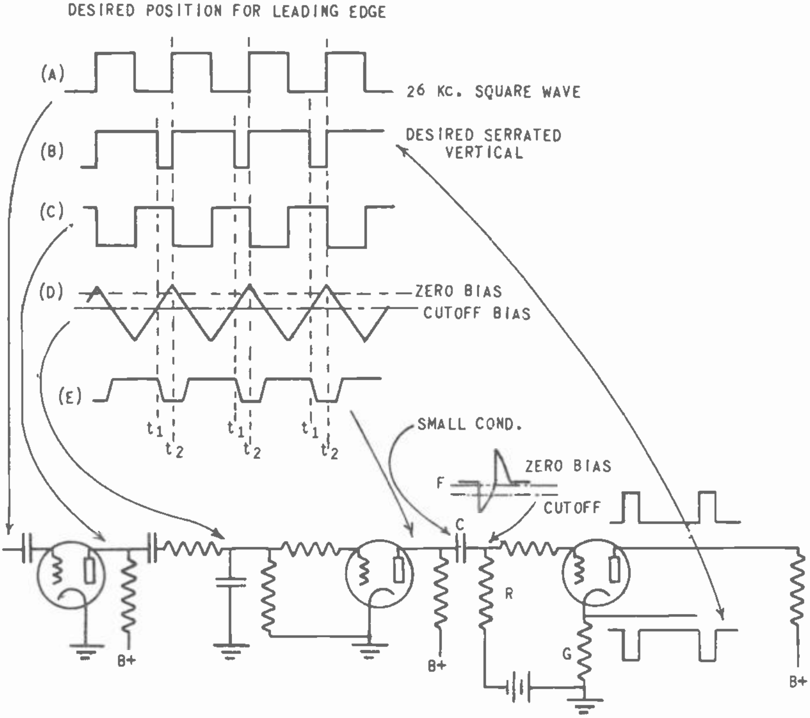


Fig.42 Circuit for the formation of serrated pulses.

In transporting the signal from one part of the chain to another over long lengths of coaxial line or through networks in various sections of the studio equipment, an unavoidable delay is occasionally introduced. When this condition exists, it is desirable to advance the phase of the pulse as it is generated. The necessity for advancing a pulse likewise occurs in the formation of the serrated vertical by widening the equalizing pulses. The circuits of Figs. 38 and 39 introduce a slight delay in the leading edge of the wide pulses which were formed from the narrow input pulses.

As mentioned previously, it is not possible to advance a pulse and still maintain a steep wavefront. The delaying of a pulse by a time equal to 95% of one full line is the same thing as advancing the following pulse by 5% of one line. This would be difficult to do in practice, but it is entirely practical to invert the phase of a symmetrical square wave, pick off the opposite half cycle, and delay it almost half the distance between pulses. This is illustrated in Fig. 42. The original square wave and the desired serrated vertical wave are shown together in Figs. 42A and B. Curve C is the square wave after it has gone through the first amplifier tube. The triangular wave at D is generated from this square wave by passing it through the low-pass filter connected in the grid circuit of the second stage. The second stage is adjusted to trim the wave at zero bias and cutoff bias, resulting in a plate circuit wave shown as curve E. This is passed through the high-pass filter RC which distorts the square wave at F. The third amplifier tube is also adjusted to trim between zero bias and cutoff bias, thereby producing in its plate and cathode circuit a narrow pulse whose leading edge occurs at time t_1. The trailing edge occurs at time t_2. These pulses which appear in the cathode circuit as curve G are easily recognized as being identical with the desired serrated vertical drawn as curve B.

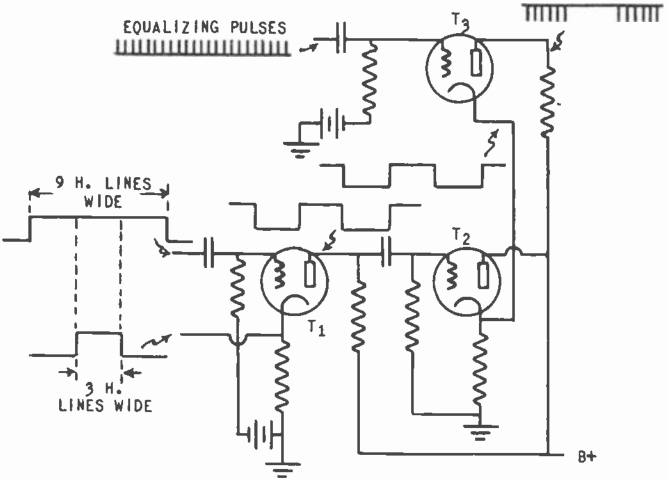


Fig.43 Detailed keying circuit.

9. **MIXING AND OUTPUT CIRCUITS.** In Sections 1 and 2 of this lesson, the functions of the various parts of the electronic generator were described together with the reasons for generating the various types of waves. In Sections 3-8, as well as in this section, detailed circuits were discussed for performing the individual functions. Before reading through this section on mixing and

output circuits, the student should turn to Figs. 1, 7, 8, 9, and 10, for a brief review of the purpose and function of the mixing and output circuits. In Fig. 10, for example, square waves of various durations and occurring at various times are used to key in and out the various waveforms that make up the supersync.

A keying circuit is essentially a modulating stage which modulates between full-on and full-off for a certain interval of time. This is done by applying the wave to be keyed to one element of a vacuum tube and applying the keying pulses to another element. Triodes or tetrodes may be used, and the keying pulses may be applied to either the control grid, screen grid, or cathode of the tube, depending upon whether positive or negative polarity is desired.

Fig. 43 shows the circuit used for keying in the equalizing pulses which occur before and after the vertical synchronizing pulses. The circuit is seen to consist of three triodes which perform the double keying function described under Fig. 10. A three line wide pulse is used to key a hole in the middle of a 9 line pulse. The resultant waveform is used to key in the equalizing pulses in tube T_3. The waveforms shown in Fig. 43 do not have the same polarity as those of Fig. 10. The reason for this is that the tubes in the schematic diagram of Fig. 43 invert the polarity as in any amplifier when a signal is impressed on the grid and taken out at the plate. All waveforms were shown in Fig. 10 as if they were positive, since this simplified the explanation of the block diagram and no actual circuits were indicated.

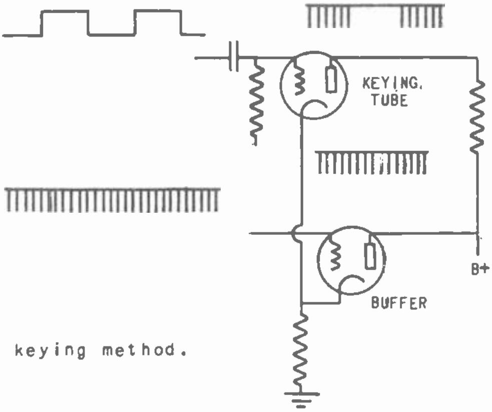


Fig.44 Alternative keying method.

Returning to Fig. 43, tube T_2 serves as a buffer and is not strictly necessary. Since it is cathode loaded, the polarity is not inverted through this stage. Tube T_1 is normally biased to cutoff, but is driven to considerable plate current by the 9 line wide pulse. During part of this pulse, the plate current is again driven to zero by a positive pulse introduced into the cathode of sufficient magnitude to cut the tube off. The resultant waveform is shown opposite the plate of this tube and is applied through the buffer to the cathode of tube T_3. Tube T_3 is also biased to cutoff and conducts only when its cathode is driven negative by the buffer tube T_2. Thus, tube T_3 passes the equalizing pulses only during the interval shown.

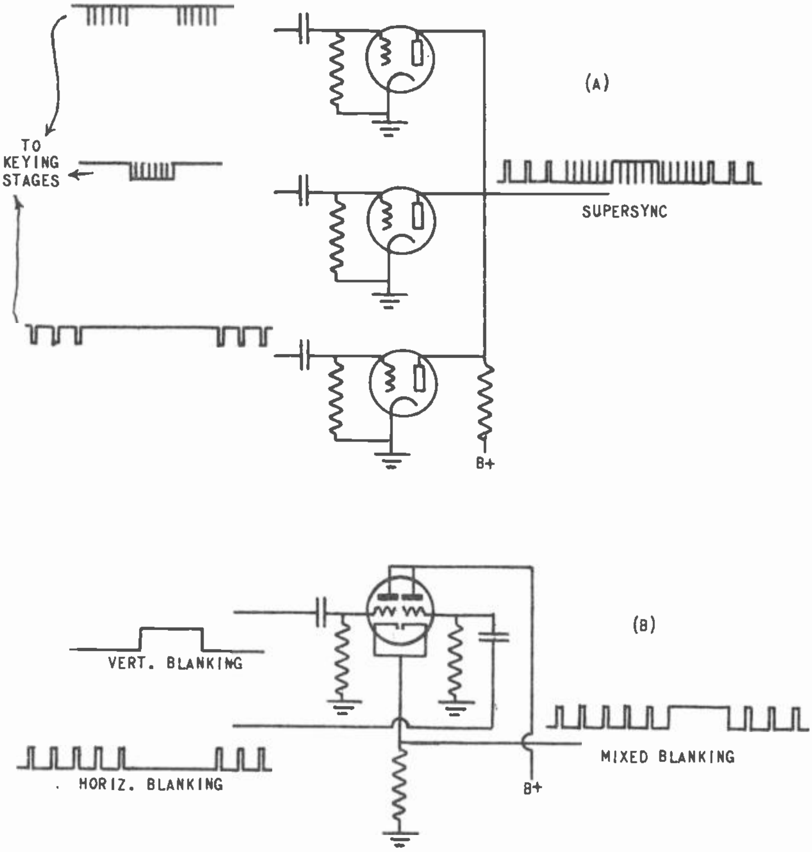


Fig.45 Signal mixing circuits.

All of the keying circuits used are variations of this fundamental type. One typical variation is to place the keying pulse on the control grid and the equalizing pulses on the cathode, as shown in Fig. 44. This simply reverses the polarity of the input wave and may be used to suit the convenience of the designer. A double triode with a single cathode could be used conveniently in this circuit.

The mixing circuits in the synchronizing generator necessary to secure the supersync and the combined blanking signals are simply amplifier tubes whose outputs are in parallel, but which have individual input terminals. The common load in the output circuit may be in either the plate circuit or the cathode circuit as shown in Fig. 45. There is very little difference between the function of a keying stage and a mixing stage, and a keying stage can be arranged to perform the function of a mixer. Fig. 46 shows how this may be done in the case of a stage which mixes the horizontal and vertical blanking. In this stage, both of the blanking pulses occur in the direction of cutoff. The tube is arranged so that either the horizontal or vertical blanking pulses drive the stage to cutoff; consequently, the two signals may be mixed without having the horizontal impulses superimposed on top of the vertical pulses during the time of the vertical blanking out.

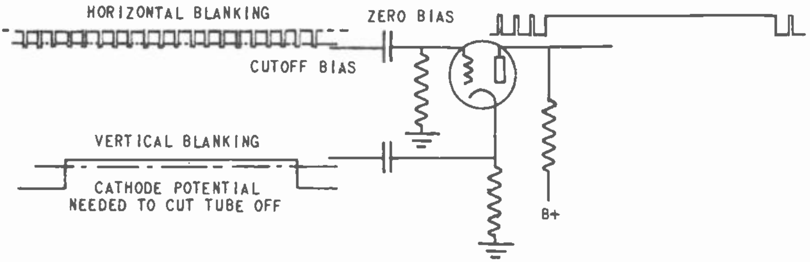


Fig.46 Alternative mixing circuit.

Turning back to Fig. 8, you will notice that five different types of output signals are required and that each output waveform has several output terminals. All of these output terminals preferably include buffers between the shaping unit and the connections to the various concentric lines, or shielded wires, which feed various portions of the studio equipment. These buffers are preferably cathode coupled, since this method reduces the output impedance of the buffer stage, and also lessens the effects of power supply "bounce" and hum in upsetting the wave shape. Where either a positive or negative signal may be desired, a load resistor can also be included in the plate circuit of the output stage. Figs. 47A and 47B show arrangements for cathode coupled output stages, and for cathode and plate loaded output stages. The advantages which accrue from cathode loading an amplifier will be discussed under a separate topic heading of this lesson.

The arrangement necessary for securing the various output waveforms were shown in the block diagram of Fig. 8. More detailed circuits necessary for shaping and delaying the pulses were discussed in the preceding two sections of the lesson. Fig. 48 shows the desired shapes of the various output pulses. There is of course considerable leeway permissible in the duration of the Iconoscope blanking and the vertical and horizontal driving pulses.

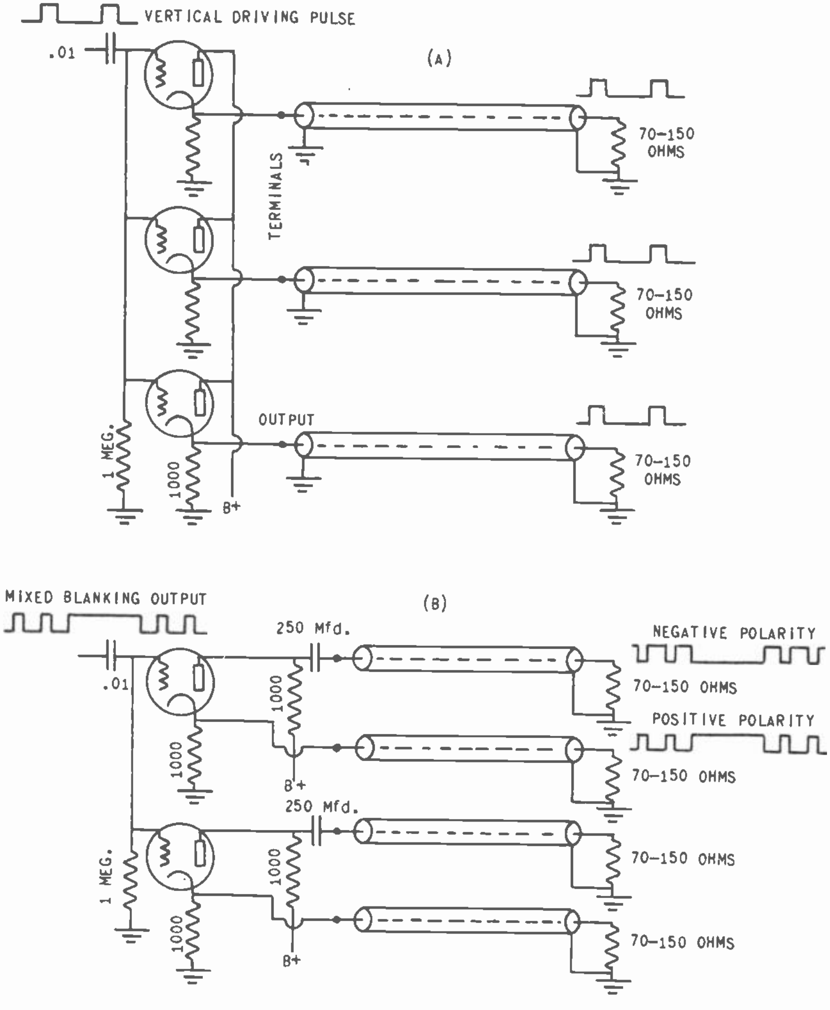


Fig.47 Synchronizing generator output circuits.

10. **MECHANICAL SYNCHRONIZING GENERATORS.** Mechanical synchronizing generators may be divided into two general classes. First, the generator whose purpose is to create the 26,460 or 13,230 cycle signal, these signals being supplied to electronic shaping circuits similar to those already described. Second, the electro-mechanical means of generating the entire RMA standard synchronizing signal and other special pulses.

In generating simple pulses at a frequency of 13,230 or 26,460 cycles, several methods are available. The oldest and most widely used method is to rotate a perforated disc between a light source and a photocell. The apertures in the disc transmit light in the form of discrete pulses which are picked up by the photocell and

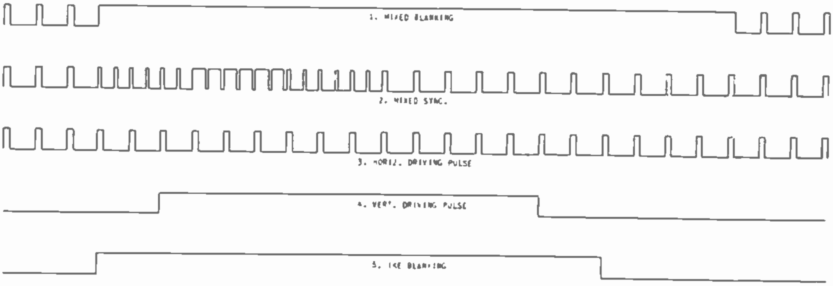


Fig.48 Relative timing and duration of synchronizing generator output pulses.

subsequently smoothed out by sharply tuned circuits in a vacuum tube amplifier. The perforated disc is driven with an 1800 rpm synchronous motor from the 60 cycle supply line. Consequently, the synchronizing disc will drift in accordance with changes in

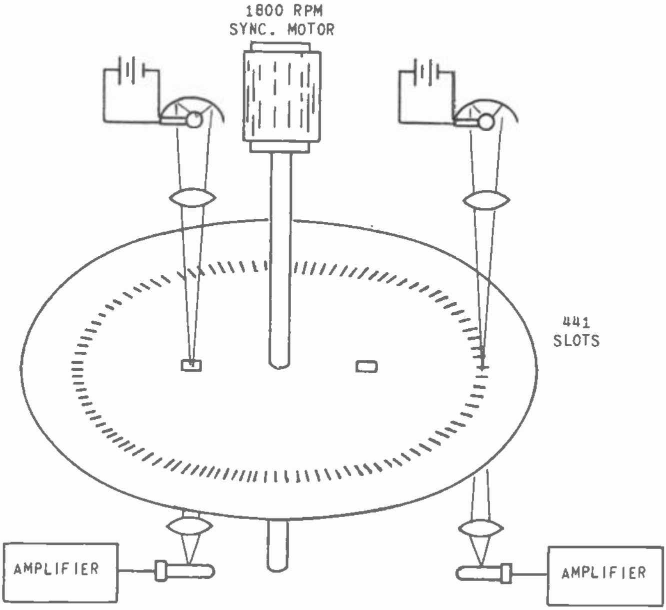


Fig.49 Slotted disc mechanical synchronizing generator.

phase of the 60 cycle line voltage. This is a desirable and necessary feature. Fig. 49 shows a sketch of such a synchronizing generator. Extreme precision is required in locating the slots, but they can be located with an optical dividing head, to the required accuracy. The slots not only must be accurately located, but must have identical widths, in order to maintain the timing and size of every synchronizing pulse exactly the same. Small particles of dust and dirt in the slots are sufficient to upset the quality of the synchronizing. However, inaccuracy between individual pulses can be ironed out as previously mentioned with sharply tuned, or even regenerative amplifiers. This system has the disadvantage that the high Q circuits will not allow the frequency of the horizontal pulses to change with slight changes in phase of the 60 cycle line voltage.

In addition to the horizontal synchronizing slots, two other slots are cut in the disc. An additional lamp house and photocell is used to project through this slot to secure the 60 cycle vertical synchronizing pulse. Other rows of slots may be cut in the disc to secure special pulses, such as the Iconoscope blanking pulse and the driving pulses for the deflection circuits.

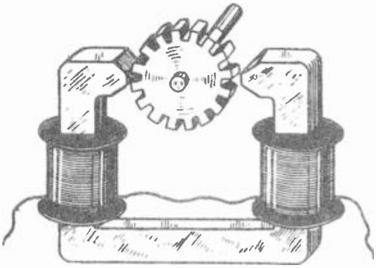


Fig.50 Phonic wheel.

Another type of synchronizing generator is an adaptation of the phonic wheel. The phonic wheel will be described in a later lesson in this unit, but a brief description will be given here. Fig. 50 is a drawing of a simple phonic wheel. If the armature is rotated between the poles of a magnet, the flux between the poles will be disturbed as each tooth of the phonic wheel passes in front of the pole. This disturbance in the magnetic field will induce, in the stator winding, current pulses which can be used for synchronizing in conjunction with a suitable amplifier and shaping circuit. Irregularities in the teeth or eccentricity in the drive would result in extremely poor synchronizing, but this can be reduced to some extent by increasing the number of poles on the stator. If a large number of poles are used, the irregularities in the teeth will tend to cancel out. The simple phonic wheel of Fig. 50 would become rather unwieldy when used for 441-line television, and since this method has been largely superseded by more satisfactory types of mechanical generators, it will not be discussed further.

A mechanical generator making use of the disturbances in an electrostatic field, as contrasted to the magnetic disturbances in the phonic wheel, is shown in Fig. 51. This generator was developed by the Scophony Company for operation on 405-line English television, but is entirely adaptable to 441-line systems. It consists of a toothed disc which rotates between two toothed stator rings. The stator rings form two plates of a condenser and as the rotor teeth pass between the stator teeth, the capacity of the condenser changes. This condenser may be used either in a tuned circuit to supply output pulses, or may be operated aperiodically by placing a large charge on the condenser and making use of the changes in current which flow in a load circuit when the capacity of the condenser is changed. The operation of the circuit is quite similar to the now obsolete condenser microphone. One practical circuit for making use of the pulses generated by the Scophony synchronizing generator is shown in Fig. 52

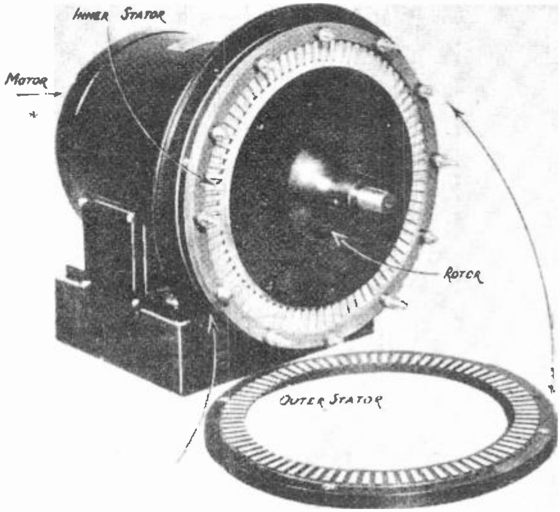


Fig.51 Scophony synchronizing wheel.

Of the several types of mechanical generators just described, the Scophony system furnishes the highest quality synchronizing pulses. One reason for this is the large number of teeth which are acting in parallel simultaneously. Thus, any irregularity is averaged out in the resultant output. Considerable eccentricity of the driving shaft can be tolerated as well as misalignment of the rotating disc. The mechanical inertia of such a system prevents rapid fluctuations in the horizontal synchronizing pulses. This is highly desirable, if not entirely essential, to the correct operation of a mechanical receiver such as the Scophony projection receiver.

An electronic generator of the type previously described tends to fluctuate slightly above and below the mean frequency at a rate of approximately 100 to 300 cycles per second. The control circuit of the electronic generator swings first to one side and then the other of the mean carrier frequency. The oscillations are quite small in magnitude, but are sufficient to upset the performance of a mechanical receiver. This is of little consequence in an electronic type of receiver since the deflection circuits can easily follow the very small changes in frequency.

As previously stated, all of the aforementioned mechanical generators still depend upon electronic means to develop the complicated wave shapes necessary for modern television. Since the shaping and output circuits form the bulk of an electronic generator, there is little to be gained from a mechanical generator except the stabilizing influence of the mechanical inertia.

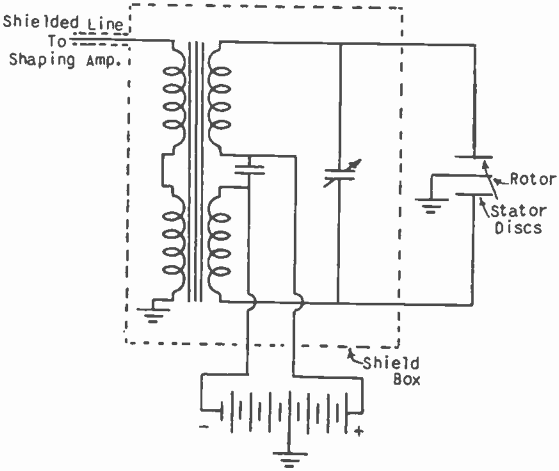


Fig.52 Circuit used in conjunction with Scophony wheel.

Optical systems, such as the one first described in this section, may be made to produce the entire group of complicated waveforms necessary for RMA standard synchronizing pulses. In this case, however, it would be impossible to utilize the smoothing circuits which are sometimes used with mechanical generators. If the entire group of standard pulses are to be incorporated upon the synchronizing disc, it is preferable to print them upon the disc by photographic reduction from a large master plate. Reflected light from the disc can be used instead of the light transmitted through narrow slots. The mechanical tolerances required might well prove prohibitive in such a system although it has not been thoroughly investigated to date.

Suppose that a master layout sheet six feet in diameter were used. The synchronizing pulses would be laid out on a circumference having a total length of almost 19 feet, or 216 inches. Even on this large a scale, a horizontal synchronizing pulse would be indicated by a line only slightly more than 1/32 inch in width (.039 inch). In order to prevent inaccuracies of the synchronizing in excess of one picture element, the synchronizing pulse would have to be located to within .0008 inch and each pulse would have to be identical to within this figure if 400-line resolution in the horizontal direction is taken as the standard. This tolerance is .001 inch, roughly, and while it would be practical to make such a master disc, it is doubtful if the same percentage of accuracy could be maintained in the photographic reduction for the synchronizing generator disc.

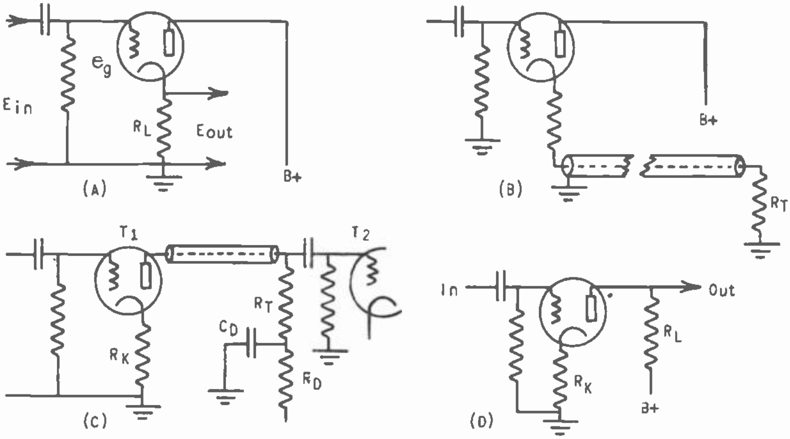


Fig.53 Applications of cathode degeneration.

11. CATHODE DEGENERATION. In this lesson and in other places throughout the course, circuits have been given with unbypassed cathode resistors when the load for the tube is in the plate circuit; in some instances the output was taken from the cathode circuit of the tube. Some interesting results are secured when an unbypassed cathode resistor is used, whether it is the only load for the tube or not. The advantages to be acquired by such operation may be listed as follows:

(1) Reduction in output impedance
(2) Improved linearity
(3) Reduction in input capacity
(4) Increase in input resistance
(5) Sharper cutoff characteristic
(6) More usable grid and plate swing

The above points will be considered in the order of listing. Consider Fig. 53A. A reduction in output impedance means that the cathode loaded stage acts to reduce the effect of changes in output impedance and changes in supply voltage. If the load is placed in the cathode circuit and some disturbance occurs in the output circuit, this disturbance will be minimized by the effect of cathode loading the tube. Suppose that the output voltage E_{out} tends to increase. This will cause the cathode of the amplifier to go positive, thereby reducing the cathode and plate current. The reduction in cathode current tends to decrease the drop through the load resistor R_L and this partially compensates for the tendency of the output voltage to rise. If the output voltage tends to diminish, the amplifier tube draws more current because of its reduced bias, and this tends to prevent the decrease in output voltage. The tube therefore has a stabilizing influence upon the output and acts like a low impedance source of voltage. The higher the transconductance of the tube, the greater will be this effect. The effective output impedance of the tube will appear in parallel with the load resistor. The value of this output impedance (not counting the load resistor which it parallels) is given by the following formula:

$$Z_{out} = \frac{1}{G_m}$$

An additional advantage of coupling from the cathode circuit is that the termination resistor of a concentric line may serve as the load resistor for the tube. If the value of the line termination is not the correct cathode resistance for the amplifier tube, more resistance can be added in series to make up the difference as shown in Fig. 53B. When the alternative method (shown in Fig. 53C) is used, and the transmission line fed from the plate circuit, the power supply impedance becomes comparable to the load resistor R_t. If the decoupling network C_dR_d is inserted to avoid the disturbing influence of the complex power supply impedance, the condenser C_d must have a very large capacity. This is due to the fact that R_tC_d must have a large time constant, and R_t is quite small (70-200 ohms).

In the circuit of Fig. 53C, any hum or voltage surges in the power supply are applied directly to the grid of T_2; while in the case represented by Fig. 53B, hum and line surges have little effect upon the output.

By incorporating an unbypassed cathode resistor in an amplifier, the linearity is improved considerably. Fig. 54 shows the grid voltage-plate current curve for an 1852 tube operated with a plate voltage of 300 volts and a load resistor of 1500 ohms. This is a typical value for video amplifiers. The left hand curve shows the performance of the tube with the cathode bypass condenser in place, while the right hand curve shows the considerable improvement in linearity which results from leaving the 1,000 ohm cathode resistor unbypassed.

The improvement is obtained at the expense of the tube's gain. This is shown by the different scales used for the input voltage, and it will be seen that it takes approximately ten times the input for this particular example, when the cathode resistor is unbypassed. The loss of gain would not be nearly as great if a normal value of cathode resistor (160 ohms) had been used, but in this case the improvement in linearity likewise would not be so great.

For values of load resistance which are low compared to the plate resistance of the tube, the gain of the stage shown in Fig. 53A will be given as follows:

$$\text{Gain} = R_L \div \left(R_L + \frac{1}{G_m}\right)$$

This value approaches unity for large values of load resistance, but never exceeds a gain of one.

The stage shown in Fig. 53D will have a gain as given by the following relation: (See next page)

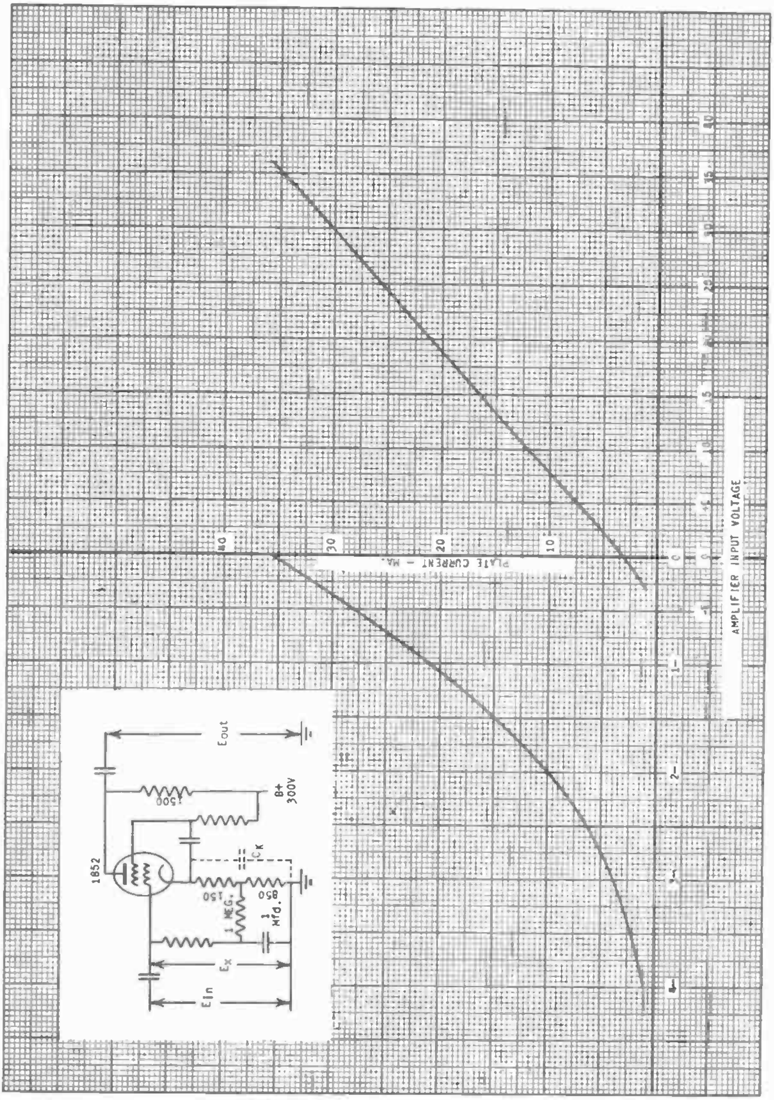


Fig.54 Effect of cathode degeneration upon linearity.

$$\text{Gain} = \frac{G_m R_L}{1 + G_m R_k}$$

The reduction in gain for a tube operated with an unbypassed cathode resistor may not be a very serious problem, due to the *reduction in input capacity* which accompanies this circuit arrangement. It is easy to obtain a mental picture of just how this reduction in input capacity occurs by considering Fig. 55. If there is a change in the potential between the grid and the other elements of the tube, electrostatic lines of force will exist between these elements and a capacity current will flow into or out of the grid. Suppose we represent a change of one volt as producing one electrostatic line of force. In Fig. 55A, an instant is considered just after the grid has swung in the positive direction by four volts.

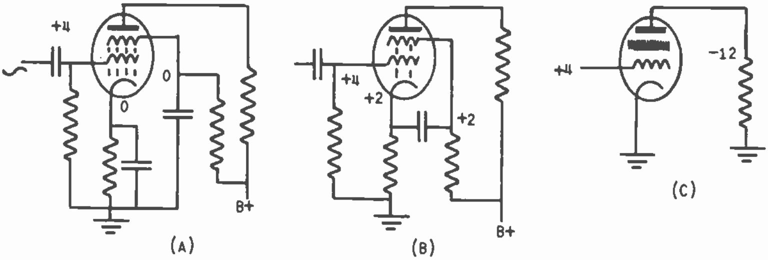


Fig.55 Effect of cathode degeneration upon input capacity.

It is seen that four lines of force exist between the grid and screen and between the grid and cathode. This represents a definite amount of capacity current which must flow into the grid to charge up this electrostatic capacity. In Fig. 55B, an unbypassed cathode resistor is employed, and the screen is bypassed to the cathode. In this case, as the grid is swung positive by four volts, a positive potential of two volts appears on the cathode. The screen is likewise changed by two volts. Thus, the change in potential between the grid and the other elements is only two volts, instead of four as previously. Fewer electrostatic lines of force will exist and less capacity current will be required to charge the interelectrode capacities. Since X = E/I, the capacitive reactance is smaller in the case of cathode loading than for a bypassed cathode resistor. For a given change in grid voltage, there has been a smaller capacity current and the effective input capacity of the tube is therefore less. In the case of a triode, only the grid-cathode capacity is reduced by this means; the grid-plate capacity still exists.

If, in Fig. 55B, the screen had been bypassed to ground instead of to the cathode, the control grid-screen grid capacity would not have been reduced, but only the grid-cathode capacity. The amount of input capacity reduction for pentode connection is given by the following formula:

$$\text{Input capacity reduction factor} = K = \frac{1}{1 + Z_k G_m}$$

The normal C_{in} should be multiplied by K to obtain the apparent C_{in}.

Returning now to the previous statement that the loss in gain of a cathode loaded tube is not necessarily an important factor, the student should realize that if the loss in gain is accompanied by a reduction in input capacity, it is often possible to increase the gain of a preceding tube due to the lower capacity loading, and this partially compensates for the loss in gain of the degenerative (cathode-loaded) stage.

The approach to the foregoing problem can be carried a little further to explain the "Miller Effect" which you studied in a previous lesson. In that case, the input capacity of a triode was given by the following formula:

$$C = C_{gk} + C_{gp}(1+A)$$

The term A denoted the gain of the stage considered. Consider Fig. 55C. If the grid voltage of the amplifier stage has been increased four volts, the plate voltage would be reduced a certain amount, depending upon the gain of the tube. This is true because the grid and plate voltages of an amplifier stage are 180° out of phase. If the triode considered had a gain of three, then the plate voltage would be reduced by 12 volts when the grid voltage was increased by 4. Thus, the difference in potential between the grid and plate would be 16 volts, and 16 lines of force would be set up between these points. To charge this condenser by such an amount would require a large capacity current flowing in the grid circuit. This corresponds to a high effective grid-plate capacity. Thus, the grid-plate capacity of the tube would seem to be $C_{gp} \times (1+A)$, where C_{gp} is the static grid-plate capacity.

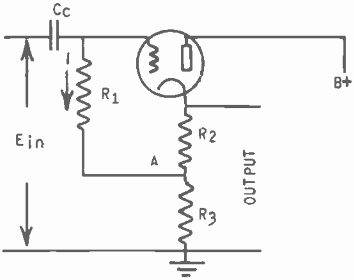


Fig.56 Effect of cathode loading upon input resistance.

Not only capacity current but in-phase or resistance current is also reduced by the application of cathode degeneration. The reason for this is exactly the same as the reasoning applied in considering the effect of input capacity reduction; namely, the changes in grid voltage are accompanied by corresponding changes in cathode voltage so that the difference in potential between these two points is reduced. Consequently, the currents flowing in the grid circuit are likewise reduced and the apparent input resistance of the tube is increased.

Referring specifically to synchronizing generators and control room equipment, the increase in input resistance has the advantage that smaller coupling condensers may be used. This comes about due to the fact that not only the tube resistance, but the resistance of the grid leak will be increased if it is returned to the cathode as shown in Fig. 56. As the grid voltage increases, the grid return at point A comes up to meet it, and the reduced drop across R_1 results in a smaller current than would normally flow in this resistor. Since $R = E_{in}/I$, the effective input resistance has been increased and condenser C_c may be chosen to have a small value. The effective input resistance is given by the following formula relative to the notation of Fig. 56:

$$R_{in} = R_1 \times \frac{1}{1 - \frac{R_3}{\frac{1}{G_m} + R_2 + R_3}}$$

Returning to Fig. 54, it is apparent that improved linearity also results in a sharper cutoff of plate current versus grid voltage. The grid swing necessary to cut the tube off may be reduced by *lowering the screen voltage.* When this is done in conjunction with cathode loading or degeneration, the limited grid swing and sharp cutoff characteristic makes the tube ideally suited to the function of an output amplifier in the control room video circuits. The student should recall that for this purpose the blanking signals were included in the video amplifier stage ahead of the output stage. The output stage was then driven to cutoff by the blanking signal which determines the absolute black level of the signal. Unless the tube has a sharp cutoff characteristic, the black of the picture, which occurs near the blanking level (cutoff point on the output stage) will be saturated due to the curvature of the tube characteristic. In other words, a sharp cutoff characteristic, which may be achieved by cathode degeneration, avoids a decrease in the contrast of the shadows or dark tones in the picture.

Another advantage of cathode degeneration which comes about due to the improved linearity, is that we are able to take advantage of the full plate current swing of the stage. Thus, the output voltage can be increased if there is sufficient available grid swing. This is relatively unimportant in low level video stages, but becomes more and more important in high level video stages, as you will discover in a later lesson.

12. REGULATED POWER SUPPLIES. In synchronizing generators, as well as in other parts of the studio equipment, regulated power supplies are necessary to maintain a constant output voltage in spite of changes in line voltage. Regulating the power supply also removes residual hum from the filter section. Some types of regulated power supplies, and most all of those used for television, also regulate against load changes. Since they maintain a constant output voltage in spite of changes in load current, they are said to have a low internal impedance. A good power supply will have an internal impedance of from 3 to 10 ohms for a 250 volt output, unless extremely large load current changes are involved. This

value is considerably lower than heavy duty dry B batteries, and approaches the internal resistance of a bank of ordinary storage batteries. The low effective internal impedance of the pack allows us to operate several stages or several units from a regulated pack without excessive crosstalk between them. The internal impedance is not low enough to obviate the use of decoupling networks, but it makes the problem somewhat easier.

Since the regulator in a regulated pack eliminates the effect of input voltage changes, it makes no distinction between line voltage changes and changes at a 60 cycle rate due to insufficient filtering. Thus, hum is considerably reduced. A typical pack will have a hum voltage of approximately .01 to .001 volt for a DC output voltage of 250 volts.

A further advantage of the regulated pack, although somewhat less important, is the fact that they may be set to a predetermined voltage, and will maintain this voltage, regardless of the output voltage of the rectifier. Thus, in a typical pack, a simple knob on the unit allows us to choose any voltage between 150 and 350 volts.

The regulated packs to be studied may be divided roughly into four classes; those employing series regulator tubes, those employing shunt regulator tubes, packs employing a bridge type of circuit, and those employing grid controlled rectifier tubes.

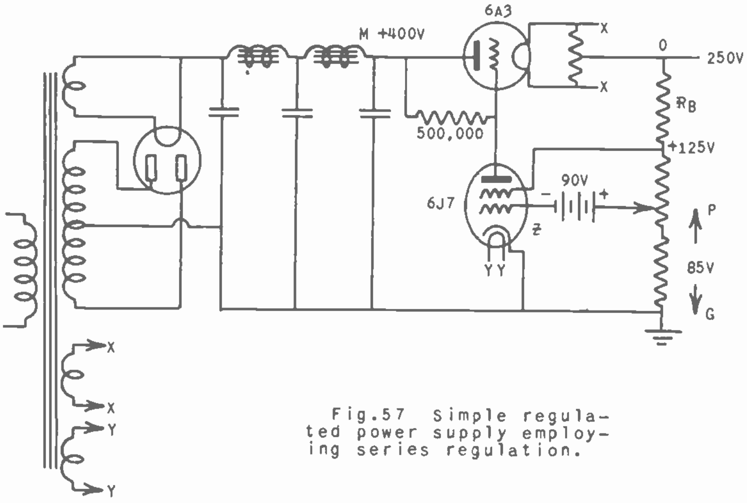


Fig.57 Simple regulated power supply employing series regulation.

Regulated power supplies employing series regulator tubes are by far the most common, and several variations of the basic circuit are often used, depending upon the degree of regulation required. Fig. 57 shows the fundamental circuit which is in itself a practical power pack. The regulator circuit is a form of two stage direct-coupled amplifier employing negative feedback. The 90 volt B battery serves to buck the voltage between point P and the grid of the 6J7 so that it operates with a normal bias of approximately -5 volts.

Before discussing the operation of this circuit, the student should become familiar with the following two terms. The series tube, shown as a 6A3 in Fig. 57, is known as the *rheostat tube,* and the 6J7 is known as the *control tube.*

The values of voltage and the size of the components in Fig. 57 were chosen arbitrarily, and could be altered to suit the particular needs of the user. Suppose the voltage at point M should increase from 400 volts to 410 volts. This would tend to increase the voltage at point O and 1/3 of this voltage change would be applied to the grid of the 6J7. (This is because the grid is tapped at about 1/3 of the way up on the voltage divider.) The 6J7 would draw more current and increase the bias of the 6A3. This would increase the plate resistance of the 6A3, and consequently most of the voltage increase of 10 volts (which we assumed) would appear as a drop across the 6A3. The voltage at O would rise only very slightly.

Exaggerating, for the purpose of illustration, suppose that the voltage at point M increases to 500 volts. If 60 ma. are flowing through the 6A3, the bias on the 6A3 must increase by 25 volts to increase the drop through the 6A3 by 100 volts. Therefore, the plate voltage of the 6J7 must change by 25 volts; this is done with .2 volt shift in the grid bias. In other words, if the voltage at point M increases by 100 volts, the maximum voltage increase at point P will be .2 volt, and the output voltage delivered at point O will be increased by .6 volt. Thus, the improvement produced by the regulator circuit is more than 80 times. Much greater improvement can be achieved with more complicated circuits to be described later.

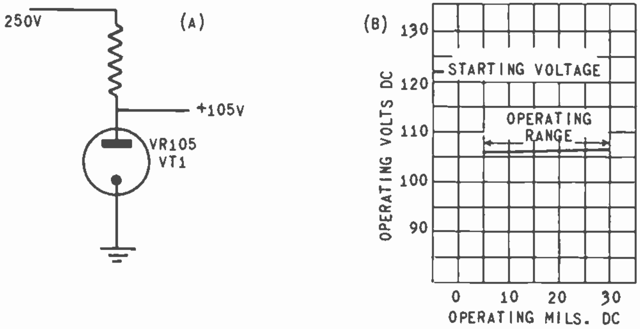


Fig.58 Performance of VR-105 gas filled regulator tube.

You have just studied how a pack regulates against line voltage changes. Consider next the effect of load current changes. The normal tendency of a load variation is to change the voltage at point O. Suppose the load is increased and the potential at O drops below 250 volts. This will lower the voltage at point P and increase the bias on the control tube. With increased bias, the 6J7 will pass less current through its load resistor and consequently reduce the bias on the 6A3, thereby lowering the drop across the 6A3 to compensate for the extra load drawn by the load circuit.

The simplest type of regulator available is shown in Fig. 58A. If a glow discharge (gas filled) tube is connected in series with a resistor, it will maintain essentially constant voltage across the tube, regardless of the current flowing through it within the rating of the tube. The regulation curve is shown in Fig. 58B for a type VR-105. Although the regulation afforded by this circuit is not good enough, and the permissible current not high enough for most television applications, the circuit finds application as a part of more complicated regulator systems.

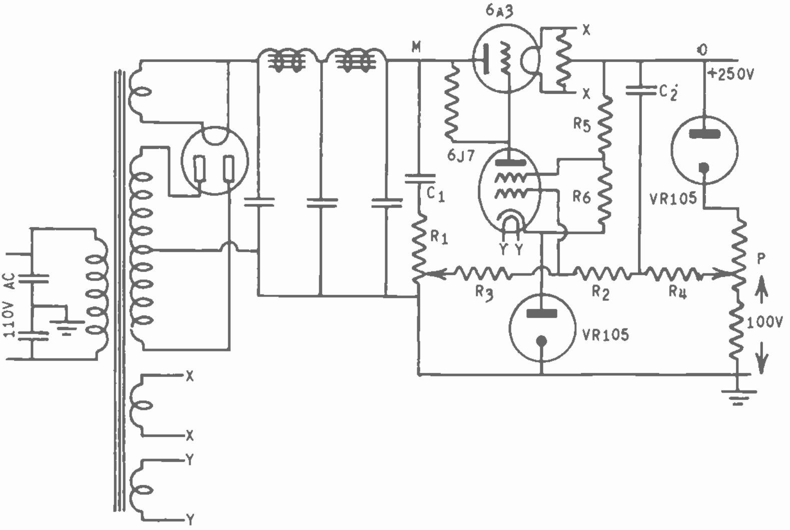


Fig.59 Regulated power supply employing special "anti-bounce" circuits.

Fig. 59 is fundamentally the same as Fig. 57. If the cathode of the control tube is now placed above ground by 105 volts, which is maintained by a gaseous regulator tube, the resistors R_5 and R_6 serve as a voltage divider to obtain screen voltage for the 6J7, and at the same time maintain a constant drain through the regulator tube from the stabilized 250 volts output of the pack. Consequently the drop through the VR-105 is made exceedingly constant. R_b of Fig. 57 has now been replaced by an additional gaseous regulator tube. Since this tube maintains a constant drop, any voltage change at point O will be applied almost directly to point P instead of through the voltage divider as in Fig. 57. This increases the quality of the regulation by a factor of almost 3. The condenser C_2 and the resistor R_4 insure that any rapid changes in the supply voltage, such as a sharp bounce due to a line surge (or a sharp increase in the current demand) will be delivered directly to the grid of the 6J7 by capacity coupling through C_2 with R_4 and the divider network serving as a grid leak. The impulses at 0 would be applied directly to the grid of the 6J7 if the network R_2R_3 were eliminated which is quite often done. This network serves a different purpose, which we shall now consider.

It is obvious from Fig. 57 that no regulation can occur until after a voltage change has taken place at point 0. If we wish to secure perfect regulation against hum or line voltage bounce, the circuit of Fig. 57 can only approach this desired result, but can never give perfect regulation. If a voltage could be applied to the control tube from ahead of the rheostat tube; that is, from point M, then it would be possible to secure either positive or negative regulation. The circuit of such a regulator is shown in Fig. 59 and with proper adjustment, almost perfect cancellation of voltage changes at point M can be had.

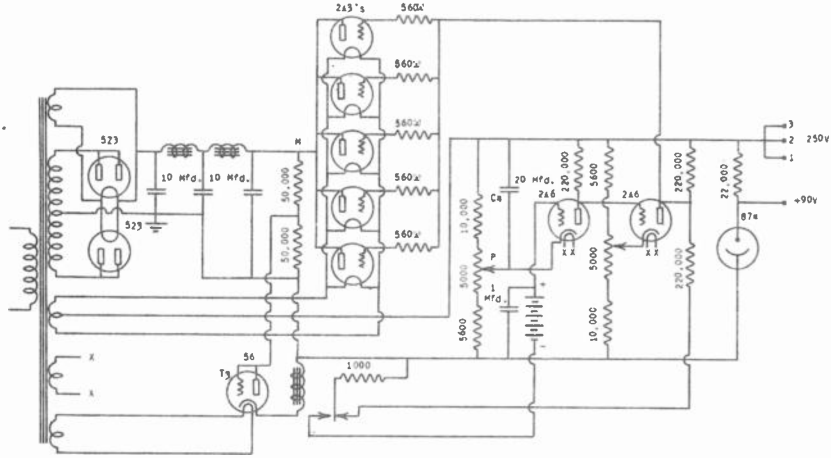


Fig. 60 Regulated power supply employing two stage DC coupled control circuit.

Suppose a decrease occurs in the voltage at point M. This will be applied by capacity coupling to the grid of the 6J7 through the voltage divider network $R_1R_3R_2$. If R_1 is adjusted so that most of the change at M is applied to the grid of the 6J7, the 6J7 will be biased completely to cutoff, reducing the resistance of the 6A3 to a minimum. This will reduce the drop across the rheostat tube to such an extent that the voltage at point 0 will actually rise above the value that it had originally. This condition is known as overregulation. The setting of R_1 can be adjusted for nearly zero regulation for the particular conditions encountered. For this circuit, it is necessary that R_1C_1 have the same time constant as R_2C_2 and that R_3 be large compared to either R_1 or R_2; otherwise, some types of surges will produce overregulation and some underregulation.

For simplicity, a single 6A3 tube has been shown in Figs. 57 and 59. If a large amount of current is required from the power supply, several rheostat tubes can be operated in parallel, allowing 60 to 100 ma. per tube. Fig. 60 shows a power supply used in a large cathode ray oscilloscope, which operates with five rheostat tubes in parallel. This circuit differs from that of Fig. 57 chiefly in the fact that two stages of amplification are employed in the control tube circuit. These stages are direct-coupled, and consist of a pair of triodes. The extra stage in the control circuit reverses the phase of the control voltage. For this reason, point P on the voltage divider across the output of the pack drives the cathode of the first stage, rather than the grid. The grid is operated above ground through a battery in contrast to Fig. 59 in which the cathode of the control tube was operated above ground, using a glow lamp. The condenser C_4 in Fig. 60 applies rapid changes in the output voltage directly to the cathode of the first tube, thereby eliminating the voltage divider in the cathode circuit for surges. This has the advantage of improved regulation for AC changes or rapid fluctuations.

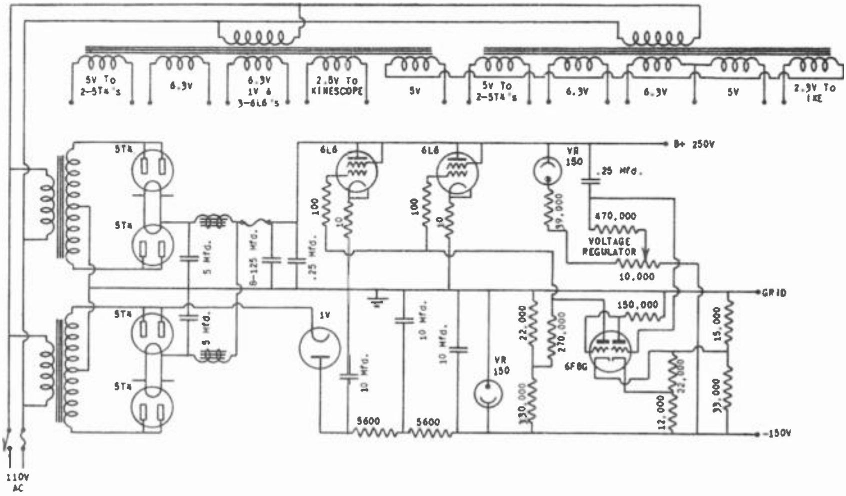


Fig. 61 Shunt regulated power supply

The 874 forms no part of the regulator circuit but is incorporated to secure a separate steady output of 90 volts. The type 56 tube T_3 is incorporated as a time delay circuit which grounds the grid of the rheostat tube and causes the output voltage to maintain a low value of 50 or 100 volts until all of the tubes in the circuit have reached operating temperature. When this time delay circuit is not incorporated, as in the case of Figs. 57 and 59, the output voltage will rise to some high value approximately 50 volts less than that appearing at point M. This high value will be maintained until the filament of the control tube (a heater type tube) comes up to operating temperature. When operating temperature has been achieved, the output voltage then drops to the normal voltage, determined by the setting of the control potentiometer at point P. By employing a two stage amplifier in the control circuit, by careful design, and by the use of heavy duty chokes and transformer, the effective internal resistance of the power supply in Fig. 60 is made approximately 3 ohms.

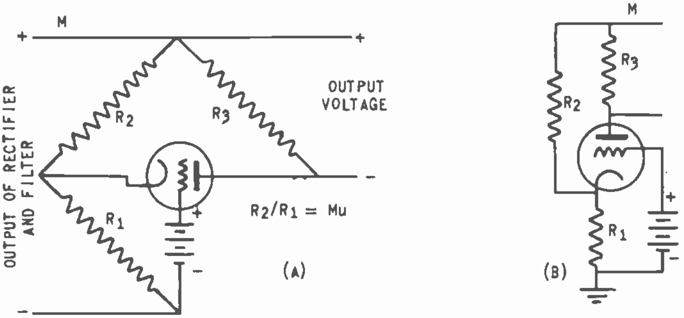


Fig.62 Simple regulated power supply for small current demands.

In the foregoing regulated power supplies, the rheostat tubes were operated in series with the load and the output of the filter. In some cases it is preferable to place the rheostat tubes across the load and regulate them in such a way that the sum of the load current and the current taken by the rheostat tubes is just sufficient to maintain the proper output potential from the power supply. When this is done, the rheostat tubes cause a variable internal drop in the power supply as the supply voltage fluctuates. This is practical since there is always several hundred ohms of unavoidable resistance in the rectifier and filter circuits. The control tubes in this circuit must operate at a potential negative with respect to ground, and this requires a separate bias supply, shown in Fig. 61 as the 1-V and VR-150. In this circuit, as the output voltage increases, the grid of the first triode in the control circuit is changed in the positive direction. Since the control tube operates in a two stage direct-coupled amplifier, the output voltage has the same polarity as the input voltage. Thus, the grids of the rheostat tubes (parallel 6L6's) are changed in a positive direction (that is, less bias is applied to them). The two 6L6's draw more current from the power supply, and thus act to lower the output voltage, which is opposite to the change originally assumed. The rheostat tubes have a stabilizing influence on the output voltage, regardless of whether the tendency of the voltage change was due to fluctuations in line voltage, hum, or a change in the load demand on the power supply.

When a small power supply is required for light duty operation and the load on the supply is relatively constant, the simple circuit shown in Fig. 62 will maintain a constant output voltage with reasonable fluctuations in input if the ratio of R_2/R_1 equals the amplification factor of the tube. For a medium or high mu tube, R_1 will be small compared to R_2. If the circuit is re-drawn as shown in Fig. 61B, it is seen that as the plate supply voltage increases, the cathode voltage is increased by a factor equal to the amplification factor of the tube. Increasing the cathode voltage is the same as placing a negative voltage on the grid. Consequently, we may say that the plate voltage and grid voltage are changing in opposite directions and at a rate proportional to the amplification factor of the tube. When this takes place there is no change in the plate current drawn by the tube and the drop across its plate load will be constant. Since the load is delivered across the plate resistor R_3, this voltage will remain constant regardless of changes in input voltage.

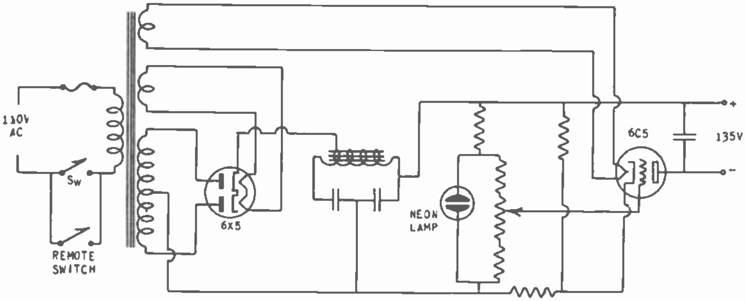


Fig.63 Practical circuit of low current regulated supply.

Eliminating the battery and re-drawing the circuit in a practical form, Fig. 63 illustrates a regulated power supply which will furnish 135 volts at approximately 10 ma. while maintaining the output reasonably constant and with less than 2 millivolts of hum. In this circuit the battery has been replaced by a neon lamp which maintains the steady voltage required for biasing the control tube.

If a regulated power supply is used to furnish bias to a television circuit, no changes in load will be anticipated. Furthermore, the current supplied by the pack will be negligible. In this case, the circuit of Fig. 64 may be used. The 6J5 functions as a half-wave grid-controlled rectifier, and by adjusting the time constant in the load circuit (choosing correct values for C_1, R_1, C_2, L) either over-regulation or under-regulation may be secured. Essentially perfect regulation is possible by adjusting the value of the load resistor.

Since the cathode of the rectifier is connected to the output circuit, it will be maintained at a fairly high positive voltage throughout the cycle. The rectifier will conduct only when the plate circuit is more positive than the cathode. The grid is returned to the plate circuit through a 1 meg. resistor and the tube behaves as a diode instead of a triode during the first part of the cycle. As soon as the plate and grid voltage reach the ignition voltage of the glow lamp, it will break down and the voltage drop across the glow lamp will suddenly drop to the normal operating voltage of approximately 70 volts. Just before the glow lamp breaks down, the cathode has charged up to a very high positive voltage, and as soon as the glow lamp breaks down, the grid voltage becomes more negative than the cathode, and the rectifier conduction ceases. The input condenser of the filter circuit then discharges until the cathode arrives at a lower voltage than the grid. At this time, the rectifier will again conduct until the plate voltage drops below the cathode voltage. Thus, there are two periods of conduction during each positive half-cycle of plate voltage.

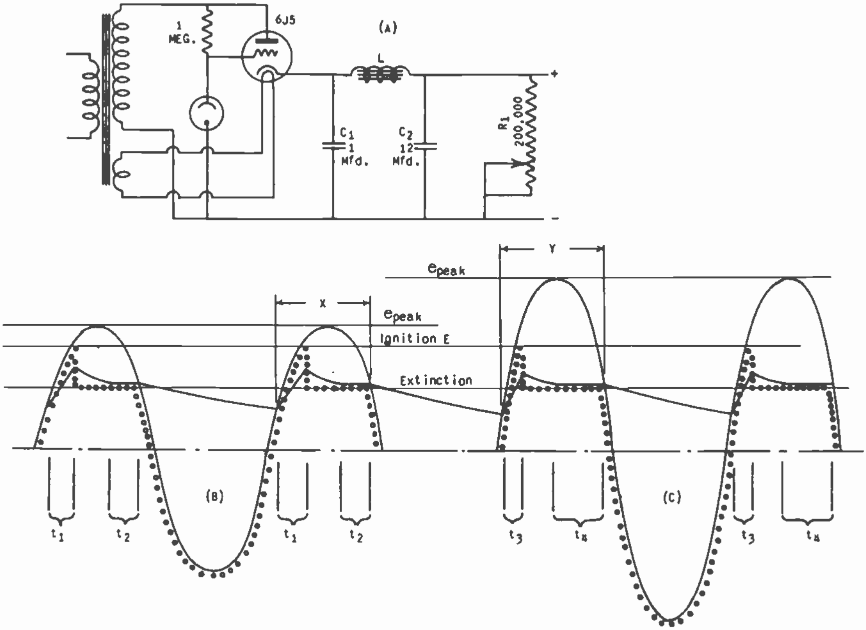


Fig.64 Regulated bias supply employing a grid controlled rectifier.

Figs. 64B and 64C show the operating conditions for two values of input voltage. The solid curve represents the plate-to-ground voltage, the dotted curve represents the grid-to-ground voltage, and the distorted sawtooth wave, labeled E_k, represents the cathode-to-ground voltage. The tube will conduct only when the grid and plate are positive with respect to the cathode, except for the brief interval of time when the grid is between zero bias and cutoff bias. The two periods of conduction previously mentioned are t_1 and t_2 in Fig. 64B and t_3 and t_4 in Fig. 64C. The ignition voltage and extinction voltage of the glow lamp do not change with the input wave.

If the input voltage becomes larger, its wavefront becomes steeper and a shorter time is required to reach the ignition voltage. Therefore, the initial charging period of the condenser (C_1

of Fig. 64A) is smaller and the peak value achieved is not as great. If C_1 discharges very slowly, then the rectifier conducts only once during the cycle, and the circuit over-regulates; that is to say, an increase in line voltage will result in a decrease in output voltage from the rectifier. If the load resistor R_1 is made smaller, then the condenser C_1 discharges very rapidly, and the tube conducts twice during the cycle. In this case, the duration of the charging time is longer for the increased line voltage as can be seen by comparing X and Y of Fig. 63. Since the charging time (t_3+t_4 compared to t_1+t_2) is longer, the power supply under-regulates; that is, an increase in line voltage will cause an increase in output voltage. By properly choosing R_1, a compromise is achieved between over-regulation and under-regulation which will give very satisfactory results. A bias supply regulated in this manner can be built very compactly and ample filtering can be provided.

At the present time, the preceding description includes all of the packs normally found in television equipment. Other varieties of regulated packs are used in scientific equipment, but have not been applied to television.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. What is the purpose of: (A) Equalizing pulses? (B) Serrations in the vertical synchronizing pulses?

2. Enumerate or sketch in block diagram form the essential timing unit circuits of an electronic synchronizing generator.

3. Describe two fundamental methods of creating rectangular voltage pulses from a sine wave.

4. Describe one method of decreasing the width of a rectangular voltage pulse; or draw a circuit to perform this function, sketching the waveform at each portion of the circuit.

5. Describe one method of increasing the width of a rectangular voltage pulse; or draw a circuit to perform this function, sketching the waveform at each portion of the circuit.

6. Describe one method of delaying a pulse; or draw a circuit to perform this function, sketching the waveform at each portion of the circuit.

7. In addition to equalizing pulse, horizontal and vertical blanking pulses, horizontal synchronizing pulses, serrated waveform, iconoscope blanking pulses, and driving pulses, at least two other pulses must be formed in the wave shaping circuits. What are the repetition frequencies, time durations, and purposes of these other pulses?

8. What advantages are to be gained by inserting the load resistance of an amplifier stage in the cathode circuit? What disadvantages?

9. What are the comparative advantages and disadvantages of the Scophony mechanical generator and a completely electronic generator?

10. Sketch a regulated power supply incorporating series rheostat tubes.

Notes

(These extra pages are provided for your use in taking special notes)

Notes

(These extra pages are provided for your use in taking special notes)

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UNIT
NO.
7

TELEVISION VOICE CHANNEL TRANSMITTERS

LESSON
NO.
4

TEAMMATES

.....SIGHT AND SOUND

This interesting booklet is devoted to a subject that is comparable to the sound track on a motion picture film.

Running along the edge of each movie film that is accompanied by sound is a small 'track'. The intensity of the 'track' varies, or appears to be shaded. It is this 'track' that produces the sound effects.....voices, music, gunfire, etc. Were it not for the marvelous accomplishments of this shaded line that runs along the edge of the film, our movies would be silent as in days gone by.

Television, to be successful, must also be accompanied by sound. Therefore, it is vitally essential that television programs have a sound 'track', just as movie film does. That is why television frequencies assigned to broadcasting stations are accompanied by ADJOINING sound frequencies. When the sound is tuned in, the television picture is automatically tuned in. This eliminates 'double' tuning, but it also means that each television transmitter must be operated in conjunction with a television SOUND transmitter. This all-important lesson is devoted to such transmitters. It should be mastered thoroughly.

Sight and sound are close teammates in the motion picture and television industries. The success of one is dependent upon the other. Ambition and knowledge are also close teammates. Ambition without knowledge cannot hope to succeed. On the other hand, knowledge is only secured through ambition.

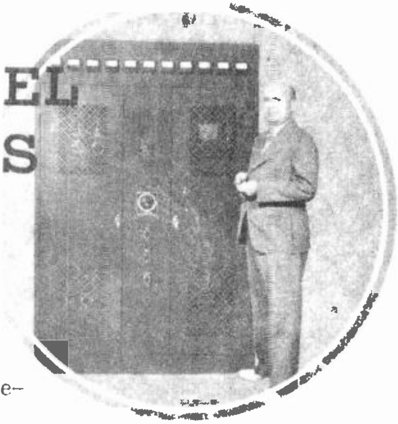
You have ambition. That is more than half the battle for success. You are gaining knowledge with each Midland lesson you complete. By successfully completing every lesson in our training plan, you will have achieved the formula for success.

Just what is this formula?

It is AMBITION, the willingness to work, PLUS a valuable fund of knowledge that will enable you to apply your ambition to a highly specialized, well-paid job in an industry that is limited to TRAINED MEN.

Lesson Four

TELEVISION VOICE CHANNEL TRANSMITTERS



"Since both the sight and sound of television programs will be broadcast on the ultra-high frequencies, it is very important that you become thoroughly acquainted with the use of apparatus operating on such frequencies.

"While the information contained in this lesson is quite important, still it becomes doubly so when considered as the foundation material for the next lesson on television transmitters".

1. CHARACTERISTICS OF ULTRA-HIGH FREQUENCY TRANSMITTERS. If the student has thoroughly mastered the basic principles of radio transmission covered in Units 3 and 4, then it is necessary in this lesson to consider only the differences that exist between ultra-high frequency voice transmitters and ordinary broadcast transmitters. The basic theory and the problems involved in the understanding of the operation of transmitters for the lower frequencies are also applicable to ultra-high frequency transmitters when due regard is taken for the limitations and peculiarities imposed by the use of these higher frequencies.

The refinements necessary or desirable for the proper operation of ultra-high frequency transmitters form the subject matter of this lesson. These refinements are not at all obvious to those familiar with only lower frequency transmitters and to some, the transmitter illustrated in Fig. 1 will seem highly unconventional. This transmitter is an extremely modern ultra-high frequency transmitter, and incorporates many essentials which you will study in this lesson.

As you progress through the lesson we will take up in detail the distinctive characteristics of ultra-high frequency transmitters. However, before considering these points in detail, it is desirable to outline the salient characteristics of these transmitters.

In the first place, the tubes are distinctly different from those used on the lower frequencies. Notice the small size of the tubes in the transmitter illustrated in Fig. 1. In Fig. 6 is shown a comparison between a small ultra-high frequency tube and an older style tube having equal plate dissipation. The comparison between these two tubes (888 and 851) illustrates many points brought out later in the lesson.

Wiring is another feature of ultra-high frequency transmitters which must be given careful consideration. In the transmitter of Fig. 1, notice the short heavy leads, and the symmetry and balance maintained in the push-pull circuit. Notice also that the tank circuits are different from those used in low frequency transmitters. Instead of coils and condensers, we find sections of concentric lines as well as open wire or balanced lines.

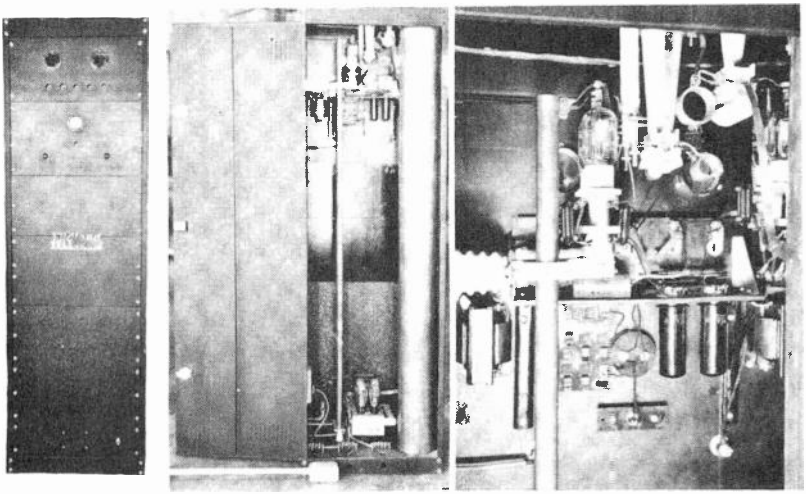


Fig. 1 Modern ultra-high frequency transmitters do not conform to low frequency convertions.

The methods of adjustments and measurements of these transmitters are dissimilar to lower frequency practices. On these frequencies, the R.F. ammeters, the dummy load resistors, and every inch of connecting leads become an appreciable part of the tuned circuit. Parasitic trap circuits and other dodges, useful on the lower frequencies are no longer applicable to the higher frequencies. The neutralizing condensers, bypass condensers, insulators and other components must be chosen with extreme care and with due regard for their decreased rating on the higher frequencies.

All of the foregoing considerations for the television voice channel transmitter are entirely applicable to the television picture transmitter which will be studied in the next lesson. The requirements of construction and adjustment of the television voice transmitter which lead to its efficient operation will also apply to the television picture transmitter. The only difference between the two is that the television transmitter must be capable of handling a much wider band of modulating frequencies. This problem forms the subject matter of the next lesson. It is therefore important that the student master the principles outlined in this lesson since they apply not only to the transmitter we are now studying, but to the picture transmitter as well.

2. POWER AND TYPE OF SIGNAL REQUIRED. The television voice channel transmitter must emit a signal of excellent quality. By this we mean it must not only be a high fidelity signal, but must be of sufficient signal strength to override all external noise. This must be so if the transmitter is to take full advantage of its wide band frequency allocation as well as the band width of the receiver.

The signal must be noise-free in order that it will not detract in any way from the enjoyment of the accompanying picture. Compared to the sound, the picture is relatively expensive to reproduce; therefore, any expenditure on the sound transmitter which will add to the enjoyment of the picture will be well worth while.

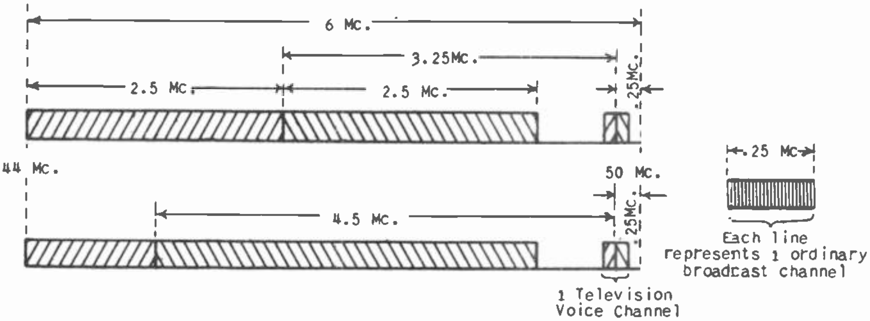


Fig.2 A comparison between television sound and ordinary broadcast allocation.

Referring to Fig. 2, we see that the available bandwidth of the television voice channel transmitter is extremely large, compared to the allocation of standard broadcast transmitters. With all danger of interference from other similar transmitters removed, we can widen the bandwidth and increase the fidelity as much as we please. The harmonic distortion of the transmitter should be as low as economically practicable. The audio response should be flat from 30 to 10,000 cycles. The sound portion of the conventional television receiver can easily handle the top audio modulation frequencies, since they are necessarily built with a wide I.F. channel, on the order of 100 kc., to handle the unavoidable oscillator drift in the receiver.

To answer the question of how much power is necessary to override all forms of external noise at the receiver, it is not necessary to talk in terms of an absolute field produced at the receiving antenna. We need talk only in terms of relative power between the sound transmitter and the television picture transmitter which is necessary to produce noise-free reception. Then, having decided upon the television picture power we intend to use, the power for the sound transmitter is automatically determined. This assumes that the voice transmitter operates either into the same antenna or into an antenna of equivalent gain as the one used by the picture transmitter. If only one receiving antenna is used to pick up the sound and the picture transmission, then this antenna delivers the two signals at the receiver with approximately the same intensity. The R.M.A. Committee on Television Standards ruled on the standardization T-114 that the voice channel transmitter and picture channel transmitter should operate with approximately equivalent power. When this ruling is followed and the antenna limitations just mentioned are adhered to, then in general the sound reception is just noticeably noisy when the picture is completely overridden with light flashes due to noise pickup. Interference which is insufficient to completely wreck the picture reception will not normally be noticed at all on the sound reception.

In order to compare the equivalent power of the sound transmitter and the television picture transmitter, it is necessary to first define the carrier power of the picture transmitter. The R.M.A. Television Standards Committee under standardization T-113 defines the carrier power of the television picture transmitter as one-fourth of the peak power. At the present time, standardization T-114 has only been approximately adhered to. The Columbia Broadcasting System television transmitter located in the Chrysler Building operates with a sound power of 7½ kilowatts for a picture carrier of only 4 kilowatts. The NBC television installation in the Empire State Building in New York City operates with a similar power rating. The Philco[1] experimental television station in Philadelphia operates with a picture channel power of 1 kilowatt for a voice channel power of only 200 watts. The General Electric station at Schenectady, N. Y. uses a power of 3 kilowatts for sound and 10 kilowatts for visual.

The British Broadcasting Company which is, of course, not subject to American standardization, nevertheless, follows quite closely to American practices in regard to the power involved. The Alexandria Palace transmitter operates with picture transmitter power of approximately 4.25 kw., for a sound channel power of approximately 3 kw.

We have shown the relation that exists in the power rating of the picture transmitter and the sound transmitter; just how much this power should be to cover a city of definite size with definite available antenna height will be discussed in detail in the next two lessons, but a fair idea can be gained from the example just cited for the cities of New York, Philadelphia, Schenectady, and London.

Next, we must consider the question of frequency allocation and frequency stability of the sound transmitter. We have already learned that a typical television superheterodyne receiver operates with a single oscillator and two fixed tuned I.F. amplifiers. This calls for a definite frequency relation between the picture and sound transmitters. Fig. 2 shows the R.M.A. standardization T-102

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[1] William N. Parker,---"A Unique Method of Modulation for High-Fidelity Television Transmitters". Proc. I.R.E., Vol. 26, PP 946-962, August, 1938.

for this frequency relation. The sound transmitter operates at a higher frequency than the picture transmitter. This is for convenience in receiver design and allows the receiver oscillator to be operated at a higher frequency than the incoming signal. This is desirable for the elimination of extraneous beats between the I.F. harmonics and the receiver oscillator. With this type of operation, the I.F. channel of the sound receiver is at a lower frequency than the I.F. channel of the picture receiver. It is desirable to have the picture I.F. operated at the higher frequency because of the bandwidths involved. The actual spacing between the television picture transmitter and the sound transmitter, as well as the width of the guard band between the sound frequency and the adjacent channel, was decided upon from considerations of the bandwidth and maximum possible steepness of cutoff of the I.F. channel of modern television receivers. The location of the television picture carrier frequency with respect to the edge of the band is a problem for the next lesson.

From a previous lesson, we learned that the bandwidth of the sound channel of a typical receiver was approximately 100 kc. From this, it would be assumed, on first thought, that the frequency stability of the sound transmitter would be of little consequence. However, receiving troubles due to frequency drift are a very real problem. It is a matter of good economics to make the sound transmitter as stable as possible to allow more freedom in the design of the receiver. One transmitter serves many receivers and the cost of good frequency stability at the transmitter is small compared to the total cost of good frequency stability in the thousands of receivers that this transmitter serves. Good frequency stability can be arrived at either through the use of crystal control, or by the use of special tank circuits to be discussed later in this lesson.

At present, the government requirements as to frequency stability for this type of service are moderately lenient. The required frequency stability is but .05% of the carrier frequency. For a typical carrier frequency of 55.75 mc., this amounts to a frequency tolerance of ±27,800 cycles. This tolerance seems quite wide and may possibly, be narrowed as time goes on. However, it is important that we think in terms of percentage frequency deviation rather than in actual number of cycles. For instance, a broadcast transmitter operating on a frequency of 570 kc. usually maintains a frequency stability to within five cycles. Doubling down from this same crystal oscillator frequency to drive a sound transmitter on a frequency of 55.75 megacycles, we would not expect a frequency stability of much better than 500 cycles. Yet, on a percentage basis, the high frequency transmitter is every bit as stable as the broadcast transmitter as is apparent from the fact that the same crystal oscillator was used to drive both.

3. BASIC CIRCUITS. Since there is no fundamental difference between television voice channel transmitters and other sound transmitters, all of the conventional circuits are possible. Some, however, are considerably more practical than others. At the present time, *the basic principle behind the design of television voice*

channel transmitters seems to be, "the simpler the design, the better". However, television is a comparatively new transmitter field and in view of the experimentation and changes that will probably take place in the next few years, it is well to mention all of the possibilities, with comments as to their relative merit. At the end of this lesson we will devote an entire section to the newer types of high frequency transmitters which have not yet emerged from the laboratory.

Probably the most talked about method of high frequency voice transmission today is the Armstrong frequency modulated system. This system is said to result in vastly improved signal-to-noise ratio, and demonstrations back up this claim. It is operated in such a way as to maintain the amplitude of the transmitted signal at a constant value, while the carrier frequency is varied by the application of voice modulation. The amount of frequency deviation from the carrier is proportional to the amplitude of the modulating wave; that is, a loud note would cause a greater frequency change than a soft note. The rate at which this carrier frequency varies is proportional to the frequency of the modulating wave. The receiver is arranged with a frequency detector to convert these changes in carrier frequency to changes in amplitude. Now, since static impulses and other types of noise occur as amplitude modulation on the carrier, rather than acting to change the carrier frequency, these noise impulses are eliminated at the receiver by the simple process of limiting their magnitude to equal that of the carrier in the receiver. In other words, the noise pulses are chopped.

Of course, considerable noise gets through in spite of the peak limiter, but only the noise components that lie within an audible band (14,000 to 20,000 cycles) from the carrier appear in the loudspeaker. Since the band width of the frequency modulated transmitter may be 100,000 cycles or more, only the weaker signals (corresponding to small frequency deviations from carrier) are affected by interference or noise.

This system is most interesting, but it is not suitable for television purposes, because it results in a complexity of transmitter and receiver circuits. Since the picture receiver is already complex and expensive, it is at present not considered worthwhile to add the complexities necessary for receiving this type of transmission. Any of the types of amplitude modulation now in use can be incorporated in high frequency transmitters. Ordinary grid modulation is a possibility for reasonable efficiencies on the order of 20% to 35%. This type of grid modulator must work with a high impedance plate load. This, of course, calls for moderately high plate voltages and loose coupling to the antenna. Such operation is difficult on the high frequencies because with loose coupling to the antenna, the losses in the tank circuit become appreciable and lower the overall efficiency of the transmitter. How this actually occurs will be better understood after we have completed this and the following lessons. At present, this type of modulation is not in common use on these frequencies for sound, but it is not impossible that it may gain favor in the future.

Absorption modulation is a possibility, but it is generally ruled out as being inefficient, non-linear, and difficult to arrange for high values of modulation; that is, 100% modulation.

Load impedance, or transmission line modulation is a new system introduced for television purposes. This will be discussed in detail in the next lesson. For the purposes of the voice transmitter, it is difficult to adjust for low harmonic distortion and lacks the simplicity of some of the more conventional forms of modulation. Being new, it is less generally understood and has not had acceptance for this type of service.

Suppressor grid modulation and screen grid modulation could be used with considerable success, but two factors discourage the application of these principles. First, is the unavailability of suitable multi-element tubes. Tetrodes and pentodes are not at present manufactured in sizes applicable to a high power, or even a medium power transmitter. Successful small multi-element, high frequency tubes have been manufactured and possibly the future will bring forth larger tubes of satisfactory characteristics. Secondly, screen grid and suppressor grid modulation suffer some of the disadvantages of grid modulation, in that the resulting tube efficiency is low and the circuits are slightly more difficult to adjust than are plate modulated systems. Plate modulation suffers none of the foregoing disadvantages, but of course requires somewhat greater modulating power.

This brings us to the next question. Should we choose low-level or high-level modulation? This has long been a potent point in the design of standard broadcast transmitters. The pro and con arguments advanced for these systems on the lower frequencies are pertinent for television voice transmitters. However, two additional factors become important on the ultra-high frequencies, and these factors swing the argument in favor of high-level modulation.

In the first place, the final stage tube complement requires considerable more dissipation in the R.F. section of a linear amplifier than in the case of an amplifier which is to be plate modulated. This is not a serious drawback on the lower frequencies, because the low-level system with the linear amplifier requires considerably less audio power. On the ultra-high frequencies, however, the size of tubes necessary to secure the required dissipation in the linear amplifier causes almost insurmountable circuit difficulties in a transmitter of considerable power. Within the next few pages, we will see just why this is so.

In the second place, the radio frequency circuits, for the low-level modulated, linear amplifier systems, are more elaborate than for a plate modulated, high-level system, of equivalent power. For this reason and also for the reason that audio modulation equipment is fairly well standardized from ordinary broadcast practice and therefore easily available, the high-level, plate modulated system will result in a somewhat simpler transmitter. This is highly desirable for a television voice transmitter in order that the time and energy of the station engineers may be concentrated upon the television picture transmitter which is inherently complex.

The foregoing considerations point to the choice of high-level plate modulation for our television voice transmitter. To date, this is the accepted method in use. This should not be considered a hard and fast rule. In any new field of transmitter work, there are almost as many ideas as there are engineers working on the problem. As more companies become interested in and work on the development of these transmitters, and as new circuits and tubes are brought forward, there are almost certain to be several answers to the problem. It is even possible that some companies may go "the long way around" to avoid patent interferences.

The circuit diagram of the UHF transmitter illustrated in the frontispiece is shown as Fig. 3. This circuit represents standard practice except that in some cases a separate modulation choke may be used.

The present day short cuts, innovations, fads, and fancies of general transmitter work are entirely applicable to ultra-high frequency transmitters. The use of inverse feedback, both R.F. and audio, as well as Doherty amplifiers, and Terman high-efficiency grid modulated amplifiers are all practical possibilities. These circuits need not be considered here, for with the exception of straight audio degeneration, all of these possibilities are avoided in modern television voice transmitters for the sake of simplicity, as mentioned previously. There is so much ground to cover in the understanding of current practices of these transmitters that we must skip these intriguing ramifications into the field of other types of sound transmission.

4. ULTRA-HIGH FREQUENCY TRANSMITTING TUBES. In the beginning of this lesson, we mentioned that tubes for ultra-high frequency transmitting service must have certain qualifying characteristics. Let us consider these in detail.

In addition to the factors which determine the value of a tube for general transmitting work such as a high transconductance factor, high available filament emission, which of course means high peak plate current and high peak grid current available, and long tube life, at least four other considerations are of prime importance in the selection of tubes for ultra-high frequency services.

The first factor which determines the performance of ultra-high frequency transmitting tubes is the interelectrode capacities involved. These should be extremely low. Large tube capacities cause the tank circuit to have an unnecessarily low LC ratio. A low LC ratio for a given output, or to put it another way, excessive tank Q makes it extremely difficult to keep the tank circuit losses to a reasonable value, because the ratio of circulating energy to the power dissipated in the load (the antenna) is too great. A large interelectrode capacity has another disadvantage. This disadvantage is the excessive R.F. currents which will be carried by the tube leads when the tube capacities are high. On these frequencies, the tube capacities are a large part, if not the major part of the total capacity across the tuned circuit. For this reason, most of the circulating tank current flows through the tube

connections and the leads to the tube elements. The R.F. currents flowing in these leads produce considerable losses and heating because of the small size and low conductivity of the leads themselves and the loss factor of the surrounding dielectric such as the glass press in the tube.

The consideration of the circulating current in the tube leads brings out the second point in the selection of ultra-high frequency tubes. These tubes must have heavy leads to carry the high currents which will flow in even a tube of moderately low capacity. These heavy leads might well be several straps about ¼" wide for tubes on the order of 1 kw. capacity. Solid leads, rather than braided or stranded leads should be used. Furthermore, the leads should be as short as possible between the external connections and the actual elements of the tube. The connections that bring the current through the glass wall of the tube must have a large R.F. current capacity and this calls for an exceedingly good seal. The ring type seal such as used in water-cooled tubes is excellent. On medium power, air-cooled tubes, the general practice is to bring several leads through the glass wall, all of which are connected in parallel. These leads are widely spaced to reduce the interaction and eddy current losses and the total current divides fairly evenly between the several leads. Several modern tubes suitable for service on frequencies around 50 megacycles are illustrated in Fig. 4. One of the tubes, which will dissipate several hundred watts, uses three wires in parallel to bring out the tube connections, each of these wires being approximately #14 wire and spaced about ¼" apart.

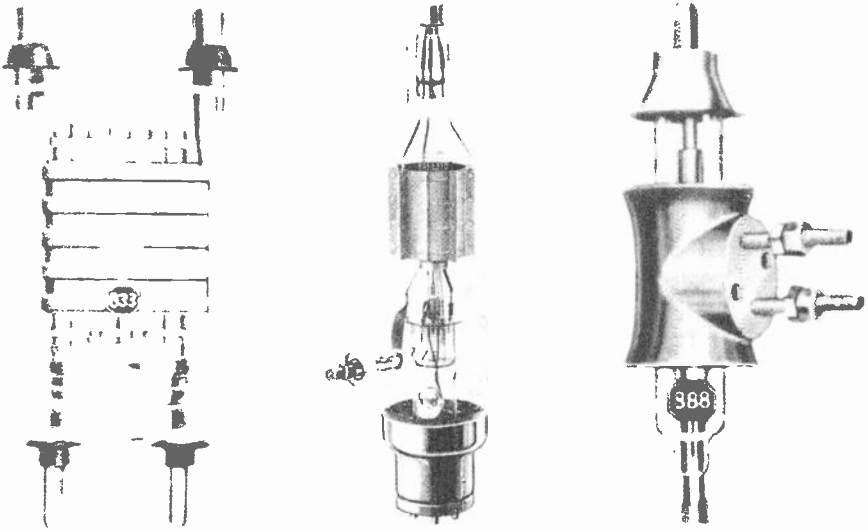
RCA 833 Eimac 450TH RCA 888

Fig.4 Tubes suitable for medium power television sound transmitter service.

The third important distinction of these tubes is the small physical size in relation to the amount of power handled. The small size not only contributes to the two factors outlined and to the reduction in transit time, (this will be explained in following paragraphs) but also reduces the standing waves which would otherwise be developed on the leads and the elements of the tubes. Since the tube's elements and leads have both distributed capacity and inductance, they act like a section of a transmission line which is open at one end. This causes a distribution of voltage along the elements of the tubes as shown in Fig. 5. In tubes of large physical dimensions compared to the wavelengths at which they operate, this voltage distribution is extremely detrimental to the performance of the tubes. It is easy to see that this uneven voltage distribution along the tube's elements would cause non-uniform operation in various parts of the elements, resulting in parts of the tubes running hotter than others. While this is undesirable, it is not as important as some of the other drawbacks of high frequency operation.

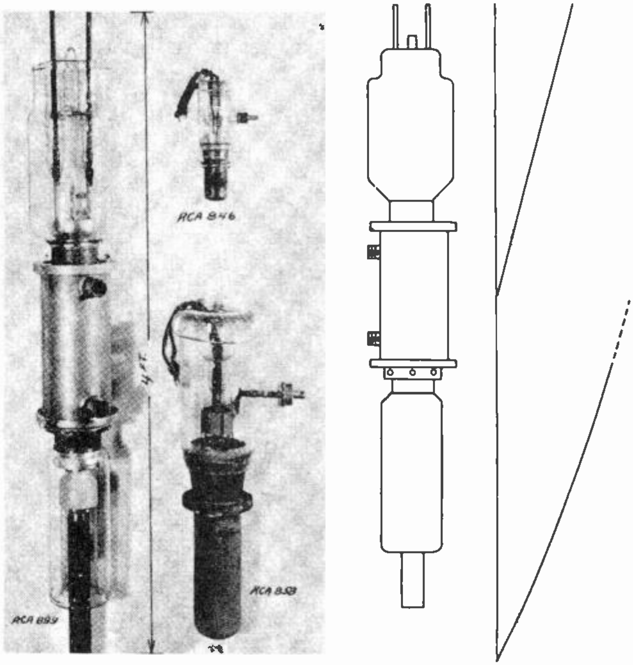

Fig.5 Small size is important in UHF transmitting tubes.

Because of the standing waves on the tube's elements, the actual voltage at the electrodes, which determines the flow of electrons within the tube, are considerably different than the voltages applied to the external connections. Furthermore, there is a difference in phase between the applied voltage and the voltage actually existing on the tube's elements, due to the definite time of propagation from the external connections to the elements. These differences make it difficult to predict or adjust the performance of the tube being used. In large water-cooled tubes, the standing waves on the tube's elements and leads may even be a quarter-wave long in the vicinity of 40 to 50 megacycles. This means that with minimum voltage applied to the external connections, maximum voltage actually exists between parts of the tube's elements as seen in Fig. 5. What has been done to overcome this limitation is shown in Fig. 6. This is a comparison, drawn to scale, of a type 888 and a type 851 tube. These tubes are of approximately equivalent power ratings, the 888 being a new ultra-high frequency tube.

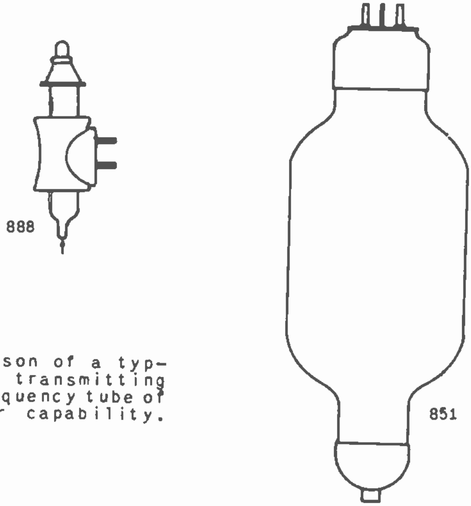

Fig.6 Comparison of a typical television transmitting tube and a low frequency tube of equivalent power capability.

The fourth important feature of the ultra-high frequency tubes is the close spacing between elements to reduce the transit time of the electrons. The electrons which flow between the grid and plate of the tube do not travel with the speed of light, but considerably slower. This time of flight of the electrons is of no particular importance for tubes operating on medium or lower frequencies. When the frequency of operation is low, the time of the electron flight (transit time) is very small compared to the time required to alter the instantaneous potential of the tube elements. When tubes are operated on the high frequencies, this is no longer the case. When the transit time cf the electrons becomes an appreciable fraction of the time to complete a R.F. cycle, there results an increase in the plate loss of the tube as well as in the grid driving power required. The tube will require grid driving power even when it is worked entirely in the negative grid region where no grid current would normally flow.

In addition to the normal input capacity of the tube, *there exists another capacity effect which causes the tube to have a high-*

er input capacity when the filament is heated than when it is cold. This effect is proportional to the transit time of the electrons and to the transconductance of the tube. We see how this effect takes place in Fig. 7. As the grid voltage increases, electrons are attracted from the filament toward the plate, but due to the finite time of travel of the electrons, there are more electrons going toward the grid than away from it. This causes a current to flow into the grid. When the grid potential is decreased, the electrons are more dense on the plate side of the grid than on the cathode side. Therefore more electrons are traveling away from the grid than are coming toward it and the current flows out of the grid into the space charge. Normally the grid current flow is 90° out of phase with the applied grid voltage so that its effect upon the input circuit is that of a pure capacity. This *electron capacity effect* is more objectionable at ultra-high frequencies than at lower frequencies for the following reasons. As the frequency becomes higher, the grid voltage changes before the electrons reach the grid because of the delayed arrival of the electrons at the grid. The capacity current is therefore shifted in phase so that it cannot be considered as a pure capacity current, but its *in-phase* component must be taken into consideration as shown in Fig. 7. *This in-phase component of electron flow to the grid results, of course, in grid loss* or an equivalent grid resistance of the tube.

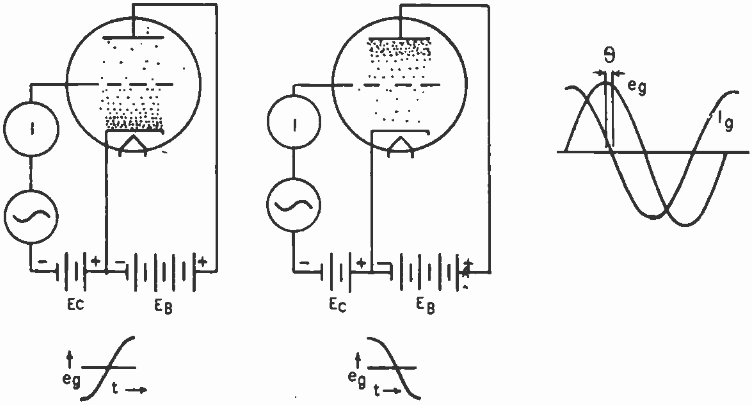

Fig.7 Effect of transit time on the input capacity and grid losses of a vacuum tube.

The plate losses of a vacuum tube due to transit time are not particularly serious in the case of an amplifier except at extremely high frequencies. For television transmitters, however, some of the tubes are operated near their limiting frequencies[1] so that this effect is of some importance.

In the case of vacuum tube oscillators, the plate losses due

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[1] Limiting frequencies: The highest frequency at which a tube can be made to function, i.e. deliver power to a load.

to transit time are considerably greater than when the same tube functions as an amplifier, as may be seen in Fig. 8. Oscillators are normally operated in such a way that the grid voltages and plate voltages are 180° out of phase. Since the pulses of plate current follow the waveform of the exciting grid voltage, the maximum plate current would normally occur when the plate voltage is at a minimum. Due to the time of travel of the electrons from the cathodes to the anodes at the higher frequencies, they arrive at the plate not when the plate is at its lowest potential, but after the plate voltage has started to rise. This causes the electrons to strike the plate when the plate is at a higher than normal potential. This is shown in Fig. 9. Since the power loss at the plate is a product of the voltage at the plate and the current (quantity of electrons), the losses at the plate are higher than for lower frequency operation. This, of course means that the plate efficiency will be lowered.

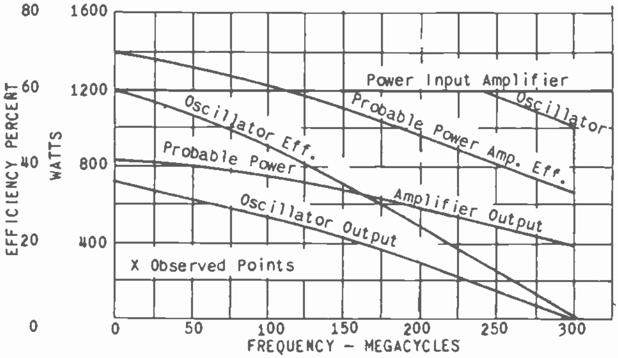


Fig.8 Performance of 888 tube.

In the case of an amplifier, this limitation does not exist, for when we tune the amplifier plate circuit for minimum plate current, we automatically adjust the phase of the plate voltage so that minimum plate voltage occurs at the peak of the plate current cycle. This could not be done in the case of the oscillator. (See Fig.9) Therefore, the efficiency of an amplifier at the higher frequencies is greater than that of an oscillator.

Another factor which causes an oscillator to have lower efficiency on the high frequencies than the efficiencies obtainable by an amplifier is the grid losses of the tube which we have discussed in a preceding paragraph. In the case of an oscillator, these grid losses of the tube must be supplied by the energy from the tank circuit. In the case of an amplifier tube, the grid losses are supplied by a preceding stage.

Some losses in an amplifier do occur, however, which are not imposed by lower frequency operation. These losses are due to the electrons present between the grid and the plate after the grid has reached cutoff voltage. Since it takes a certain length of time for these electrons to be entirely collected by the plate, on the higher frequencies the plate will have reached a higher voltage by the time these electrons are collected than in the case of low frequency operation. Therefore, greater heating of the plate will occur when these electrons arrive.

The problem of transit time effect is not quite as important as one would be led to believe when considering the large size and spacing of the elements in transmitting tubes necessary to withstand the high voltages generally used. These high voltages, coupled with the fact that grids of transmitting tubes are normally operated considerably into the positive region, produce high accelerating fields for the electrons leaving the cathode, so that the distance the electrons have to travel is partially offset by their high velocity.

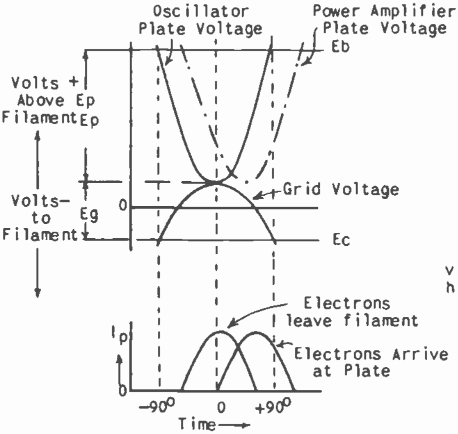


Fig.9 Phase relations in a vacuum tube operating at ultra-high frequencies.

Two other factors which are important in ultra-high frequency tube consideration have been inferred, but not directly named in the preceding discussion. These are the necessary low lead inductances and the absence of dielectric in the field of the electrodes. Lead inductance in the cathode causes losses in a tube in a somewhat similar manner to the effect of transit time. Our previous requirement of heavy, short leads to carry the high R.F. current automatically takes care of this consideration by lowering the inductance of the leads. High frequency tubes use as little dielectric internally as possible. The spacings between the elements are maintained only by the rigidity of the elements themselves which are normally supported directly from the glass seals.

For high frequencies when powers on the order of several hundred watts to several kilowatts are required, a water-cooled tube is generally to be preferred to an air-cooled tube of equivalent power. The elimination of the large glass bulb, the increased dissipation for a given plate size, the heavier, shorter leads, and the superior glass seals are favorable points consistent with the specifications we have set forth in the preceding paragraphs.

The use of a pure tungsten filament in a water-cooled tube allows the tube to clean up residual gases besides making the tube considerably more rugged with respect to temporary overloads. However, the emission efficiency of tungsten filaments is low. This is a decided drawback for television picture transmitters and even for television voice transmitters to a lesser degree. We shall consider this point in our discussion of television picture transmitters.

Ordinary triode and pentode transmitting tubes have been vastly improved in the past few years and it is intriguing to speculate upon the possible developments of the future. Two things are badly needed to improve the performance of ultra-high frequency tubes. One is a rugged filament with high emission efficiency to permit high values of peak current in the operation of the tube. The other is a plate material or cooling medium which will increase the dissipation per unit area from the grid and plate. Some form of liquid cooling which could operate at a higher temperature than water, or water under considerable pressure, might meet these requirements. An increase in dissipation would allow the tube to be made considerably smaller, possibly even to the point where high-power pentodes might be practicable.

5. WIRING AND ARRANGEMENT OF PARTS. The care and precision used in the assembly and wiring of ultra-high frequency transmitters is a constant source of amazement to radio men whose experience has been entirely with low frequency apparatus. Oftentimes it is desirable to forsake conventional arrangements of parts when such

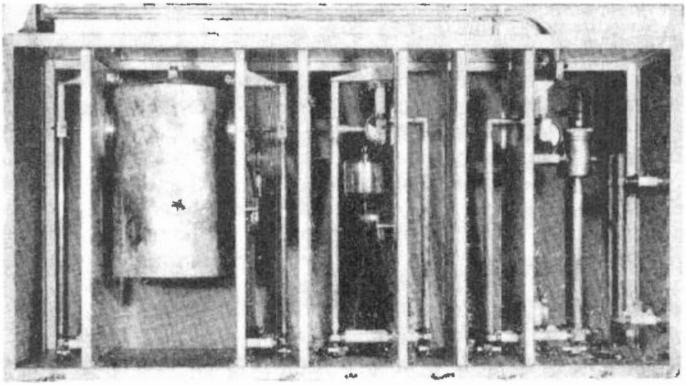


Fig. 10. High frequency transmitters are of functional design.

procedure will shorten lead lengths and improve the balance of the circuit. "Utility first and appearance second" might be called the high frequency slogan. Refer to Fig. 10, or back to Fig. 1 for an example of this functional design.

Possibly the most important point in the mechanical design and in the assembly of these transmitters is in arranging for a good ground connection. The ability to secure a good ground largely determines the ease with which we can control the mode of operation of the circuit. At the risk of being somewhat elementary for this advanced lesson, we will define ground and then explain how we can secure it. For our purposes, *the ground part of the circuit is the point which does not change its absolute potential with respect to the average potential of the universe.* Notice we said that it must have no change of voltage; in other words, no R.F. potential. We do not care whether this point is at B- or B+ or whatever its DC voltage may be, provided it has no AC voltage with respect to any fixed potential.

There are two ways we can secure a good ground. One is by the *use of a large electrostatic capacity,* such as a big chassis or a power supply. Such a large mass of metal has considerable electrical inertia. Trying to change its fixed potential is like grasping the side of a building and trying to shake it. The second way in which we can secure a good ground and perhaps the best way for high frequencies is by the *use of symmetry and balance in the placement of parts.* A symmetrical or push-pull circuit oscillates about the point of symmetry, just as a see-saw moves up and down on its pivot. The center point of our push-pull circuit we may call the neutral plane and this neutral plane is our electrical "see-saw" pivot. The use of a push-pull circuit gives us our balanced type of ground and for a single ended circuit we must arrange for a large capacity type of ground.

If we explain why the balanced type of ground is superior to the single-ended type of ground, we bring up another important point in our wiring problem. This point is the increased impedance of the connecting leads on the higher frequencies. Once more we will return to mechanical analogies for an explanation of this point. First, if we had a large flywheel with an off-center axle turned by a small motor, the motor as well as the flywheel would vibrate. This is equivalent to an ungrounded, single-ended circuit. Now, if we wanted to stop the motor from vibrating; that is, "ground it", we could mount it on a base with heavy steel studs. This is equivalent to tying the ground portion of our single-ended circuit to the chassis with heavy leads. These leads have a certain amount of inductance so that their impedance is proportional to the frequency at which we operate. On the high frequency with which we are concerned, the impedance of even a few inches of #10 or #8 wire becomes considerable. Returning to our mechanical analogy, we can no longer say that we have mounted the motor with heavy steel studs; our comparison is now as if we had mounted the motor on rubber studs so that the motor would vibrate almost as badly as when it was not connected to a base at all. As the frequency becomes higher, even the inductance of the chassis becomes important. In our analogy, it would be almost as though our base were made of rubber. With considerably care, it is always possible to make a single-ended high frequency circuit perform satisfactorily but it is generally easier to arrange our circuits in a push-pull fashion.

While the ground connections for a single ended amplifier impose the most stringent requirements, other circuits must also be wired with care. Wire becomes almost a misnomer when used in connection with television voice transmitters. Wires are definitely not good enough. Copper tubing or heavy copper straps are generally used to form all connections, since the greater area results in less inductance per unit length. The length of all connecting leads must be just as short as it is practical to make them. A typical 500-watt transmitter might have connecting leads made of copper straps ¾" wide, with not a lead in the entire transmitter exceeding four inches in length if it is to carry R.F. current. A calculation of the reactance of a piece of #10 wire 4" long at 60 mc. emphasizes the importance of this point.

$$L = .002l \; 2.303 \log_{10} \frac{41}{d} - 1 + \mu\delta + \frac{d}{2l} \quad \times 10^{-8} \text{ Henries}$$

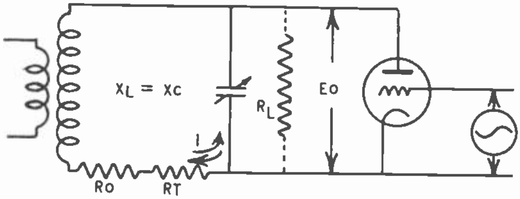
Where l and d are in cm.
δ becomes .009 for #10 wire at 60 mc.
$L = .0834 \times 10^{-8}$ Henries
$X = 2\pi f L = 31.4$ ohms.

All circuit components should be laid out with care, particularly in push-pull circuits where all leads should be carefully arranged for perfect symmetry. A difference in lead length between complementary tubes in a push-pull circuit of one-half an inch might cause unbalance in some cases. Output coupling links are particularly bad offenders since they are often called upon to feed a single-ended load such as a coaxial transmission line. In spite of extreme care in the layout of circuit components, some slight unbalance will always occur. These unbalances result in an in-phase component of R.F. current, and a return path to the center of the tube filament should be provided for these currents. The circuits in which these currents flow are indicated by the X mark in Fig. 3.

Unless the leads in the transmitter are extremely short, an appreciable portion of a standing wave appears across each of the leads. Television transmitters operate on wavelengths below six or seven meters so that the distance from a complete voltage maximum to a voltage minimum will be less than five feet. Furthermore, the inductances required in the tank circuits are quite small because of the large capacity involved. Unless care is taken, most of this inductance will be used up in the connecting leads so that the tuning condenser is no longer across the tank inductance. The total circuits associated with the tube must be considered as a complicated network of condensers, coils and transmission lines. Add to this the fact that the time of propagation of the current along the connecting leads causes phase shift between the voltages appearing at the tube and other portions of the associated circuit and you will see how important it is to use short leads.

We previously mentioned that large diameter tubing or flat copper straps were desirable in interconnecting the circuit components. When parts are to be moved such as in the replacement of tubes, one is often tempted to use flexible braid. This is to be guarded against. The losses in copper braid or in stranded cables are considerably higher than in copper straps or tubing of anywhere near the same size. On a lower frequency, such is not the case, but the losses on the higher frequencies are considerably greater, as may be understood from a consideration of skin effect. Skin effect is the tendency of an R.F. current to travel only on the outer portion of the conductor. The higher the frequency, the thinner the layer of copper in which these currents are carried. On television frequencies, the current may travel in only a few thousandths of an inch of the outer surface of the connecting leads. Braid is made up of many strands of fine wire which are interwoven. When one strand of wire goes underneath other strands of wire, the R.F. currents leave this wire and travel on the surface of the other strands in such a way that the currents always travel on the outside surface of the braid. Each time the current passes from one strand to another, it suffers losses due to contact resistance of these contacting strands.

6. TANK CIRCUITS. In radio transmitters, the tuned circuits associated with the tubes are called tank circuits and are conventionally formed from resonating coils and condensers. We proportion the inductance, the capacitance, and the antenna coupling so as to give a Q of from 10 to 20. The loaded Q of a tank circuit is the ratio of the reactance of the coil or condenser to the total resistance in the circuit. In Fig. 11, this total resistance is $R_o + R_T$, and consists of the reflected resistance of the output circuit and the equivalent loss resistance of the tank. In television transmitters, the tube capacities, the capacities of the neutralizing condensers, the stray capacity, and the capacity of the tuning condenser, set a certain *minimum limit* which is generally higher than the value we would like to use. Excessive tank capacity causes increased losses in the tank circuit as we shall see from the following discussion.



R_o = Reflected Resistance of Output Circuit.
R_T = Loss Resistance of Tank Circuit.

Fig. 11 Effect of LC ratio on tank circuit loss.

If we maintain the proper load and secure a constant power output from the transmitting tube shown in Fig. 11, the voltage Eo developed across the tank will be constant regardless of the LC ratio. As we increase the tank capacity (decrease Xc and Xl) the circulating current I, will be increased. If we double C, we must use only one-half the value of L to maintain resonance. If we use a coil of only one-half the inductance of our original design, the *loss resistance* of the circuit (Ro) will be halved and the circulating current doubled. Since the power dissipated in the loss resistance is I^2R, the *loss* in the tank (Ro) will be doubled:

$$W = I^2R$$
$$= (2)^2 \times \tfrac{1}{2} = 2$$

The tank losses in an ordinary coil-condenser combination are considerably increased at higher frequencies not only because of the decreased LC ratio, but also because the eddy current losses and skin effect are greater. The tank losses cause considerable heating of the coils and condensers causing them to expand and go out of tune as the transmitter warms up. Furthermore, the heating leads to early corrosion which increases the resistance of the coil and in turn increases the losses in the tank. The obvious expedient of choosing a larger diameter conductor is not a satisfactory remedy since the distributed capacity between turns increases in such a way that the decreased resistance of the copper is more than offset by the increase in circulating current in the coil. For low or medium power tubes where considerable care can be taken to reduce stray capacities, conventional tank coils are quite satisfactory, but for some applications, particularly for higher power installations, other forms of tank circuits must be sought.

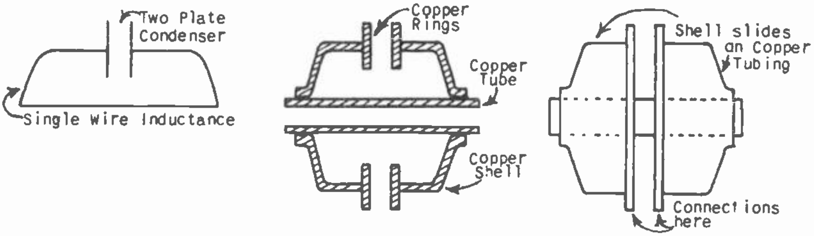
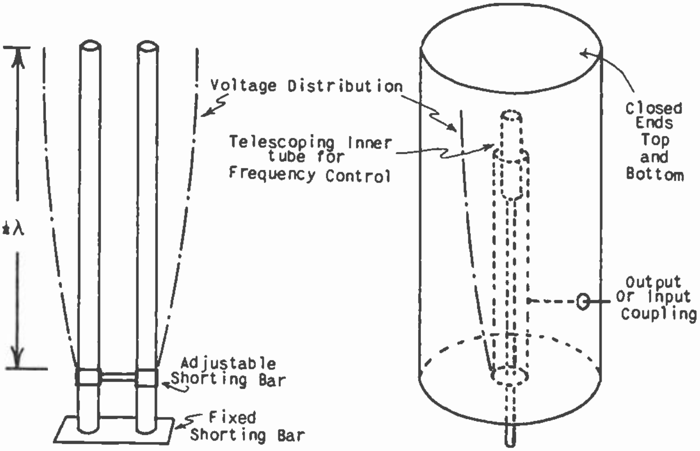


Fig. 12 Kolster low-loss tank circuit.

The tank circuit which is the nearest approach to the coil-condenser combination is the single turn torroid, developed by Dr. F. Kolster. This circuit is illustrated in Fig. 12. Because of the peculiar shaped tank circuit resembling two derbies placed together, this is often known as the Kolster "hat". This is a single turn coil with the tuning condenser formed by the brims of the hats. For a given amount of circulating energy, this coil develops relatively small losses for several reasons. Since the current flows on the inside surface of the hat, the flux produced by this current is entirely contained within the hat. Thus, there is negligible radiation loss. An ordinary coil has considerable loss of this type due to the leakage flux. Conventional coils tend to concentrate the current in a very small portion of the individual turns due to the manner in which the flux links these turns, causing eddy currents and crowding of the electrons. In the case of the Kolster circuit, the current flows uniformly around the inside of the hat and because of this even distribution of current and the large copper area involved, the resistance losses in the circuit are exceedingly small. A third and perhaps less important reason for the good performance of these tank circuits is that there is no dielectric in the electrostatic or electromagnetic fields as there is likely to be in conventional coils and condensers.



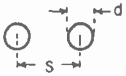
(A) Parallel Line Tank (B) Concentric Line Tank

Fig.13 Voltage distribution on transmission line tank circuits.

A more commonly used form of high frequency tank circuit is the parallel rod or transmission line tank illustrated in Fig. 13A. Here we substitute the distributed inductance and capacitance of the line for the lumped inductance and capacitance which we are accustomed to using. Notice in Fig. 13A the standing waves of voltage that develop when we feed energy into a quarter-wave transmission line. The theory underlying the performance of transmission lines will be considered in detail when we study television transmitting antennas and their feed lines. Although a section of line used as a tank circuit is electrically one-quarter wave long, its physical length is generally somewhat shorter than this because of the tube capacity and stray capacity which exist across the end of the line. In order to resonate these capacities we must make the transmission line tank circuit slightly inductive which is done by diminishing its length. For large tubes with their attendant high capacity, the physical length of the tank circuit might be as short as one-eighth of a wavelength.

The parallel line type of tank circuit will develop more circulating energy for a given loss than will conventional coils and condensers, because the area of the copper may be increased as much as we please without changing the LC ratio (provided we increase the spacing between the lines). In this case, L and C are distributed values and their ratio depends only upon the surge impedance of a line which in turn depends upon the ratio of the line spacing to the line diameter in the following manner.

$$Zo = 276.5 \log_{10} \frac{2S}{d}$$



This is an approximate form but is satisfactory for all practical values of line; that is when S÷d is greater than 2.7 or when Z is greater than 200 ohms.

Another type of transmission line having most of the advantages of the parallel lines, with some others besides, is the concentric, or coaxial type of line. A tank circuit formed by a quarter-wave coaxial line (illustrated in Fig. 13B, and seen in Figs. 1 and 10) has even lower losses than the parallel lines just described. Like the Kolster "hat", the field of a concentric transmission line is contained entirely within the outer conductor which acts as a shield, thus lowering radiation and dielectric losses. Furthermore, the currents that flow on the outer surface of the inner conductor and on the inner surface of the outer conductor are uniform about the circumference of any cross section. In the case of the open wire or parallel type of lines, there is some crowding of the electrons. Concentric lines do not lend themselves quite as readily to push-pull amplification and, in general, they are somewhat more difficult to arrange mechanically. The parallel line tank circuit is tuned by adjusting a short circuiting jumper which varies the length of the useful portion of the line. The same idea can be used in tuning the concentric line, but the doughnut-shaped disc to perform this action is difficult to build. If we can tap the load or input connection to the line at other than the open end of the line, we can vary the tuning by changing the length of the inner conductor previously arranged for telescoping as shown in Fig. 13B.

The use of the high Q tank circuits just discussed not only lowers the losses in the tank but improves the selectivity of these tuned circuits when they are lightly loaded. The use of a high Q tank in an oscillator is a method of frequency control ideally suited to ultra-high frequency transmitters. Using a Kolster hat or a concentric transmission line tank circuit, we can build an oscillator having frequency stability equivalent to or exceeding that of crystal controlled transmitters. Let us emphasize again that we must consider frequency stability in terms of percentage of the carrier frequency.

All other circuit constants being equal, Q is the measure of percentage of frequency stability of the transmitter. The Q for a good crystal might run upwards of 10,000. Let us compare these values with what can be obtained by use of concentric quarter-wave transmission line tank circuits. It has been determined that there is an optimum ratio of outer conductor to inner conductor diameter for any given outside diameter of coaxial line to obtain maximum Q. This ratio has been found to be 3.6. Curves showing the unloaded Q of transmission lines of various sizes and at various frequencies are illustrated in Fig. 14. It is seen that for frequencies above 40 or 50 megacycles, the diameter of line having a Q of 10,000 or greater becomes quite reasonable. Fortunately, the Q of a transmission line goes up with the frequency. Since it becomes increasingly difficult on very high frequencies to obtain sufficient excitation from doublers following a crystal oscillator, there is a double advantage in the use of this type of tank circuit on frequencies above 100 megacycles.

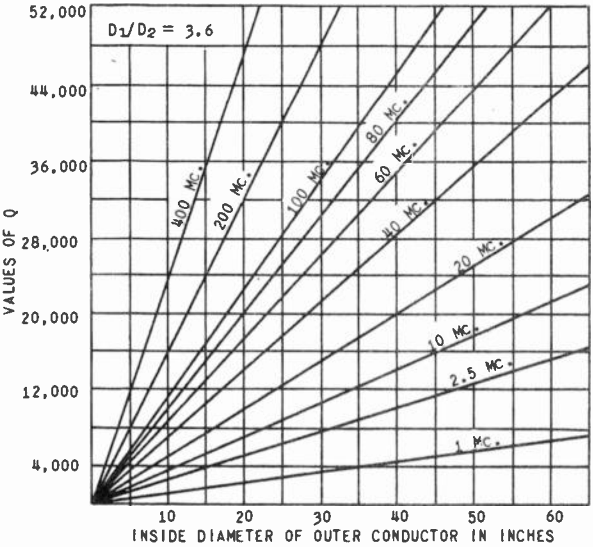


Fig.14 Q versus OD of coaxial copper tank circuits.

At frequencies now used for television voice transmitters; that is, in the vicinity of 40 to 50 megacycles, there is some debate as to the superiority of crystal control versus transmission line control of transmitters. In the case of crystal control, there is a multiplicity of tubes and circuits. In the case of transmission line control, certain other refinements required are partial disadvantages. In the first place, the lines themselves are moderately bulky and expensive. The lines must be mounted so as to be vibration-proof. Furthermore the values of Q as indicated in Fig. 14 are unloaded values and are not approached in practice unless special precautions are taken in the design of the oscillator circuits. The principal requirement is that the oscillator must be very lightly coupled to the concentric line and must feed back a minimum amount of energy to this line from its plate circuit. To accomplish this, the oscillator is partially neutralized and the neutralizing condensers are used as a feedback control. Furthermore, the grid of the oscillator is tapped near the grounded end of the coaxial tank circuit so that only a small portion of the energy circulating in the tank circuit is delivered to the oscillator. Another requirement is that the oscillator itself must be lightly loaded by the buffer amplifier in order that it will work with a minimum amount of excitation and a minimum reaction to the grid circuit by changes in the plate circuit. This requirement means that the oscillator will be working with relatively low efficiency and therefore larger tubes are required than would be the case in the crystal controlled circuit. Since the tank circuit does carry considerable circulating energy, and since it is subject to ambient temperature changes, the

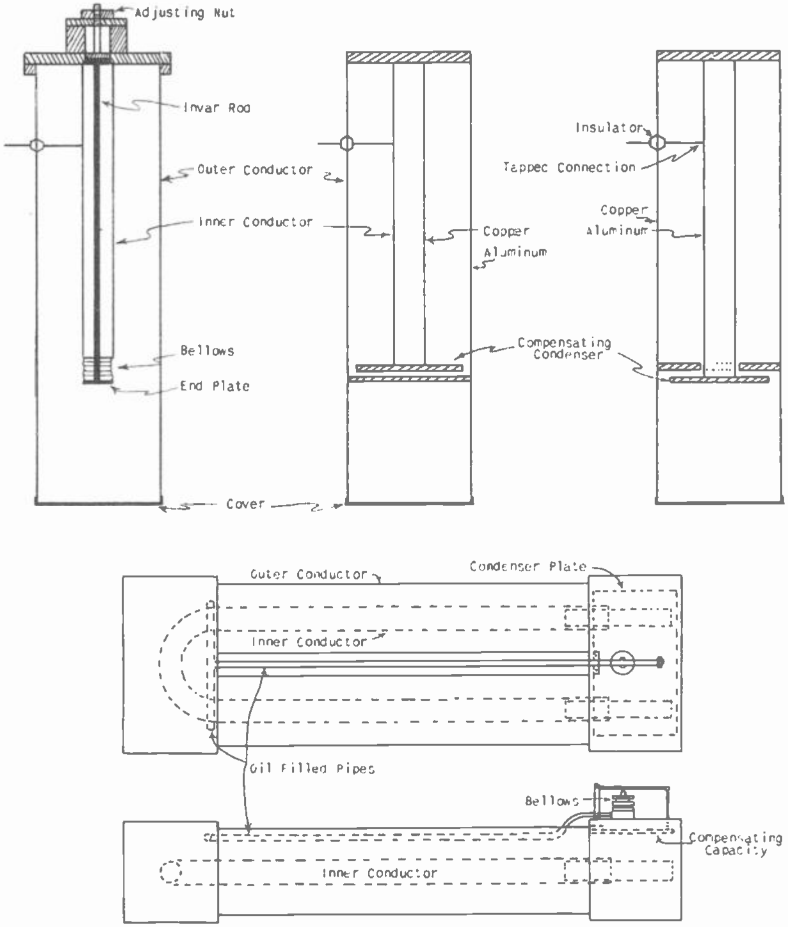


Fig.15 Temperature control of coaxial lines.

expansion and contraction of the copper during normal operation of the transmitter will cause detuning of the tank circuit. Several ingenious means have been devised to compensate for this effect. The simplest and most common method is the use of an invar rod inside the center conductor of a coaxial line. This rod is connected to a flexible bellows on the end of the inner conductor. This invar bar, which does not change its length with temperature changes, maintains the length of the inner conductor and thus determines the frequency of the tank circuit, regardless of temperature changes. Several more precise, but more complicated means include the use of a bellows filled with oil, connected to an oil expansion chamber which changes with temperature; and the use of a compensating condenser whose spacing changes in such a way that the capacity variation compensates for the changes in tuning of the tank circuit. Some of these means are illustrated in Fig. 15. A typical transmitter circuit using a line controlled oscillator is shown in Fig. 16. This is the new Columbia television transmitter installation in New York City.

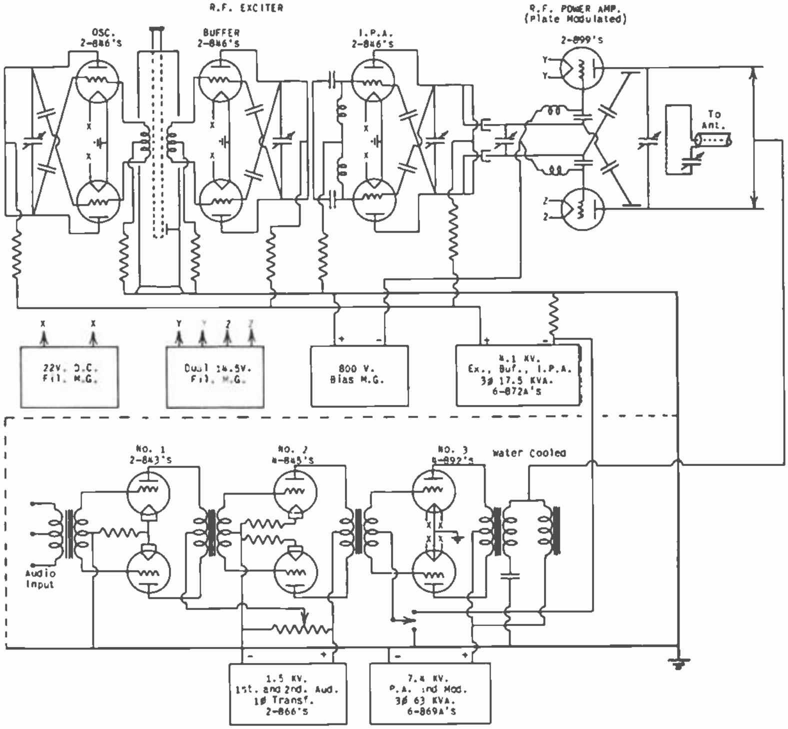


Fig. 16 Circuit of the CBS television sound transmitter in New York City.

7. FURTHER APPLICATIONS OF TRANSMISSION LINES AS CIRCUIT ELEMENTS. One transmitting application of a quarter-wave section of transmission line is known as a metallic insulator. This application could also be thought of as a single frequency choke coil. Thus we can ground one end of a quarter-wave rod and the other end will exhibit an extremely high input impedance. A section of quarter-wave coaxial line shorted at one end presents an even higher input impedance because of its lower losses. Thus we might use a quarter-wave steel bar to support an antenna structure so that while the antenna would be effectively insulated from ground with respect to R.F. currents, a good DC ground would exist for lightning protection and for heating current when it is desired to melt sleet and ice on the antenna. A quarter-wave section of transmission line bypassed to ground at one end might be used to feed plate voltage or bias voltage to an amplifier, just as a tuned R.F. choke coil might be used to feed in plate voltage or bias voltage to an amplifier on the lower frequencies.

Sections of transmission lines not only form good insulators, but when desired can be so arranged as to form exceedingly effective short circuits. A low loss line one-quarter wave long, open at the far end reflects a short circuit at the input end. Thus, if we had two points several inches apart, we could secure a much better short circuit between these points by using a line a quarter of a wave long, open at the far end than we could obtain by wiring directly between the two points. This would be an effective R.F. short circuit only. If we also wanted to supply a path for direct current we should use a half-wave line, short circuited at the far end. The half-wave line reflects the same impedance at the sending end as we place across the far end of the line. In this case we can nullify the reactance of the shorting bar on the end of the half-wave line by adjusting the line to a slightly shorter length than a full half-wave. Here again we can secure a better short circuit than we could by wiring directly between the circuit elements. This almost seems like a case where the shortest distance between two points four inches apart is a half-wavelength.

The most useful application of this principle is in inter-connecting vacuum tubes. You will recall from our section on ultra-high frequency transmitting tubes that one of the limitations in the design of these tubes was the inability to build tubes which were sufficiently small to eliminate the effect of standing waves on the tube elements and the leads to these elements. One way to eliminate the effect of these leads is simply to extend them until the total length of the leads plus the extension becomes a half-wavelength long. Thus if we wish to eliminate the effect of the standing wave on the filament leads, in the tube illustrated in Fig. 5, we operate two of these tubes in push-pull and connect the filaments together through a half-wave section of transmission line. By shortening this transmission line we can cause it to have a reactance which will nullify the inductive reactance of the tube leads. The voltage distribution along the half-wave line connecting the filaments of a push-pull stage is shown in Fig. 17.

A serious problem is often presented on the ultra-high frequencies by the necessity of operating several tubes in parallel, to secure the desired power. Half-wave lines provide a satisfactory means of accomplishing this. One of the tubes is connected directly to the tuned circuit to be used. The other tubes are placed in a row and connected together and to the first tube by means of half-wave lines which are subsequently tuned to balance out the reactance of the tube elements and tube leads. The operation is then almost as if the tubes were connected internally, without any leads whatsoever.

One interesting property of transmission lines is the ability of a quarter-wave section of line to act as an impedance transformer. This property plays an important part in the television antenna system and in the television picture transmitter. Since the impedance transformer may find occasional use in voice transmitters, we will describe it briefly here. You will recall in our discussion of insulators and short circuits that if we opened one end of a quarter-wave line, the other end appeared to the circuit as a short. When we closed one end of the quarter-wave line, the opposite end appeared as an open circuit. For terminations other than a short circuit or an open circuit, the line transforms to intermediate values of resistance in such a way that the line is always the mean between the sending end and the receiving end impedances. Thus if we have a line whose characteristic impedance is 100 ohms, and we terminate a quarter-wave section of this line in ten ohms, then the opposite end of the line will appear to the circuit as a 1,000 ohm resistor. This is represented mathematically as follows:

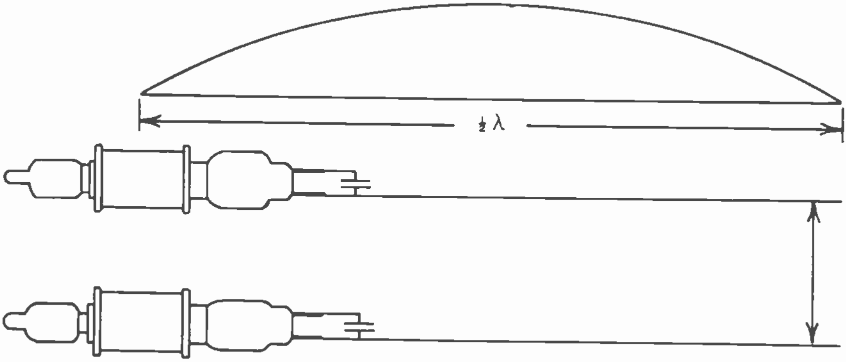


Fig.17 Use of half-wavelength interconnecting line to eliminate the effect of filament lead inductance.

For a line of length = $\lambda/4$:

$$\frac{Z_{in}}{Z_o} = \frac{Z_o}{Z_{out}} \quad \text{or} \quad Z_{in} = \frac{Z_o^2}{Z_{out}}$$

Z_o = Characteristic impedance of line (surge impedance).
Z_{out} = Output or terminal impedance of line.
Z_{in} = Impedance reflected at input of line.

Occasionally this type of line might be used to match from the impedance of an antenna transmission line to the plates of the R.F. amplifier. In this case, the line would replace the tank circuit and it could be considered either as a tank circuit or as an impedance transformer.

Another property of a transmission line which is sometimes used in the voice transmitter is that of a phase inverter. When a voltage is applied to a transmission line, it takes time for it to travel from one point on the line to another. For a half-wave section of the line, it takes the time equivalent to one-half wavelength, or 180°. When we apply a R.F. voltage to a half-wave transmission line, there appears at the opposite end an equivalent voltage which is 180° out of phase. This useful property permits us to use a half-wave section of the line wherever it is desirable to rotate the phase of the voltages. In the transmitter illustrated in Fig. 1, such a line is used to couple from the grid of one oscillator tube to the grid of the other oscillator tube. One of these tubes is also supplied with voltage from the large concentric tank circuit which acts as a frequency control. The half-wave line then serves to invert the phase of this voltage and apply it to the opposite tube so that a push-pull oscillator is controlled from a single ended tank circuit.

8. COMPONENT PARTS. At the present time, few transmitting parts are rated for ultra-high frequency application, and this makes the selection of parts for a television voice transmitter an arduous task. This is further complicated by the fact that three mutually opposed situations exist. First, it is desirable to use components of an exceedingly small size to reduce standing waves and inductance effects. Second, the size of apparatus generally increases as the power handling capability is increased. Third, the power handling capability of most apparatus decreases as the frequency increases. Therefore, our high frequency application requires components of high power handling capability, but unfortunately, such apparatus is generally exceedingly bulky and, therefore, undesirable. Such is the dilemma of ultra-high frequency transmitter design. Let us consider a few of the component parts specifically.

The R.F. bypass or coupling condensers present a serious problem, especially for high power work. Due to the ultra-high frequency and the unavoidable tube and minimum circuit capacity, high R.F. currents are encountered, therefore the bypass or coupling condensers may be called upon to carry 30 or 40 amperes. Furthermore, since the circuit and tube capacities present a fairly low reactance, the condensers must carry this current without introducing appreciable reactance into the circuit. Specifically, the reactance should not exceed 15 or 20 ohms for most applications and this calls for a capacity of several hundred micro-microfarads, even at television frequencies. At 50 megacycles, a 500 mmfd. condenser would have a reactance of 6.4 ohms, provided that the inductance of the condenser plates and the associated leads could be kept to a negligible value. For a given capacity, the voltage drop across a condenser goes down as the frequency goes up. On the other hand, the dielectric loss factor increases with frequency. In general, the result is that the current-carrying capacity of condensers are considerably lower for television frequencies than for ordinary broadcast frequencies. For transmitters on the order of 1 kw., this is no particular problem except one of cost. Standard mica condensers are available which will carry several amperes at 50 mc. Their physical size, 3" to 6" cube, is reasonable and the cost generally runs between 5 and 30 dollars. For a high powered transmitter, bypass condensers must be selected with more care in order to avoid trouble due to their larger size.

The subject of tuning condensers is not extremely important because they are seldom used in high frequency transmitters of 1 kw. or over. In large transmitters, the bulk of the tuning process is accomplished by adjusting the length of the resonant line tank circuits in conjunction with the unavoidable fixed tube and stray capacities. In this case, a trimmer may be used for fine tuning adjustment, consisting of two parallel disc plates, one of which is controlled with a micrometer type of screw adjustment. These plates are mounted directly upon the resonant line tank circuit without the use of insulators and are somewhat similar in appearance to the disc type neutralizing condensers widely adopted by amateurs.

In lower power transmitters where it is sometimes desirable, for compactness and ease of adjustment, to use ordinary coil and condenser type of tanks, the condensers are chosen with several points in mind. First, they must have an exceedingly low minimum capacity and their maximum capacity should not be greater than necessary for ease of tuning and coil adjustments. For split-stator condensers used in push-pull operation, 30 mmfd. per section is usually ample. Secondly, these condensers should be of the low-loss variety. There should be a minimum quantity of high quality insulating material, so arranged as to give a moderately long leakage path. The plates should have well rounded corners or edges and the connecting lugs to the plate assembly should be short and heavy. The wiping contacts that carry current to the rotor should have sufficient area to carry the high currents encountered at these frequencies. They must not be long and they must make positive connection to the rotor. A third consideration is the design of the framework and the method of applying current to the rotor structure. This should be accomplished without having closed loops in the supporting mechanism of the condenser frame which might resonate near the operating frequency. The connection to the rotor is preferably made at the center, in order that the use of the condenser in a push-pull stage will not unbalance the circuit.

Neutralizing condensers deserve careful consideration both as to their placement in the circuit and to their mechanical design. The most important points are to keep the leads extremely short, and to preserve balance and symmetry in the case of a push-pull stage. The function of the neutralizing condensers is to form a balanced bridge in such a way that no energy feeds through from the grid circuit to the plate circuit. In a push-pull stage, two arms of this bridge are already formed by the grid-plate capacity of the amplifier tubes. Now, if the leads to the neutralizing condensers are of appreciable lengths, then the other arms of the bridge formed by the neutralizing condensers consist of both capacity and inductance. The neutralizing condensers must than be adjusted to balance out the reactance of their connecting leads as well as to compensate for the capacitance in the tubes themselves. Since the reactance of the connecting leads increases with frequency while the reactance of the neutralizing condensers decrease with an increase in frequency, it is obvious that the bridge is balanced only for one specific frequency. When the lead lengths become appreciable, we may neutralize an amplifier with the plate voltage off to prevent feed-through of energy from the grid circuit to the plate circuit; yet, when we apply the plate voltage, the amplifier will oscillate on some other frequency where the bridge neutralizing circuit is no longer balanced. In low power transmitters, good neutralization is obtained by the use of the smallest possible neutralizing condensers, placed as close as possible to the tubes themselves. In the case of high powered transmitters, small neutralizing condensers and short leads are not possible. Even when considerable care is taken in the layout of the circuit it often becomes necessary to heavily load the amplifier, by tight coupling to the antenna circuit, in order to prevent self-oscillation. In the large water-cooled tube illustrated in Fig. 5 (type 899), the grid lead inside the tube is of considerable length. Fortunately, it is of sufficient size to allow an expedient which greatly improves the neutralization of this particular tube. It has been found that a cylinder of metal slipped over the glass envelope that contains the grid lead can be adjusted until the capacity to the grid lead itself, in conjunction with this cylinder, forms the neutralizing condenser. When two of these tubes are operated in push-pull and are placed close

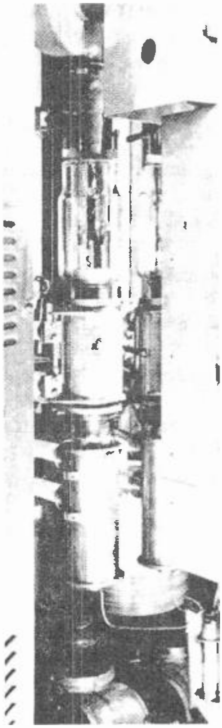


Fig.18 The capacity between the internal grid lead and a sleeve enclosing the end of the tube forms a convenient neutralizing condenser for a type 899 tube.

together, it is easy to see that the criss-cross neutralizing leads connecting opposite plate and grid neutralizing sleeves are exceedingly short. This forms a very satisfactory means of neutralization, considering the size of the tubes involved. One of these neutralizing condensers is illustrated in Fig. 18.

In order to mount the various components, it becomes necessary to use a certain number of insulators. These are used as sparingly as possible, and wherever losses are important, they are placed at a point of low voltage and out of the intense radio frequency field. Rigidity of construction is frequently more important than eliminating the last vestige of power loss, and when insulators are used, they are chosen with due regard to their mechanical strength and loss characteristics. Since the dielectric constant of insulating material is considerably greater than unity and since there is considerable R.F. voltage developed across the insulators there results a R.F. current flowing through them. This current produces heating in the insulators due to the dielectric losses of the material. In medium power transmitters on the order of one to several kilowatts, even good Isolantite insulators run warm to the touch. Since the capacity across the insulator is constant and the dielectric of the insulator is constant, the current flow, and, consequently, the losses produced increase as the carrier frequency is increased. Thus our television voice transmitter, which operates in the vicinity of 50 megacycles or higher, is subject to considerable insulator loss.

Bakelite, which is a fairly good insulator on the lower frequencies is entirely unsuited to high frequency application. Bakelite insulators used in spacing open wire transmission lines carrying only 1 kilowatt will heat to the point of blistering and, occasionally, catch on fire. Hard rubber is much superior, and certain types of special hard rubber have very low-loss characteristics. Aluminized hard rubber, commercially known as X2B, is excellent. Forms of good ceramic insulators, such as Isolantite, Steatite, etc., are quite satisfactory from the standpoint of strength, as well as having good insulating properties. Such insulators are the most common in use at these frequencies. It is practically impossible to machine these ceramics, and they must be molded and fired in their final shape. For certain applications, other types of insulating material would be selected. Micalex is fairly satisfactory and can be machined, but it is quite brittle. Victron, Amphenol, and other phenolic compounds exhibit extremely low-loss characteristics on the ultra-high frequencies. They are very easy to machine, but unfortunately they are brittle; also they become soft and pliable when subjected to heat. The losses and heat developed from various causes in a transmitter prohibit the use of these materials for most applications.

In high power transmitters, even Isolantite and ceramic insulators must be selected with care. It is necessary to avoid concentrating the electrostatic flux at any point in an insulator in order to prevent local heating. For instance if a metal screw projects into the body of an insulator, the largest R.F. currents in the insulator will emanate from the point of the screw. The heat produced at this point in the insulator may cause the insulator to shatter. This can be avoided to some extent by the use of corona shields in the form of caps into which the insulator sets. These tend to divert the current uniformly around the outside of the insulator.

9. MEASUREMENTS AND ADJUSTMENT TECHNIQUE. To successfully adjust and operate our sound transmitter, we must be able to measure the power input and output as well as the operating frequency.

To obtain our approximate frequency, we may use one of the commercially available wavemeters, or we may resonate a section of transmission line and observe the distances between the voltage maximums or minimums. If the line is loss-free, the current will travel on the wire with the velocity of light and we can calculate our frequency or wavelength from the following:

$$\lambda = \frac{2l}{39.37} \text{ Meters}$$

$$\text{Or:} \quad f = \frac{300}{\lambda} \text{ Mc.} = 300 \times \frac{39.37}{2l} \text{ Megacycles}$$

Where: l = distance in inches between successive voltage maximums or minimums.

In using the transmission line or letcher system, we can tune it by adjusting the position of a short-circuiting bar near one end, shown as S in Fig. 19. The distance from this bar, which will be a voltage minimum to the next voltage minimum, should not be taken as an indication of the operating frequency, because the residual inductance of the short-circuiting bar causes this first section of line to be somewhat shorter than our frequency would indicate. To avoid these end effects we should measure the distance between voltage maximums or minimums.

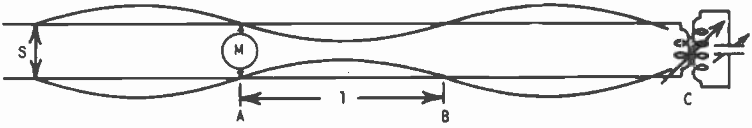


Fig.19 Letcher wire measurement of operating wavelength.

A thermo-galvanometer (M of Fig. 19) is used as a low range voltmeter and moved along the line until a point of minimum voltage is indicated. As the voltage decreases, the coupling at C can be increased, raising the voltage on the line and increasing the effective sensitivity of the voltmeter. After one voltage minimum, such as A, has been determined, the voltmeter should be moved to the next minimum, B, and the distance between these points entered as l in the formula above. If the letcher system is sufficiently long to obtain several nulls, the distances between them can be averaged to obtain an l of greater accuracy.

Having determined the approximate frequency, we must resort to a heterodyne type of frequency meter for an accurate indication. At the present time, the television frequency deviation, for both the sound and the picture carriers should be held to within .05%. To maintain this stability, they should be continuously monitored with a frequency meter whose accuracy is at least twice as good as the frequency tolerance; that is, the frequency meter should be able to measure to within .025% of the indicated frequency. At least two commercial frequency meters are available which will meet these requirements.

The type 620A Heterodyne Frequency Meter, manufactured by the General Radio Company can be used to measure frequencies up to 300 megacycles with an accuracy of .01%. It consists of a calibrated oscillator which beats with the received signal in a triode detector. The audible output is supplied through one stage of audio amplification. The oscillator is continuously calibrated by a one megacycle crystal oscillator of high stability which generates multiple harmonics in such a way that a beat is obtained on every one megacycle division on the main dial.

The new type 303A RCA Frequency Limit Monitor is suitable for frequencies up to 45 megacycles, with an accuracy of .005%. It is entirely possible that this could be adapted for higher frequency use.

Radio frequency currents in the television band can be measured with ordinary R.F. ammeters when certain correction factors are applied. The thermocouples incorporated in these meters are quite short and reasonably straight so that the current flowing in the thermocouple is uniform. The appreciable diameter of the thermocouple and its low resistance (in the case of the higher range meters) causes the thermocouple to exhibit skin effect which we have already discussed. The skin effect increases the resistance of the thermocouple for high frequencies so that more power is dissipated in the instrument for a given current. Since the meter is, in reality, a power measuring device, this causes the meter to read high. In Fig. 20 we see what corrections must be applied for a particular meter operating at a particular frequency.

The power input to a high frequency transmitter can be measured as in low frequency practice. Power output measurements may be made in three ways. First, we may obtain a non-inductive resistor which we substitute for the antenna load and measure the radio frequency current flowing in this resistor. Second, we may take any load available, such as an antenna system, or a dummy antenna; then after tuning to resonance, measure the input resistance of the system. The process of resonating the antenna system as well as the methods of RF impedance and resistance measurements will be described in the lesson on television antenna systems. The antenna can then be used as a transmitter load in conjunction with a current meter to obtain the output power of the transmitter. The third way that power measurements may be made is by the use of a group of electric lamps. The light output is measured by a photocell. This system can be calibrated on DC and applied with fair accuracy to the high frequencies. If a photocell is not available, a similar bank of lights may be supplied with 60-cycle current and the brilliance of the two light banks compared visually. The non-uniform current distribution in the lamps due to their appreciable inductance causes some error in either of these methods.

The first method using carbon resistors is generally applied to low power installation on the order of a few watts. Where high power is involved it becomes increasingly difficult to secure a non-inductive resistor. A resistor capable of high dissipation has recently been manufactured by the International Resistor Company which consists of a metallized coating on a glass tube an inch or more in diameter and about 10 inches long. This resistor may be mounted in a jacket and water-cooled. A group of these water-cooled resistors will form a load for a transmitter of several kilowatts and the power output of the transmitter may be measured either by measuring the current flowing in the resistor or by observing the temperature rise and volume of the water that circulates through the jacket.

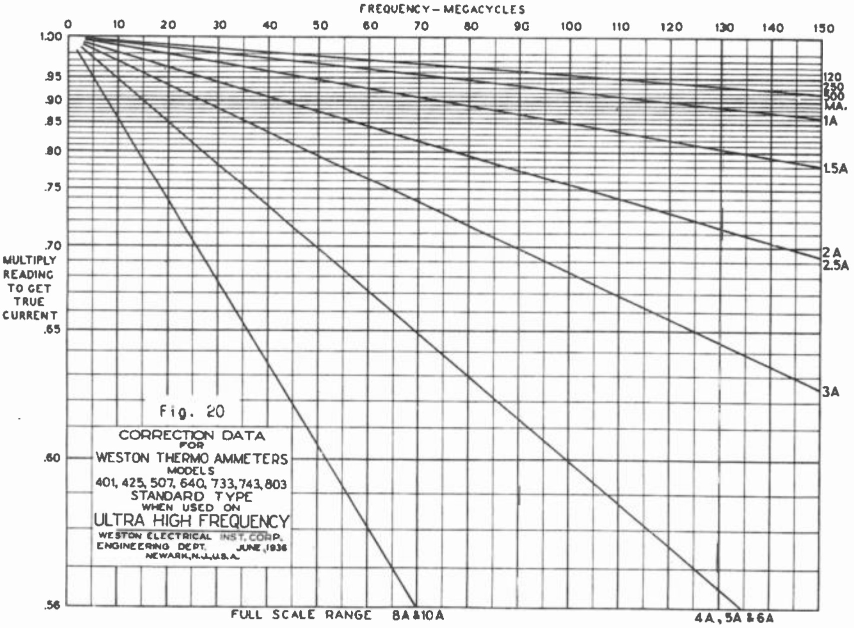


Fig. 20

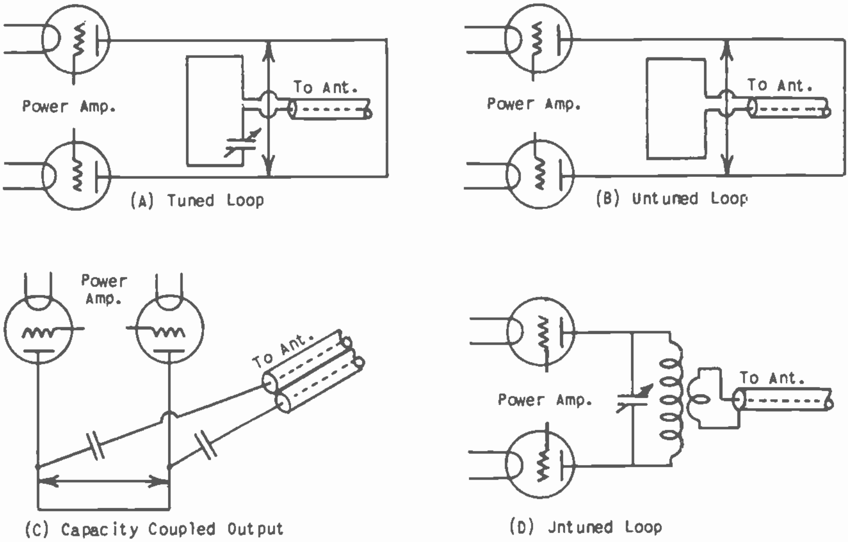
CORRECTION DATA
FOR
WESTON THERMO AMMETERS
MODELS
401, 425, 507, 640, 733, 743, 803
STANDARD TYPE
WHEN USED ON
ULTRA HIGH FREQUENCY
WESTON ELECTRICAL INST. CORP.
ENGINEERING DEPT. JUNE, 1936
NEWARK, N.J., U.S.A.

The adjustment of a television sound transmitter is carried n in exactly the same manner as the adjustment of a low frequency ransmitter. Parallel line tank circuits are tuned by varying the osition of the short circuiting slider until resonance is obtained. eutralizing condensers are adjusted for minimum feed through of nergy between the grid circuit and the plate circuit of a stage hen the plate voltage is removed. As in lower frequency operation, e can do this either with indicating lamps or thermogalvanometers, or we can adjust the neutralizing by observing the dip in the grid meter as the plate tank is tuned through resonance. One precaution, which is more important for high frequency transmitters than for transmitters operated on low frequencies, is the detuning effect of the neutralizing adjustments on the grid and plate tank. When we neutralize the stage, we must vary the neutralizing condenser in small steps and continually readjust the grid and plate tanks to resonance. Variations in hand capacity or altering the positions of adjacent parts must be guarded against, due to the increased effect of these capacity changes on the high frequencies. When the coupling between stages is by the use of direct mutual coupling between the plate coil of one stage and the grid coil of the next, or when the stages are coupled together by the close proximity of the transmission line tank circuits, very efficient transfer can be had with moderately loose coupling. However, in the low power stages where sections of a solid dielectric transmission line are used in conjunction with link coupling, considerable loss will take place in the coupling lines unless the coupling loops are adjusted until standing waves in the transmission lines are avoided.

In the case of line controlled oscillators, there is one major point to consider in their adjustment. The oscillator tubes should be as lightly loaded and as loosely coupled to the frequency controlling tank circuit as is permissible. This was covered under the section on tank circuits with particular reference to frequency control.

After the transmitter is completely adjusted, there still remains the problem of transferring the power output to the antenna. Current practice calls for the use of a *terminated transmission line* to feed the power output of a transmitter to the antenna system. Regardless of the type of antenna used, the impedance reflected at the transmitter will be the characteristic impedance of the line. We could remove the transmission line from the output of the transmitter and replace it with an equivalent resistor without affecting, in any way, the operation of the transmitter. Our problem then is simply one of developing the greatest power output from the transmitter into a certain value of resistance, usually 70 ohms. At the present time, no special means are incorporated to suppress the harmonic output of the transmitter other than good design in the transmitter itself and in the tuning of the antenna system. The simplest type of output coupling may be used. Usually, we either tap off the output circuit through condensers a few inches from the short circuiting bar of the line tank or simply couple to the tank with a single turn loop. Occasionally this coupling loop is tuned. These methods are illustrated in Fig. 21.

Before dismissing the subject of adjustment technique we should add a word of caution about parasitic oscillations. In our previous study of transmitters, parasitics were often eliminated by the use of parasitic chokes and resistances. These components cannot be used in television voice transmitters. The frequencies at which these transmitters are operated are not greatly removed from the limiting frequency of the tube and the frequencies at which the parasitics occur. Therefore, when parasitics develop, any attempt to remove them by the use of chokes and resistors would greatly lower the efficiency of operation at the carrier frequency. The only cure for high frequency parasitics is altering the circuit in such a way as to remove them or by extreme care in the design of the transmitter in the first place. Electron oscillations in the tubes themselves can generally be eliminated by altering the plate or grid voltages or the driving power of the tube.



(A) Tuned Loop

(B) Untuned Loop

(C) Capacity Coupled Output

(D) Untuned Loop

Fig.21 Typical antenna coupling methods in use at ultra-high frequencies.

A frequent trouble encountered is that of resonance between the power supply leads and the by-pass condensers, or between the filaments of the tubes and the by-pass condensers. Altering the length of these leads or changing the capacity of these condensers changes the tuning of these circuits, or we may remove unnecessary bypass condensers. Occasionally we are bothered with the simultaneous push-pull and parallel operation of two tubes which are intended to operate only in a push-pull manner. To test for this, you can short circuit the plates or grids of the tubes in question. If they continue to oscillate, or amplify, as the case may be, they are working in a parallel mode. This operation can be prevented by detuning the circuit through the addition of by-pass condensers or choke coils in the supply leads.

10. OTHER TYPES OF HIGH FREQUENCY TRANSMITTERS. As television comes into popular use, there will be a demand for more channels than the seven assigned between 44 and 108 megacycles. The only available portion of the radio spectrum is at the higher frequencies.

Circuit and tube limitations prevent present day *high power* ope tion above 100 megacycles, but tubes and circuits are continua improving. We would do well to consider both the advantages higher frequencies and the circuits and tubes that have been c sidered as possible sources of RF power for these frequencies. we go higher and higher in frequency, the shadowing effects of buil ings and hills becomes more noticeable. Likewise, the power sorption of intervening objects is increased. This indicates t the field delivered at the receiver is less in the case of v high frequency transmission. If the same physical size of anten is incorporated for a 50 mc. as for a 500 mc. transmitter, it c be a more complex array in the case of the 500 mc. transmitter, and therefore concentrate the energy in a useful direction so that a net gain would result by going to the higher frequency. Furthermore, the noise encountered is less as the carrier frequency is increased. By the time we have reached 600 mc., even automobile ignition interference, which may ruin reception at 60 megacycles, has little effect. When we operate on an exceedingly high frequency, the receiver can tune over a band of many megacycles with a reasonable percentage change in the constants of the tuned circuits so that the reception of a large number of channels becomes practicable. The large number of available channels on wavelengths around 1 meter or below will allow the use of numerous relay stations which may eventually become important in the chain operation of television stations.

The three best known ways of securing R.F. power on the centimeter wavelengths are the use of specially designed triodes and pentodes operated in conventional circuits, the Barkhausen Kurz electron oscillator, and the split-anode negative resistance magnetron. Of these three, the first and last mentioned seem to have the best possibilities.

Small push-pull pentodes have been constructed which will give a power output of 10 watts at 150 mc. These tubes have two pairs of elements in one glass envelope and the filament center tap is connected to a metal sheet and to the suppressor. The screens are connected to another metal sheet which forms the screen by-pass in conjunction with the closely spaced filament-suppressor connecting sheet. This by-pass condenser also serves as an electrostatic shield between the input and output circuits. A novel triode has also been developed in the laboratories whose elements are, essentially, a continuation of a transmission line which goes completely through the tube. The tube's leads are extended in the form of transmission line tank circuits on both ends of the tube. Thus the tube capacities and lead reactances are divided between the two tank circuits. This double-ended tube is capable of a power output of 60 watts at 300 mc. A smaller model of this type will give an output of 2 watts at 1500 mc. These particular tubes are not commercially available, but at least one American commercially-available tube (WE 316A) is capable of an output of 6 watts at a wavelength of 60 centimeters (500 mc.). The WE type 304A is capable of 30 watts at 220 mc.

The Barkhausen Kurz oscillator utilizes an ordinary triode having cylindrical grid and plate elements. The grid is operated at a highly positive potential and the plate is operated somewhat negative. Under such conditions, the electrons are accelerated rapidly toward the grid, but most of them pass through the interstices of the grid mesh. Before reaching the plate, they are turned around by the negative plate potential and attracted by the positive grid potential and may or may not strike the grid on their return path. The electrons will oscillate about the grid wires a random number of times and at random frequencies. If, however, a tuned circuit is placed between the grid and plate in the form of a parallel line tank circuit, the electrons will oscillate in the form of clouds and deliver power to the tank circuit. Such oscillators will deliver as high as 8 watts at a wavelength of 50 centimeters, but the power becomes vanishingly small at their limiting frequency of around 6 centimeters or 5,000 megacycles. The efficiency of this type of oscillator is inherently low, ranging from 1% to 7% with values on the order of 2% and 3% being typical. Furthermore, most of the dissipation occurs at the grid and there is no convenient means for cooling this electrode.

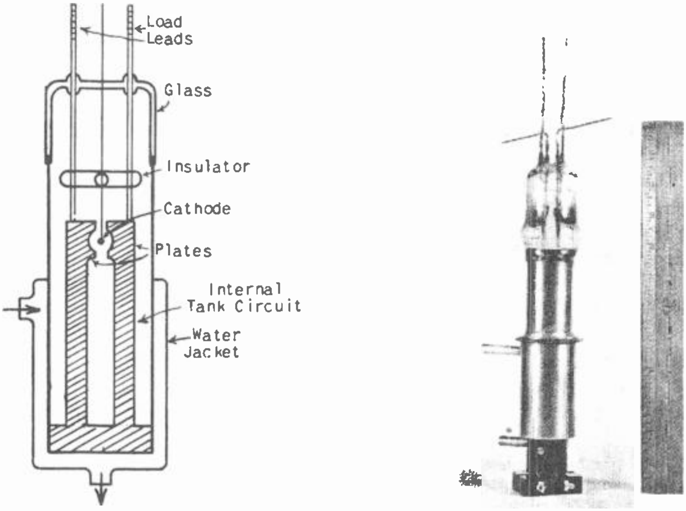


Fig.22 High power experimental magnetron. The large coil that surrounds the tube and produces the intense magnetic field is not shown in this figure.

The frequency of the Barkhausen oscillator is not particularly stable. It is influenced chiefly by the structure of the tube and the applied grid voltage, but to some extent by the tuning of the tank circuit, the plate potential, and the filament temperature.

The magnetron is another form of electron oscillator. The magnetron tube consists of a straight wire filament, coaxially surrounded by two plates formed of a split cylinder as illustrated in Fig. 22. A magnetic field is applied to the tube in such a way that the lines of force are parallel to the axis of the tube. A positive voltage is placed on the plates of the magnetron. The positive voltage tends to attract the electrons in a straight line from the filament to the plates but the magnetic field produces a force on the electrons perpendicular to their normal direction of travel. Therefore, the electrons, instead of traveling in straight lines to the plate are forced in circles about the filament. If the plate voltage is slightly increased, some of the electrons will eventually reach the plate, while if the plate voltage is lowered, the electrons will continue to circle about the filament and eventually return to the filament. A negative resistance is found to exist between the two semi-circular plates or cylinders. If a tuned circuit is placed between the plates, an electron oscillation occurs within the tube when the magnetic field is adjusted to the critical value such that some of the electrons leaving the filament just reach the plate. Under these conditions, when one plate swings positive, the electrons are attracted to the plate and energy is delivered to the load. As the plate voltage becomes more negative, the electrons are forced by the magnetic field to circle the filament and alight on the opposite plate which is now becoming positive. To maintain satisfactory oscillation, we must incline the axis of the magnetic field at an angle from 3° to 10° with respect to the axis of the cylinder. If the magnetic field is not inclined, oscillations can be maintained by placing two small plates at the ends of the main cylinder and applying a small DC voltage between them to cause an electron drift down the axis of the tube. The electrons will then travel in a spiral and be collected by the end plate of the magnetron. One or the other of these two methods must be applied in order to prevent the space charge effects of the electrons from distorting the field.

The frequency of the magnetron oscillator is determined by the shape and spacing of the electrodes, the plate voltage applied, the strength of the magnetic field, the tuning of the tank circuit, and the temperature of the filament. The magnetron oscillator is both difficult to modulate and unstable in frequency.

The efficiency of the magnetron oscillator varies inversely as the transit time. For exceedingly small magnetrons having low transit time, efficiencies on the order of 55% are possible. Since magnetrons are generally used on extremely high frequencies, efficiencies on the order of 20% to 25% are more usual. Because of the simplicity of construction, the split anode magnetron can be arranged to dissipate considerable power and, consequently, have high power output. Tubes have been built with water-cooled anodes, which give power outputs on the order of 450 watts at a wavelength of 45 centimeters. A smaller tube has been reported which would deliver 2.5 watts at 10 centimeters. This is a frequency of 3,000 megacycles. The highest frequency obtained to date by means of a vacuum tube occurred in a magnetron, operating on a wavelength of 1.1 centimeters, corresponding to a frequency of 27,000 megacycles.

Television transmitters are complex but intriguing. The art progresses rapidly and the ramifications are many. In this lesson we have endeavored to be complete. Consequently, in a lesson of this length, it is difficult to explain each circuit and mode of operation from the beginning. A certain knowledge on the part of the student has been assumed, but the background of preceding lessons has been adequate. In many cases, the explanations have been brief, but after repeated study of any point which may at first seem obscure, the student will find that his mastery of the subject has been well worth the time and effort spent.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. What type of modulation is generally employed in television sound transmitters? Is it incorporated at high level or low level? Give brief reasons for your answers.

2. Name at least four factors which are important in the design of transmitting tubes for ultra-high frequency operation.

3. What is transit time? How does it affect the ultra-high frequency performance of vacuum tubes?

4. How can we secure a good radio frequency ground?

5. Name several important considerations in the assembly and wiring of television sound transmitters employing push-pull circuits.

6. Describe briefly four types of tank circuits. Which are most desirable for high power UHF transmitters? Why?

7. Sketch one temperature compensated tank circuit useful for frequency stabilization of UHF transmitters.

8. Describe briefly several applications of transmission line sections as circuit elements.

9. What insulating materials are suitable for UHF services and what are their distinguishing characteristics?

10. (A) Sketch two or three methods of coupling the antenna transmission line to the output tank circuit. (B) What added precautions due to the high frequency are necessary in neutralizing a radio frequency amplifier of a television sound transmitter?

Power & Light Building
Kansas City, Missouri

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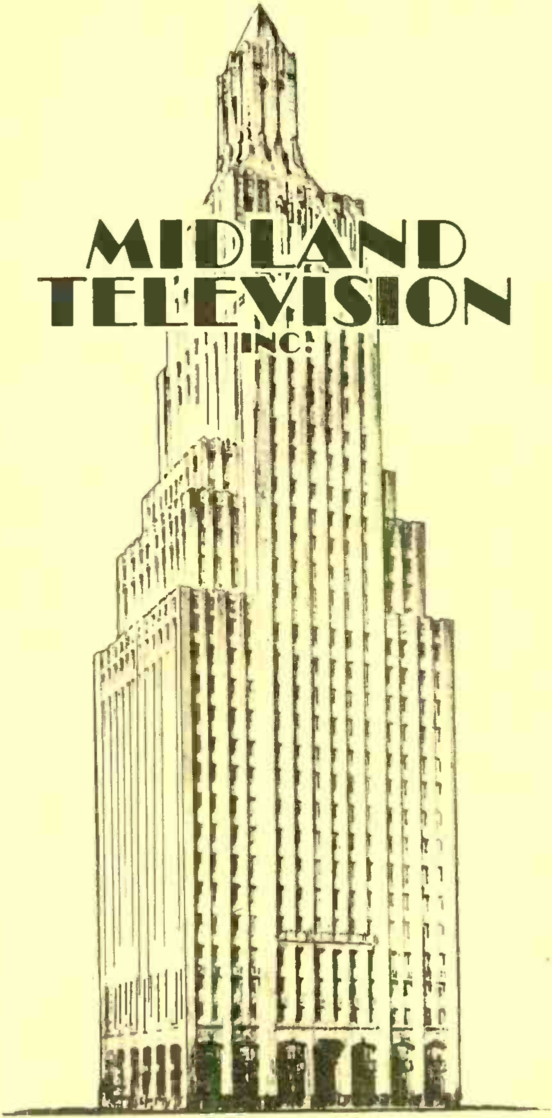

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT
NO.
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TELEVISION PICTURE TRANSMITTERS

LESSON
NO.
5

LIGHTING THE WAY

.....for human progress.

Although a variety of important inventions have completely altered the mode of living for millions of people during the last one hundred years, radio and motion pictures have been of greater benefit to the masses than any other mechanical scientific contributions we know of.

Radio brought the world to the homes of people living in remote sections of our nation and to the world. Mass education became a reality. Public health has benefitted materially from talks by medical men. People have a more intelligent understanding of governmental problems. Our states have become more closely united.

Through the medium of the motion picture, those of us who are unable to travel extensively, have become intimately acquainted with life in our own country and foreign nations. People living in the plains states have witnessed the beauty of the eastern and western mountains even though they have never visited them. The knowledge that only travel could bring in the past, is now available through the movies.

While the movies had sight and sound, radio had sound only. But that has been changed with the advent of television. Leaving out the unlimited entertainment possibilities of television, we find that the educational advantages of radio and television combined, are vast. The wedding of sight and sound in the home will contribute much to the progress of the human race now and in the future.

In contributing your knowledge and energy to the further advancement of radio and television, you will gain personal financial rewards and have the satisfaction of knowing that you have done your part in LIGHTING THE WAY FOR HUMAN PROGRESS.

Lesson Five

TELEVISION PICTURE TRANSMITTERS

"This lesson is designed to give you a complete understanding of the unusual problems encountered in the transmission of a video signal. Because of the wide band of frequencies to be handled, practices followed in broadcast procedure cannot be used in the design of a video transmitter.

"While you may never be called upon to design or build a television transmitter, still you must have a thorough knowledge of their theory of operation if you are to satisfactorily maintain and operate such equipment."

1. SALIENT FEATURES OF TELEVISION TRANSMITTERS. We have learned that to successfully reproduce a television picture having high definition and freedom from all types of distortion, it is necessary that the entire video chain from Iconoscope to Kinescope have a flat frequency characteristic with a freedom from delay distortion over a wide band of frequencies. It is this necessary requirement which prohibits the use of transmitters of ordinary design for television picture service.

In transmitters for audio service, the plate voltage, the RF driving power, the power input, and the RF load on the plate circuit are all varied in such a way that the RF stages operate with the maximum possible power output consistent with the tube ratings and reasonable plate efficiencies. Normally, these operating conditions are also varied to secure essentially linear modulation. In transmitters designed for picture service, the load impedance into which the RF tube must operate is no longer an independent variable. For a given circuit capacity, the load on the tube is a function of the desired bandwidth in a manner that we shall presently see. For high-definition television, the load on the RF stages is quite low, and a result is to lower the stage efficiency and decrease the power output in comparison with the performance that may be obtained from the same tube for aural purposes.

The modulator, as well as the RF stages, operates at reduced power output, due to the wide frequency band of the modulating voltages. In order for the modulator to faithfully reproduce its wide band video input, it must operate into a low-impedance load. This limits the modulating power and the modulating voltage that the tube will furnish. Even when the stage to be modulated requires very little power, the modulator itself must develop many times this required output when it must furnish a high voltage across the comparatively low load impedance in its plate circuit (this low load impedance being necessary to obtain the required bandwidth).

In the last lesson, we learned that for high-fidelity sound broadcasting, it was not necessary to reproduce the frequency components below 30 cycles. For television picture transmitters, it is standard in this country to transmit the DC component. Therefore, all transmitters must handle, in addition to an extremely wide band of frequencies, the slow variation which corresponds to the average background, or average illumination of the picture. This DC transmission is highly desirable for three reasons.

First, it is desirable to preserve the background information of the picture in the transmission process in order that the receiver will not necessarily have to depend upon a diode or other type of DC restoration circuit. If the background information is transmitted, the receiver can then be direct-coupled from the second detector, should this prove desirable, or certain changes may be made in the light of future inventions, without the limitations imposed by AC transmission.

A second desirable feature of DC transmission is the latitude of design allowable for the receiver synchronizing pick-off circuit. If DC transmission is not used, it is necessary to either level the video circuit before applying the synchronizing pick-off, or the receiver must possess an extremely good automatic gain control circuit (abbreviated AGC) in order that a DC-coupled type of pick-off may be successfully employed. AGC is desirable at the receiver in any case, but it may prove economically practical to eliminate it in the very cheapest receivers when DC transmission is employed.

The third reason for employing DC response at the transmitter is to increase the power output by making full use of the dynamic range of the transmitter. If direct coupling or its equivalent is employed in the video and modulator stages of the transmitter, the peak of the synchronizing will always occur at maximum output of the transmitter. If AC-coupling is employed at the transmitter, the blanking level and the tip of the synchronizing will occur at various levels of output power, depending not only on the amplitude of the picture signal involved, but also upon its character. Unlike normal sound waveforms, the waveform of the picture signal is highly asymmetrical.

Let us consider two radically different types of picture signals, such as a group of white bars upon a black background as contrasted to a group of black bars on a white background. In Fig. 1A, we have the basic diagram of a typical grid modulated transmitter. If the signal input was capacity-coupled at X, then the transmitter does not contain the DC component (background information of the picture), and the picture signal shifts about the bias axis of the tube in such a way that the area of the signal above this axis is

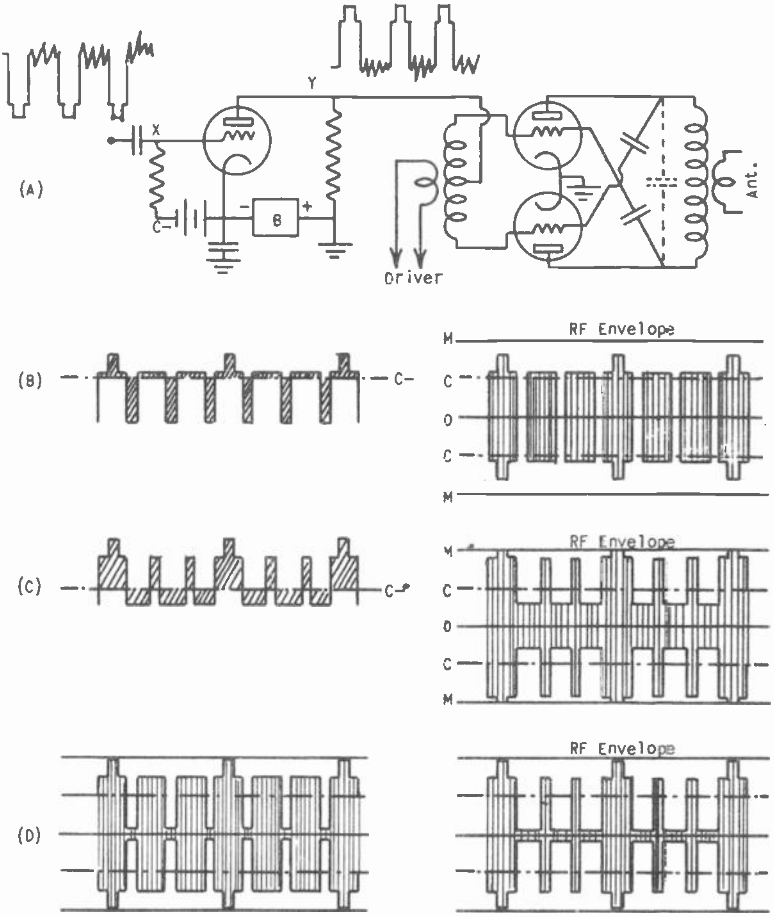

Fig.1 Effect of different types of picture signals upon the modulation capability of the transmitter.

exactly equal to the area of the signal below this axis, as shown in Fig. 1B and 1C. These two figures illustrate the different types of picture signals as they appear at point Y in the schematic diagram at the top of Fig. 1. If the value of the signal is increased until distortion of the received signal is observed, it will be found that two entirely different types of overmodulation have occurred. In the case of the white bars upon the black background of Fig. 1B, the negative peaks of modulation have reached zero output, even though the tips of the synchronizing have not yet attained the peak output limit of the transmitter. In this case, we have obviously overmodulated the transmitter without exceeding the peak power limitations. Under precisely the same operating conditions, and with an exactly equal amount of picture signal, the signal generated by a group of black bars on a white background, as illustrated in Fig. 1C, will overmodulate the transmitter in the positive direction, thus clipping the synchronizing peaks long before the negative carrier swings have reached zero output. Had the transmitter been DC-coupled, the magnitude of output voltage attained by either type of signal would have remained constant, and the signal which overmodulated the transmitter, as illustrated in Fig. 1B and 1C, would not have overmodulated the transmitter for the DC case illustrated in Fig. 1D. Obviously, then, the transmission of the DC component will result in a transmitter with greater modulating capabilities than an AC-coupled transmitter of otherwise similar design.

In spite of the poor circuit efficiency and the low power output obtainable from a specific tube for use in television picture transmission, it becomes necessary to develop relatively high power output, in some manner or other, in comparison to the power necessary for equivalent coverage under ordinary broadcast conditions.

In broadcast work, a field of 2 to 10 millivolts per meter is considered satisfactory for primary urban service. In television service, a figure of 5 millivolts per meter has been established as the necessary field for suburban New York City.

The required field for primary sound broadcasting and primary television is seen to be comparable, but it takes more power to lay down a given field strength at ultra-high frequencies than on the lower frequencies, due to the increased attenuation of the signal in passing over the open country, and particularly, in passing through buildings, over hills, and other obstructions. There are other reasons, also, for this difference, which will be discussed in the following lesson. The combined need for greater power and the increased difficulty of obtaining it on the ultra-high frequencies, results in a large and costly transmitter for a moderate coverage.

In order to transmit the wide band of modulating frequencies required to produce a high-definition picture without occupying an excessively large slice of the RF spectrum, it is necessary to suppress one sideband of the television picture transmitter. Since the receivers are normally designed to accept only the higher sideband of the transmitted signal, the lower sideband can be filtered out at the transmitter without loss of picture quality at the receivers.

2. POWER REQUIREMENTS. The detailed discussion of television coverage will be taken up in the next lesson on television transmitting antennas. At this point, however, we can consider the conclusions of that lesson in arriving at the necessary power output of the transmitter. Elaborate field tests throughout greater New York City have shown that a suburban field of 5 millivolts per meter is necessary to give a picture which is satisfactorily free from interference. In smaller cities, where less interference exists in the suburban areas, a field strength of 1 millivolt per meter will produce a satisfactory picture, and occasionally an excellent location is found where the picture reception is limited only by the sensitivity of the receiver and the noise generated in the receiving

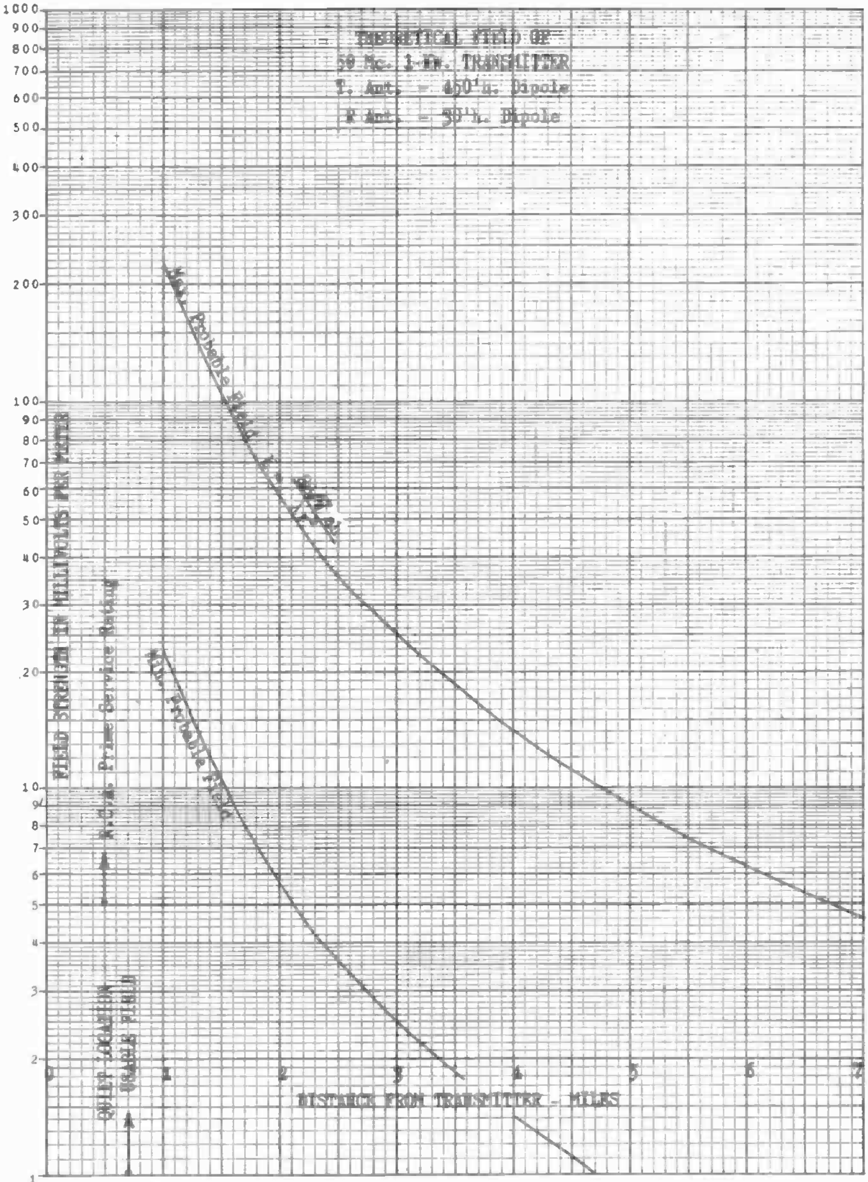

Fig.2 Probable field strength of a 1 kw. transmitter located at an elevation of 450 feet.

tubes. The latter estimate of 1 millivolt per meter for medium sized cities has not been very well substantiated, since television transmission and reception is, at present, limited to a few of the larger cities.

The field strength that a transmitter of particular power will develop depends upon a number of variables. The chief of these are the power of the transmitter, the height of the transmitting antenna, and the height of the receiving antenna, the ability of the antenna to concentrate its radiation in a useful direction, and the type of terrain over which the signal must be propagated. To obtain the approximate power necessary for a given coverage, some of the variables are eliminated by assuming that the antenna is a dipole,[1] and that the ground is level and free of obstructions. In this case, the field delivered at the receiving antenna will be given by the following formula:

$$E \text{ (volts per meter)} = \frac{88\sqrt{W}\,ah}{\lambda r^2} \qquad (1)$$

Where: W = power in watts, λ = wavelength in meters, r = distance between transmitter and receiver in meters, a = height of receiving antenna in meters, and h = height of transmitting antenna in meters. This specifies that the received field varies directly as the height of the transmitting or the receiving antenna, and that it varies as the square root of the power and as the inverse square of the distance between the transmitter and the receiver. Thus, if the height of either the receiving or transmitting antenna is doubled, the received field would be doubled. If the transmitter power is doubled, the received field would be increased by only 41%. A receiver located two miles from the transmitter would receive a field only one-quarter as strong as a receiver located one mile from the transmitter, all other things being equal.

Having obtained the expected field from a transmitter of given power by neglecting absorption and irregularities in the terrain, we can then estimate the probable field that will exist by allowing certain factors as multipliers for the calculated field. In general, the received field will run between 10% and 100% of the calculated value. On the outskirts of town and in the suburban areas, somewhat less variation occurs. In these areas, a figure of 30% to 60% is applicable. This practical reduction of the calculated value is due to the scattering effect of intervening objects and to absorption by these objects and by the terrain. Occasionally, the absorption which would tend to reduce the calculated value of field strength is offset by an unusually advantageous receiver location, such as at the crest of a hill, since this gives the receiving antenna a greater height than that used in the formula to calculate the theoretical field. Assuming that the received field will vary between 10% and 100% of the calculated value, we may then plot the maximum and minimum probable field as in Fig. 2.

Having found the maximum and minimum probable field at a certain

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[1] Simple center-fed, half-wave Hertz.

distance from the transmitter, we can estimate much closer than 10 to 1 by observing the particular site and noticing if it commands a direct line of view to the transmitter, free from any obstructions. If the path is clear and the receiver site is at the top of a hill, the probable field will lie near the maximum line on the curve. If the receiving location is lower than normal, or surrounded by buildings or other obstructions, the received signal is apt to fall near the minimum probable field of our graph.

From Fig. 2, we observe that a transmitter of 1 kw. would be satisfactory for a town, such as Kansas City, having a population of approximately half a million people, extending to a radius of about seven miles from the center of town. In this case, an antenna height of at least 450 feet is available, and satisfactory coverage could be had over most of Greater Kansas City. Even this power would not give perfect coverage, and unless special antennas were incorporated, one or two of the suburbs lying about 10 miles from town could not be adequately served. In larger cities, the suburban population would extend from 15 to 35 miles from the metropolitan center. With a transmitter height of 450 feet, the service range of 35 miles would require a transmitter power of approximately 25 kw. However, in a city of such large dimensions, a location could probably be found which would give a greater antenna height than that upon which our calculations are based. The height of the Empire State Building in New York City is approximately 2½ times the antenna height assumed in Fig. 2. Consequently, a 10 kw. transmitter located in the Empire State Building would have a service range of about 25 miles. At present, the NBC station occupying this location has a carrier of 4½ kw. which gives a satisfactory signal throughout the greater part of suburban New York City. The CBS station located in the Chrysler Building has an antenna height of approximately 950', with a corresponding service range approximately 2/3 that of the NBC transmitter in the Empire State Building, since the two transmitters are of equal power output.

3. MODULATION METHODS. In the last lesson we discussed the application of the various modulating methods for high frequency transmitters. These principles are, of course, applicable to television transmitters, but it becomes necessary to review them in order to point out the effect of the wide band frequency requirements upon the various systems contemplated. After reviewing the various possibilities, we shall select several of the most desirable methods and compare them on the basis of transmitters designed for 1 kw. output power.

It was concluded in the last lesson that absorption modulation applied to the tank circuit of a transmitter was impractical for voice transmitters, and that this system has further disadvantages for television. An absorption modulated transmitter, illustrated in Fig. 3, operates upon the principle that the power delivered to the load is the available power which the modulator does not absorb. Consider a square wave having steep sides (involving a wide band of modulating frequencies). If this wave were applied as a positive impulse to the modulator tubes, the tank circuit would be rapidly damped, and the output would quickly fall to a low value. Upon the rapid removal of the impulse, it would take a certain length of time for the tank circuit to acquire enough energy to build up to its original output, distorting the square wave in the manner shown. This is equivalent to a combination of amplitude and phase distortion of the signal. To overcome this defect, additional damping would have to be added to the tank circuit, decreasing the efficiency of the output stage to a value even lower than the already low efficiency inherent in this type of modulation.

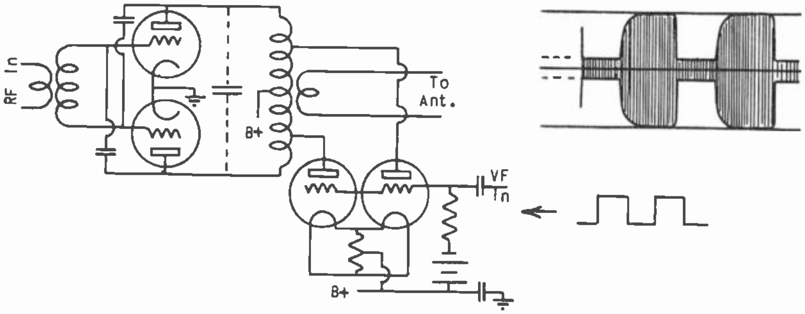


Fig.3 Absorption modulation and its effect upon signal quality.

Transmission line modulation, or load impedance modulation, illustrated in Figs. 14 and 15, removes the limitation of bandwidth from the power amplifier stage by introducing the modulation between the final tank circuit and the antenna, thus allowing the power amplifier tube to run at the maximum efficiency obtainable at the carrier frequency. The operating principle of this type of modulation will be discussed in following paragraphs. The transmitters which use this modulating method operate with an overall efficiency of approximately 25% in the high-level stages. This is very good, compared to the other types of modulation normally used for television purposes. In this type of transmitter, the modulator becomes a sideband generator, while the power amplifier tubes amplify only the carrier. By segregating these duties, it is possible to use somewhat smaller tubes than for other types of picture transmitters. The advantages of this will be recalled from the previous lesson. This type of modulation is theoretically quite linear as well as being moderately efficient, and of wide band characteristics. However, it is necessary to operate the grids of the modulators to extremely high positive voltages. The positive grid swing limits the choice of modulator tubes to those having grids capable of considerable dissipation and current handling capacity, as well as placing a low impedance upon the video drivers which the modulators are to follow. Furthermore, this low impedance is non-linear with amplitude variations so that unless the video driver has low internal impedance, some modulation non-linearity will result.

Suppressor grid and screen grid modulation have the same disadvantages for application in picture transmitters as in television voice transmitters, the chief of these being the unavailability of suitable tubes. A second disadvantage is that the video voltage swing required for 100% modulation is greater than the swing required for control grid modulation. The greater video voltage requirement results in the necessity for larger modulator tubes than in the case of control grid modulation.

For television sound transmitters, we concluded that plate modulation was most satisfactory. For television picture services, this type of modulation is difficult to achieve for two reasons. The first difficulty is the necessity for a modulation reactor which will pass the wide video frequencies involved. Plate reactors have been built which will pass a band of frequencies from 30 cycles to 2.5 megacycles. When these are used, the DC variations must be applied as series modulation which will be discussed shortly. It is doubtful that a reactor suitable for a high power transmitter could be built which would be flat from 30 cycles to 4.5 megacycles. Even if such a reactor could be built, its cost would likely be prohibitive, and the problem would still remain of introducing the low frequency into the transmitter as some other form of modulation. A resistance can be substituted for the modulation reactor if the plate voltage is run to twice its normal value. Then the resistor will dissipate a power approximately equal to that consumed by the transmitter. This immediately halves the efficiency of the high-level stage. A non-inductive resistor capable of dissipating 5 to 50 kw. is in itself somewhat of a problem.

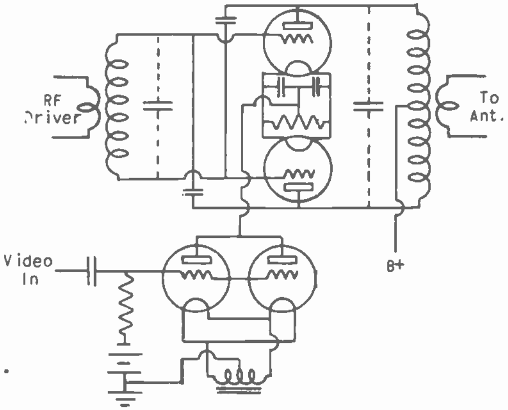


Fig. 4 Series modulation.

Series modulation of the plate supply as shown in Fig. 4 is a way out of this particular problem, and is a much-used system from the older days of television. Any system of plate modulation, however, requires a relatively high modulating voltage which is difficult to achieve at video frequencies, because of the electrostatic capacity into which the modulator must work. We will see, a little later on, that this problem is so serious as to present an apparent impasse——a power limit, beyond which it seems impossible to go with present day tubes.

Grid bias modulation requires a reasonably small video swing; in fact, smaller than any other method except transmission line modulation. A grid modulated transmitter for television services must, like the plate modulated transmitter, operate into a relatively low load impedance, because of the damping required in the plate circuit to secure the necessary bandwidth. This damping is, of course, the load reflected by the very tightly coupled antenna circuit. The result is that the plate voltages for television picture transmitters are low and the grids must be swung into the highly positive region, resulting in considerable grid current. This, of course, is drastically different from ordinary grid bias modulation in broadcast practice. Such a transmitter operates with a reasonable plate efficiency of 20%, but the linearity of modulation is not all that could be desired. The output cannot be modulated to much below 25% of the peak output without considerable curvature. Also, due to the grid current which flows at modulation peaks, some curvature of the output characteristic will occur at maximum output unless the modulators have an effectively low internal impedance. In spite of the non-linearity that results from this type of grid bias modulation and its none-too-high plate efficiency, this method is the chief system in use today. It is employed not only in several

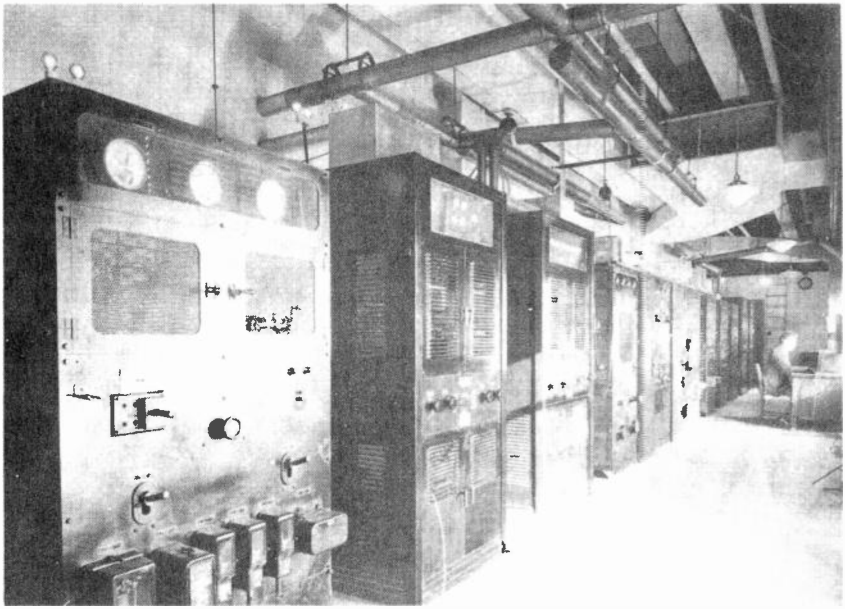


Fig.5 The N.B.C. television transmitter in the Empire State Building, New York City.

foreign transmitters, most prominent of which is the British Broadcasting station in London, but it is also employed in the NBC television transmitter located in the Empire State Building (pictured in Fig. 5), and in the Columbia Broadcasting System station diagrammed in Fig. 6. It is also the method employed in the 1 kw. commercial transmitter offered for sale by RCA and pictured in Fig. 7.

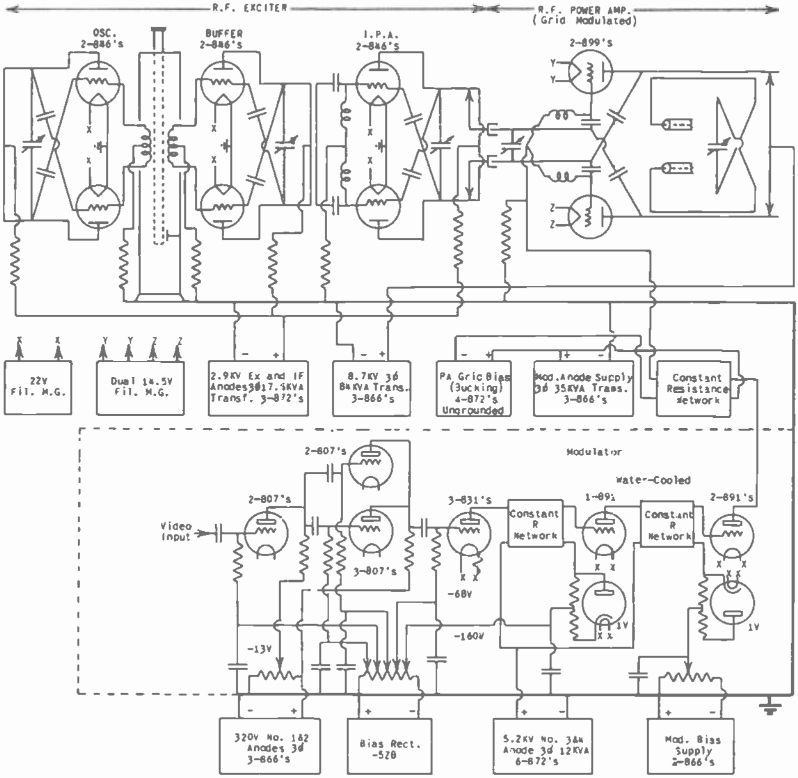


Fig.6 Basic circuit of the C.B.S. picture transmitter in the Chrysler Building.

Linear amplifiers present a way out of the modulating problem. If the modulation is introduced at a sufficiently low level, several systems become practical. The problem then remains of amplifying this signal to a level sufficient to deliver the required power to the antenna. This is the system employed by the General Electric Co. in their 10 kw. station at Schenectady. In order to secure the required bandwidth in a linear amplifier, it is necessary to heavily damp the tuned circuits and to employ overcoupling similar to the method used in IF design of television picture receivers. When high-level modulation is employed, the bandwidth of only one tuned circuit need be considered, but in the case of the linear amplifier, the attenuation of the sidebands is the product of the attenuation of the individual stages. Consequently, the individual tuned circuits of the linear amplifier must be designed for a greater bandwidth than the output tank of a high-level modulated stage, or else a succession of double-tuned and single-tuned circuits in successive stages must be carefully designed and adjusted. Under these conditions which result in low-impedance loads for the various amplifier tubes, the gain per stage is quite low and, consequently, the number of stages that must be employed may become excessive if the modulation is introduced at too low a level. The system, however, is quite practical and, undoubtedly will become more widely adopted as high power, high frequency tubes are improved from time to time.

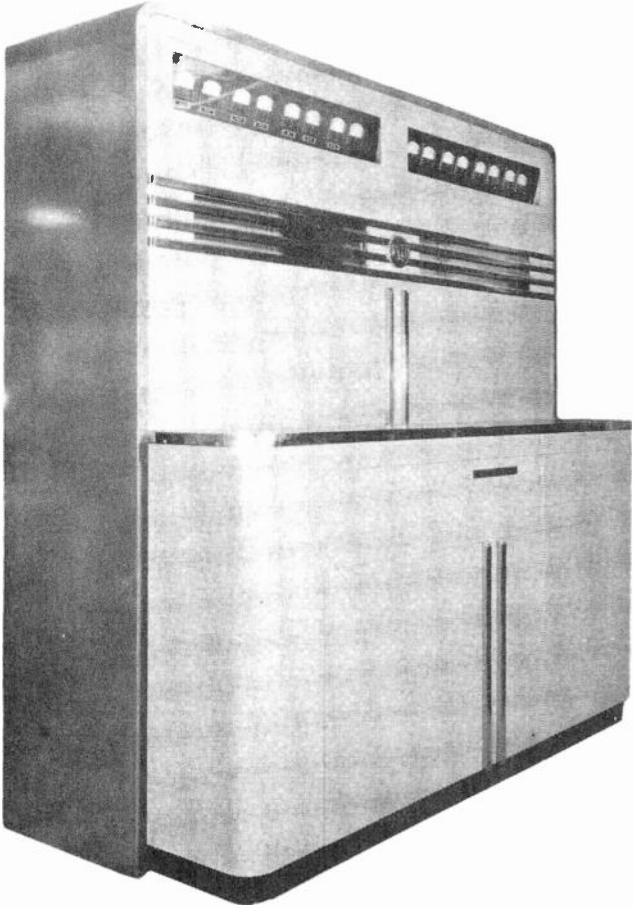


Fig.7 R.C.A. 1 Kw. Commercial Television Picture Transmitter.

4. PLATE MODULATION. For simplicity, let us consider the type of plate modulation illustrated in Fig. 8. This eliminates the necessity of a wide band modulation reactor as does the series modulation system indicated in Fig. 4. Although the resistance-coupled, parallel type of plate modulation requires something over twice the input power because of the dissipation in the modulation resistor, it is otherwise similar to series modulation in the tube complement and operating conditions for optimum performance. Many of the design problems of the plate modulated transmitter will carry over into the design of grid modulated transmitters and linear amplifiers which will be considered in following paragraphs. This is especially true of our first consideration, namely, the RF load into which the tubes must work in order to fulfill the bandwidth requirements.

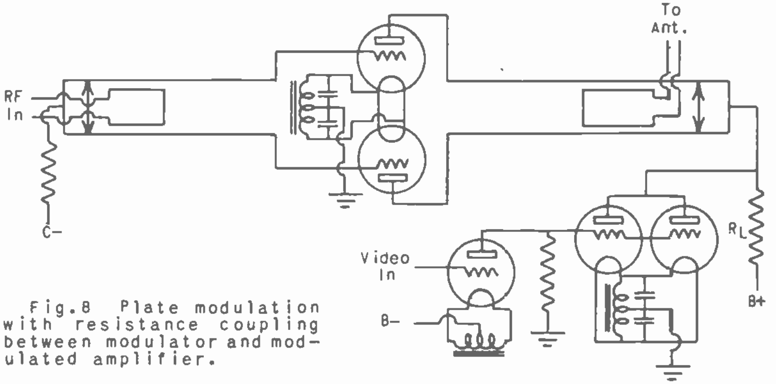


Fig.8 Plate modulation with resistance coupling between modulator and modulated amplifier.

For a simple tuned circuit, such as a tank circuit of a transmitter, the bandwidth is a direct function of Q in the following manner:

$$Q = \frac{f_o}{f_1 - f_2} \qquad (2)$$

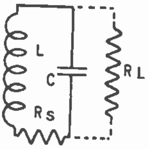
Where: f_o = carrier frequency, f_1 = the highest sideband frequency, f_2 = the lowest sideband frequency. Therefore, $f_1 - f_2$ = twice the top video frequency. When the tank Q is adjusted from the above relation and the tank circuit is fed from a high-impedance generator, the response will be down to 70% at the edges of the band. For the sake of economy in television transmitters, the loading on the tank circuit is so adjusted that the response of the amplifier will be down to 70% at the top modulating frequencies that it is desired to handle. Then, if necessary, the video amplifiers and modulators of the transmitter are peaked at the higher video frequency in order to compensate for the discrimination against the highest and lowest sidebands in the RF circuit.

If we consider that the tank circuit of a transmitter is a pure inductance, tuned only by the tube capacities and the unavoidable stray capacities of the tube, then there is a very definite load impedance reflected to the plate of the modulated amplifier of

Fig. 8. This is given by the following formula:

$$R_L = QX_C \qquad (3)$$

Where: R_L = the radio frequency load impedance into which the modulated amplifier must work, X_C = the reactance of the tank capacity at the carrier frequency, and Q is obtained from formula (2). This relation is derived as follows:



$$\frac{X_L}{R_S} = Q, \text{ but } X_L = X_C \text{ at resonance.}$$

$$\frac{X_C}{R_S} = Q \qquad \frac{X_C^2}{R_S} = R_L \qquad R_S = \frac{X_C^2}{R_L}$$

$$\frac{X_C \times R_L}{X_C^2} = Q \qquad \frac{R_L}{X_C} = Q$$

$$R_L = QX_C$$

In the practical case, the tank inductance is a section of a transmission line. If the section of line is chosen of such a characteristic impedance that it is physically only 1/8 wave long, then for practical purposes, it behaves like a pure inductance over the television band. If the line tank is 60 degrees long, then the bandwidth of the modulated stage is decreased by about 25% from the pure inductance case, unless the antenna circuit is also tuned and tightly coupled to the tank of the modulated stage.

In broadcast transmitters, it is customary to decide upon an economical value of plate supply voltage consistent with the ratings of the tubes to be used, after which the bias, excitation, and load resistance are varied to obtain the maximum power output from the tube without exceeding the dissipation limit or other ratings of the tube. For television transmitters, however, the load resistance is immediately fixed as soon as we decide upon the bandwidth desired and the minimum capacity achievable [from formula (3)]. This load impedance will be very much lower than the values we are accustomed to using for other services. If we retain the normal plate voltage and the normal excitation to the tube as indicated by other services, a terrific peak plate current would result.

The second limitation which we must consider, in addition to the load impedance on the tube, is the maximum value of instantaneous plate current that the tube will deliver without shortening its life. In the case of water-cooled tubes, this peak plate current plus the grid current, is the available emission of the tube, since saturation currents do not damage a tungsten filament. Most air-cooled tubes, however, are manufactured with thoriated tungsten, and these filaments cannot be operated at their maximum emission, since the thorium would evaporate from the surface of the filament faster than it could be boiled out from the interior of the filament, and this would damage the tube. The peak current usable from a thoriated tungsten filament depends upon the type of service involved. If the peak current occurs only at the peaks of modulation, higher values of filament emission can be used without damage to the tube than if the peaks occur at each radio frequency cycle. For modulated services, a good tube will deliver a useful peak current of approximately 35 ma. per watt of filament heating power, but the peak current for any particular tube should be arrived at through the manufacturer's rating; for example:

TYPE 833

Class C telephony rating at carrier.........Ip = 400 Ma. Max.
At peak of modulation........................Ip = 800 Ma.

Assume 150° plate current flow angle (normal).

From Fig. 11, we observe: $\frac{\text{Peak Current}}{\text{DC}} = 3.8$

Then: Peak plate current = 3.8 × 800 = 3.04 Amperes.

Check: Filament heating power = $E_f \times I_f$ = 10 × 10 = 100 watts.

Peak current per watt of heating power = $\frac{3.04}{100}$ = 30 Ma. per watt.

With the above consideration as a starting point, we shall proceed to design a 1 kw. plate modulated transmitter, in order to illustrate the application of these principles. Many features of television picture transmitters will be shown which are not in common with any other type of radio transmitter work.

If the peaks of modulation are not to exceed the tube's rating, our design must be based upon this instantaneous condition. From a preceding calculation, we arrived at a peak current of 3 amperes for the 833. For simplicity, we will neglect grid current, as well as the calculations necessary to arrive at the grid driving power and grid bias. From the plate characteristics of the type 833, illustrated in Fig. 9, we draw a line to represent the maximum usable emission at 3 amperes. We draw another line, called the diode line, which represents the condition for equal grid voltage and equal plate voltage. Normally, we operate the tube so that at the peaks of modulation, the minimum plate voltage is just somewhat greater than the instantaneous grid voltage. In other words, no point on our load line should fall to the left of the diode line for the peak of the modulation cycle. Likewise, no point on our load line should fall above the line representing maximum usable emission. If the load line falls to the left of the diode line, excessive grid driving power will be required, and large grid currents will result with possible overheating of the grids. If the load line crosses the line representing maximum usable emission, the tube's life will be materially reduced.

To obtain the slope of the load line, we proceed as by formulas (2) and (3) to obtain the RF load into which the modulated amplifier must work. To do this, we sum the minimum capacities across the tank circuit as follows:

Allow 10 mmfd. for wiring.............10
C_{pk}.............. 8.5
$2C_{pg}$ = 2 × 6.3.............12.6
31.1 mmfd.

Assume a carrier of 51.25 mc.

Then:
$$X_C = \frac{1}{2\pi fC} = \frac{10^6}{2\pi \times 51.25 \times 31.1}$$

$$X_C = 100 \text{ ohms.}$$

Top video frequency = 4.25 mc.

$$(2) \qquad Q = \frac{51.25}{8.5} = 6$$

$$(3) \qquad R_L = QX_C = 6 \times 100 = 600$$

The factor, 2 × grid to plate capacity, comes about due to the necessity of neutralizing. This is true because the neutralizing capacity

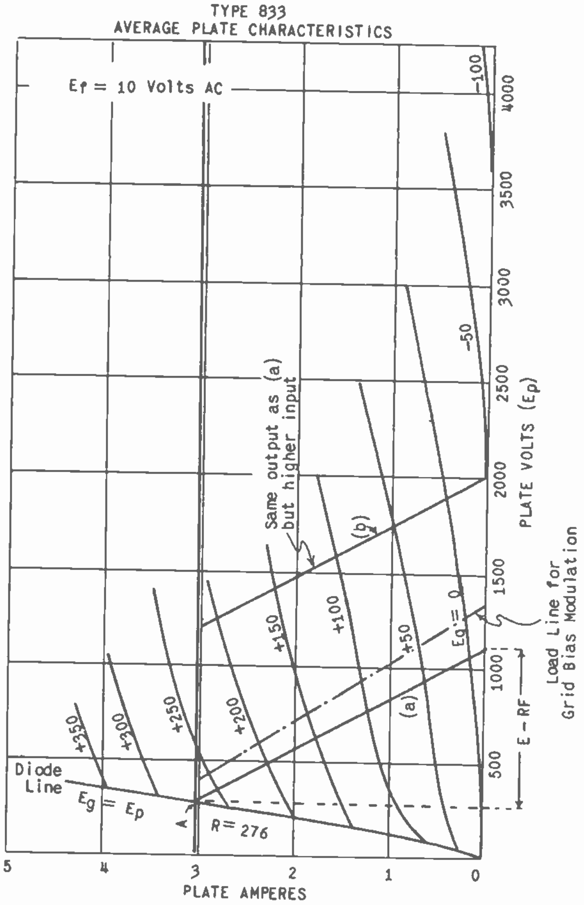


Fig.9 Application of plate characteristics of vacuum tubes in picture transmitter design.

approximately equals the grid-plate capacity of the tube and for any type of capacity neutralization, it can be shown that the neutralizing condenser is effectively across the tank circuit, as far as bandwidth is concerned.

Let us consider the case of plate feedback to obtain neutralizing voltage, as shown in Fig. 10A. The equivalent circuit of the tank could be shown as in Fig. 10B, since the grid-plate capacity is in series with the neutralizing condenser across the tank. Since the impedance into which the tube looks is only one-quarter of the total tank impedance, the load on the tube could be replaced by the equivalent circuit of Fig. 10C. Consequently, as far as bandwidth is concerned, the tank acts as if twice the grid-plate capacity appeared across it.

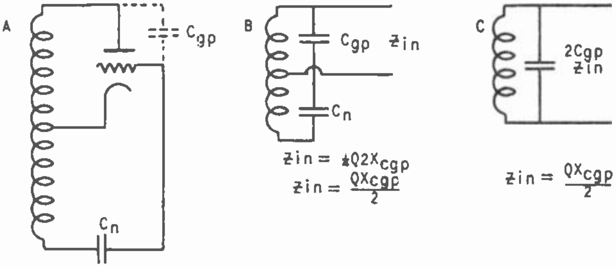


Fig.10 the effect of the neutralizing condenser upon the bandwidth and load impedance of a modulated amplifier.

From the foregoing calculation, the reactance of the capacity branch was found to be 100 ohms per tube and, choosing a Q of 6, the load resistance per tube is 600 ohms. We arrived at a Q of 6 by choosing a top modulation frequency of 4.25 mc. which made the widest sidebands fall 4.25 mc. above and 4.25 mc. below the carrier frequency.

For audio applications, the load line is plotted directly upon the plate characteristic as you learned in Lesson-6, Unit 3. This cannot be done in RF work because the load is a tuned circuit and the value of 600 ohms, which we calculated, applies only to the fundamental component of the RF current delivered by the modulated amplifier. Since the tube is operating for less than half of each cycle, the output current contains many harmonics. It is assumed that all the harmonics are shorted to ground through the tank circuit. For rigorous analysis, it would be necessary to use a point-by-point method of calculation which consists of entering on the tube characteristics an assumed value of RF sine wave plate voltage to obtain the plate current that the tube will deliver. From these values which are selected to conform with the known condition of plate load, it is possible to accurately predict the performance of a tube. This method is beyond the scope of this lesson, but a good approximation is obtained as follows.

At the peak condition of modulation, the plate current of a class C amplifier normally flows for approximately 150°. The ratio of the peak of the current wave to the fundamental component will not vary greatly for various shapes of current pulses, even when the wave is considerably distorted. Consequently, we can obtain the ratio of peak value of the wave to the peak value of an equivalent wave of fundamental frequency; that is, one which is entirely free of harmonics. We obtain this from Fig. 11. For the condition of 150° flow angle, we note that the ratio of peak current of the plate pulse to the peak of a fundamental wave is approximately 2.2. Hence, we can plot a load line for the distorted current pulse by dividing the calculated load (into which the fundamental component of plate current flows) of 600 ohms by the factor 2.2. The tube will behave, as far as the distorted current wave is concerned, approximately as if it were working into a load of 276 ohms. The next step in calculating the performance of the 833 is to draw in this new load line from point A of Fig. 9. Actually, the load line is not quite straight as shown, but an exact analysis is not necessary for the purpose of this lesson. The power output of the tube will be:[1]

$$W_{out} = E \times I$$

Where E and I are RMS values of carrier frequency sine waves.

$$W_{out} = \frac{E_M \times I_M}{2}$$

Where E_M and I_M are crest values of the fundamental radio frequency plate components. E_M is obtained from the total plate voltage swing of Fig. 9. I_M is obtained by dividing the distorted current wave of 3 amperes from Fig. 9 by the form factor obtained from Fig. 11 for 150° flow angle.

$$W_{out} = \frac{(1120-280)\left(\frac{3}{2.2}\right)}{2} = 572 \text{ watts.}$$

The input will be the product of the DC plate voltage times the DC plate current as follows: W_{in} = (.8 amperes DC at peak of modulation) × (1120V DC supply voltage from Fig. 9) = 895 watts. These conditions are for the peak of modulation. If the transmitter is adjusted for linear modulation, the values at carrier condition will be one-quarter of the peak values, or as follows:

$$W_{out} = \frac{572}{4} = 143 \text{ watts.}$$

$$W_{in} = \frac{895}{4} = 224 \text{ watts.}$$

$$\text{Efficiency} = \frac{143}{224} = 63.8\%$$ (tube only—not counting circuit losses)

$$\text{Dissipation} = 224-143 = 81 \text{ watts.}$$

A 1 kw. transmitter designed to use 833's in the final amplifier would require eight of these, and it is doubtful if eight tubes could

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[1] Review Lesson 6, Unit 3, for a study of load lines.

be made to perform satisfactorily in push-pull-parallel, at television frequencies. We cannot obtain greater output by increasing the plate voltage, since to do this would only move the load line of Fig. 9 to the right. Since the power output is the product of the AC voltage and AC current swing, which remains constant regardless of the position of the load line, it is impossible to increase the output without altering the slope of the load line. The slope of the load line, of course, is fixed by the bandwidth requirements. Moving the load line to the right, however, does decrease the amount of excitation required, but increases the plate dissipation and lowers the efficiency very rapidly. If we increase the plate voltage of a tube by moving the load line to the right, we increase the input to the tube without changing the output, which accounts for the rapid rise in plate dissipation at higher plate voltages. For television services, the peak emission of the tube is usually reached before the dissipation rating of the tube is exceeded because of the low value of load impedance. When this is true, the output is limited by two things, namely; the peak emission that the tube can stand; and the minimum capacity that can be had across the tuned circuit for a particular bandwidth. Therefore, the output of a RF stage can be calculated approximately from the following formula:

$$P = \frac{113\ I_b{}^2}{f_m C} \qquad (4)$$

Where: P = output in kw., I_b = maximum plate current rating for class C telephony, C = minimum tank capacity in mmfd, and f_m = top video frequency. (150° flow angle is assumed). Check this formula against the calculations on the preceding pages for the performance of the type 833.

Applying this formula on the basis of 1 kw., it is seen that no pair of tubes smaller than 858's (20 kw., dissipation) or 889's (5 kw. dissipation) will give the required output. For broadcast services, these tubes in a plate modulated stage would give a power output of 16 kw.

While it is difficult to achieve sufficient power output in a modulated amplifier for television services, still the real problem lies in obtaining a satisfactory means for modulating the power amplifier.

The plate voltage of the 833's at carrier conditions will be half of the peak 1100 volts. To modulate this plate voltage 100% we must have a peak-to-peak modulating voltage of 1100 volts, and this must be generated across a capacity of approximately 200 mmfd. as calculated below:

$2\ C_{gp} = 2 \times 6.3$..........12.6 μμfd.
C_{pk}.......... 8.5
21.1
8 tubes.......... 8
168.8
Capacity of plate tank to ground........ 30.0
Total capacity........ 198.8 μμfd.

In modulator design, as in the design of video amplifiers, we set the load resistance equal to the capacitive reactance of the total capacity at the top modulating frequency. Doing this, we obtain for the modulator load a value of 188 ohms as follows:

$$X = \frac{1}{2\pi fc} = \frac{10^6}{2\pi \times 4.25 \times 200} = 188 \text{ ohms}$$

Since the current change is given by I = E ÷ R, we see that the tubes must develop a peak-to-peak current of 6 amperes across the load resistor in order to fully modulate the class C stage. This would require at least two water-cooled type 891 tubes operating with a plate voltage of 4700 volts at 1.6 amperes each, and dissipating a total of nearly 15 kilowatts. Even then, a special type of coupling network would have to be used in order that the output capacity of the 891's would not appear in parallel with the capacity of the load (explained in section 9), and therefore, further lower the load resistance and increase the necessary power. Under these conditions, the 891's would require a grid swing of 750 volts across a capacity of 174 mmfd., and another 891 would be required to drive the modulator.

A pair of 889 water-cooled tubes would be a more practical stage than the eight type 833's which we have considered. From formula (4), we see that it is possible to realize a 1 kw. carrier from a pair of these tubes.

$$(4) \qquad P = \frac{113\, I_b^2}{f_m C} \qquad \text{Then:} \quad P = \frac{113 \times 1}{4.25 \times 47.8} = .56 \text{ per tube.}$$

TYPE 889:		
Cgp = 17.8 × 2 =	....	35.6
Cpf = 2.2	....	2.2
Cgf = 18.8		
Strays	....	10.0
Total	....	47.8 mmfd.

Maximum rating Class C telephony: Ip = 1.0 amp.
" " " B " " " "
Ef = 11 volts. If = 125 amperes.
Dissipation = 5 kw. maximum.

However, it would not be practical to secure this output at a plate voltage much below 1200 volts. This requires that the modulator must have a peak-to-peak output of 2400 volts, or over twice the voltage necessary for the 833's. This greater swing is only partially offset by the reduced capacity into which the modulators must work. If we use type 891's for modulators, at least three, and preferably four tubes will be required. If four tubes are used, the capacity of the modulator tubes themselves determine the value of the modulator load instead of the capacity of the modulated amplifier:

MODULATOR

TYPE 891:	
Cgp....	27
Cpf....	2
	29
4 tubes....	4
	116
Strays....	10
Total Load Cap.....	126

MODULATED AMPLIFIER

TYPE 889:	
2 × Cgp = 2×17.8 = ...	35.6
Cpf	2.2
	37.8
Two Tubes	2
	75.6
Capacity tank to ground	30.0
Total load Capacity	105.6

This becomes about the limit of voltage which this type of modulator tube will deliver. If peak-to-peak voltages on the order of 4,000 to 5,000 volts are required, the ratings of the type 891 will be exceeded when the load resistor is set equal to the output capacity of the tube. Under these conditions, the ratings of the type 891 are exceeded even though the tube is supplying no power to the load, or is completely disconnected from the modulated amplifier.

It may be seen therefore, that for any particular modulator tube, there is a limit of voltage swing that can be developed, and that this voltage swing is proportional to the peak plate current that the tube will deliver, divided by its output capacity. The water-cooled type 899 would make an excellent modulator except that because of its high amplification factor, the grid is swung into the positive region at moderate plate voltages.

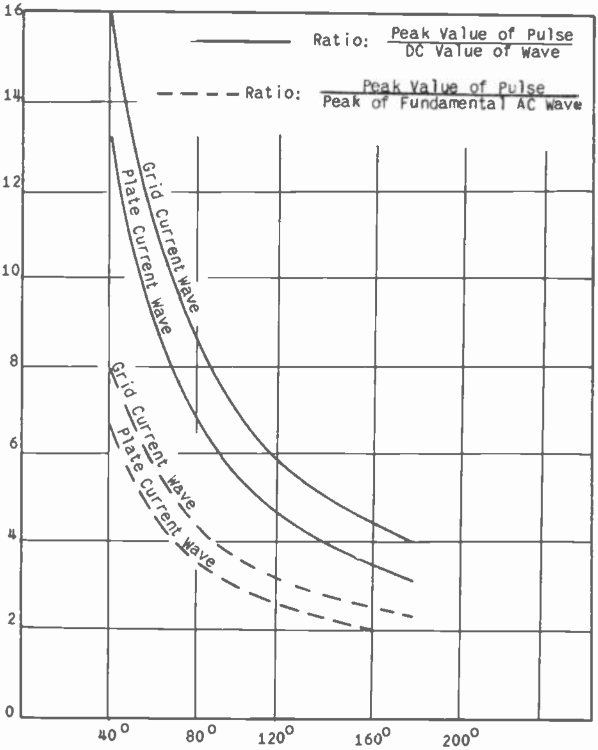

Fig.11 Effect of flow angle upon the plate and grid currents.

5. GRID MODULATION. Grid modulation requires a great deal less video swing for 100% modulation than does plate modulation, and therefore, the modulator tube requirements are considerably less. This improvement is achieved without any loss in peak power output

from a given modulated amplifier tube. Actually, slightly greater peak power outputs are obtainable from a tube using grid modulation than for the same tube when it is plate modulated. This equal or greater peak power output is obtained at lower efficiency, and consequently, greater plate dissipation, but since plate dissipation is not normally the limiting factor in the tube performance for television services, this is of little consequence. If the same tube complement in the modulated stage is used for grid modulation as for plate modulation and the same bandwidth is maintained, obviously, the RF load impedance with respect to the fundamental current will remain the same. For grid modulation, we can tolerate a somewhat greater flow angle of plate current than for linear plate modulation. This results in a greater ratio of fundamental current to peak current as obtained from Fig. 11, and consequently, the load line as illustrated in Fig. 9 can have a longer slope, with correspondingly greater voltage swing. This means that a tube operating with a longer flow angle will have greater power output for a given *peak* current, even though the dissipation will be considerably higher and its efficiency reduced.

At the present time, grid modulation is the most popular method employed for television picture transmitters. It is used not only by RCA and by the NBC and Columbia installations in New York City, but also is the accepted method used in England and on the continent. The RCA 1 kw. transmitter illustrated in Fig. 7 is block diagrammed in Fig. 12. The performance of this transmitter will be briefly considered.

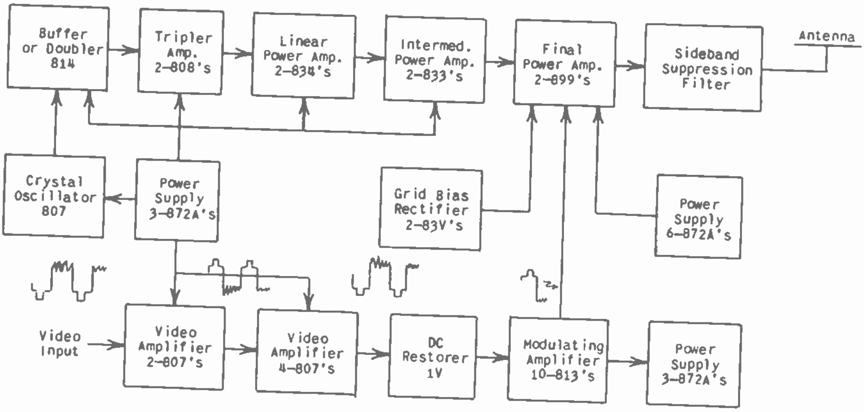

Fig.12 Block diagram of RCA 1 kw. picture transmitter.

At the peak of modulation, a grid modulated amplifier for television, operates in the same manner as a Class B linear amplifier. Under some conditions, the grid modulated stage may operate with a smaller flow angle, but generally it is arranged so that plate current will flow during 180° of the RF cycle exactly as in the case of a Class B linear. For grid modulation, or for linear amplifiers of the wide band television variety, the power output is limited by the peak usable emission of the tube, and by the minimum capacity obtainable across the output tank. This was true for the case of plate modulation also, but formula (4) must be changed when calculating the performance of grid modulated or linear amplifiers, due to the different flow angle and the consequent different form factor (Fig. 11) of the plate current pulse. When these differences are considered, the performance of the grid modulated tube, or of the linear stage, is predicted by the following expression:

$$P = \frac{98 (I_D)^2}{f_m C} \text{ Kw.} \qquad (5)$$

Where: P = carrier power in kw., I_D = maximum DC plate current rating, for linear amplifier service, and C = output tank capacity in mmfds. Applying this to the case of the 889's used in the 1 kw. transmitter, we obtain an output of just under one kw. for the pair of tubes using the following tentative data:

CLASS B TELEPHONY

DC plate current		1.0	Amp. Max.
Cgp = 17.8	2Cgp =	35.6	Mmfd.
Cpf = 2.2		2.2	
Cgf = 18.8		37.8	
Wiring Strays		10.0	
Total Capacity		47.8	

$$(5) \quad P = \frac{98 \times (1)^2}{4.25 \times 47.8} = .483 \text{ Kw. per tube} = .97 \text{ for two tubes.}$$

From this, it would appear that it is not possible to obtain a full kilowatt from this transmitter; however, a more detailed consideration of the problem will show that an output of one kilowatt is quite practical.

In the first place, the plate current drawn by the 889 stage is the average plate current over the modulation cycle. For television signals which are not symmetrically disposed about any particular axis, the average plate current over the modulation cycle will change with various types of picture signals. This should be apparent from Fig. 1. Due to the short duration of the synchronizing and blanking pulses, it happens that the axis of a typical television signal is displaced from the center of the wave toward the edge of the wave representing the highlights of the picture. For negative polarity of modulation, this means that the average current of the grid modulated stage will be less than for sine wave modulation. Since formula (5) is based upon symmetrical modulation, the value of current entered in this calculation can be higher than the actual current that the tube will draw, provided that the peak emission of the tube is not exceeded.

In the second place, the 899 is rated for the same current for Class C telephony as for Class B telephony, resulting in a greater peak current for the former case. If the value of current entered in formula (5) is altered to produce the same peak plate current as resulted in the calculation of the plate modulated stage, it is seen that the 889's will readily furnish the 1 kw. rating of the RCA transmitter of Fig. 12.

$$(5) \qquad P = \frac{98 \times (1 \times 1.14)^2}{4.25 \times 47.8} = .627 \text{ Kw. per tube.}$$

$$\text{Then:} \quad P = 1.25 \text{ Kw. per pair of 889's.}$$

A pair of 889's will deliver 1 kw. when operating with a plate supply voltage of 3000 volts. Under these conditions, the plate efficiency will be approximately 17% and the peak-to-peak video voltage required will be approximately 900 volts. During modulation peaks, the peak of the RF grid voltage will reach something over 700 volts positive. Considerable grid current will flow during modulation peaks, and some grid current will exist even at carrier conditions. Such conditions are never obtained in broadcast practice.

If the plate voltage is increased without increasing the peak plate current, no increase can be had in the output power. This follows from the same reasons discussed in preceding paragraphs for the case of plate modulation. Unlike plate modulation, however, as the plate voltage is increased, the modulator requirements are slightly decreased, rather than being made more severe. As the plate voltage is increased, the grid current becomes less, and a slight reduction is effected in the grid swing for a given output. It can be seen from Fig. 13, that the peak-to-peak modulating voltage is just equal to the peak RF grid swing. Consequently, an increase in plate voltage not only effects a slight decrease in RF driving voltage, but in modulating voltage as well. At about 5500 plate volts, the dissipation limit is reached for the type 889 with a plate efficiency of 9%. While an increase in plate voltage produces a severe reduction in plate circuit efficiency, the improvement in modulating requirements is only slight. Technically, this improvement is due to the divergence of the grid voltage lines on the plate characteristics away from the diode line, and can be seen to a slight extent in Fig. 9. The effect is much more apparent in large water-cooled tubes.

If the transmitter of Fig. 12 is operated at approximately 4500 volts, a video swing of 800 volts, peak-to-peak, will be required for 100% modulation. This will require ten type 813's operating at their maximum current rating. In this case, the tube load is determined by the modulator capacity, rather than by the capacity of the modulated amplifier as follows:

MODULATOR

TYPE 813:		
Cout	14	mmfd.
Ten tubes	10	
	140	mmfd.
Wiring	10	
Total Capacity	150	mmfd.

MODULATED AMPLIFIER

TYPE 889:		
	Cgp x 2	35.6
	Cgf	18.8
		54.4
	Two tubes	2
		108.8
	Grid tank to ground	25.0
	Total Capacity	133.8

$$R_L = X_C = \frac{1}{2\pi f c} = \frac{10^6}{2\pi \times 4.25 \times 150} \qquad R_L = 250 \text{ ohms.}$$

To secure 800 volts across this load resistor will require a current change of 3.2 amperes or 320 ma. per tube. If a waveform distortion of approximately 12% is tolerated, the modulators could be operated with an average current of 160 ma. per tube.

Applying the same procedure to the transmitter of Fig. 6, a power output of approximately 4 kw. is obtained, which is readily modulated by a pair of 891's.

899's GRID MODULATED:

Max Ip for Class C Telephony.......2 Amp.
" " " " " " ....2 Amp.
Cgp.........23 mmfd.............2 Cgp = 46
Cpf......... 5 " 5
Cgf.........10 51
Wiring Strays..... 10
Total Capacity..... 61 mmfd.

$$P_{out} = \frac{98 \times (2 \times 1.14)^2}{4.25 \times 61} = 1.97 \text{ Kw. each.} \tag{5}$$

$$P_{out} = 4 \text{ Kw. for two tubes.}$$

The plate voltage can be between 5000 volts and 11,000 volts without exceeding the tube rating. In Fig. 6, it is seen that the CBS transmitter is operated with 8900 volts. The peak-to-peak video voltage, as well as the peak RF voltage required, will be a trifle less than 1500 volts which can be achieved with a pair of 891's as seen below.

MODULATOR

TYPE 891'S:
Cgp.........27
Cpf......... 2
29
Two tubes.... 2
58
Wiring.......10
Total........68 mmfd.

MODULATED AMPLIFIER

TYPE 899'S:
2 x Cgp............. 46
Cgf............. 10
56
Two tubes...... 2
112
Grid tank to ground... 25
Total..........137 mmfd.

$$R_L = X_c = \frac{1}{2\pi fc} = \frac{10^6}{2\pi \times 4.25 \times 137} = 275 \text{ ohms.}$$

$$\text{Peak-to-peak current} = \frac{1450 \text{ volts}}{275 \text{ ohms}} = 5.26 \text{ Amp., or } 2.6 \text{ Amp. per tube.}$$

Modulator average current will be approximately 1.5 amperes.

Grid modulation for television services operates in a somewhat different manner than ordinary grid bias modulation in broadcast transmitters. A rough comparison of this is shown in Fig. 13. This type of graph is *not strictly accurate for RF purposes*, since, as mentioned before, the tank circuit is not a pure resistance, and, therefore, the slope of the dynamic E_g–I_p characteristic will change as the fundamental component of the plate current pulses change. There are a number of comparison points between the two types of grid modulation that this graph will show clearly: first, because of the lower impedance load, the plate current pulses are considerably larger for television transmitters; secondly, the tube is operated into the positive grid region and is driven to the peak current that the tube will stand. Furthermore, due to the decreased plate voltage used, the cutoff value of plate current occurs at a lower bias in television transmitters.

The high values of grid currents required, demand a video modulator of low internal impedance if good fidelity is to be preserved. Because of the bandwidth involved, the modulators are either direct-coupled or resistance-coupled into the grid circuit of the modulated amplifier, and consequently, their internal impedance is simply the plate resistance of the modulator tubes. This results in curvature on the positive modulation peaks, which will tend to reduce the amplitude of the synchronizing pulses.

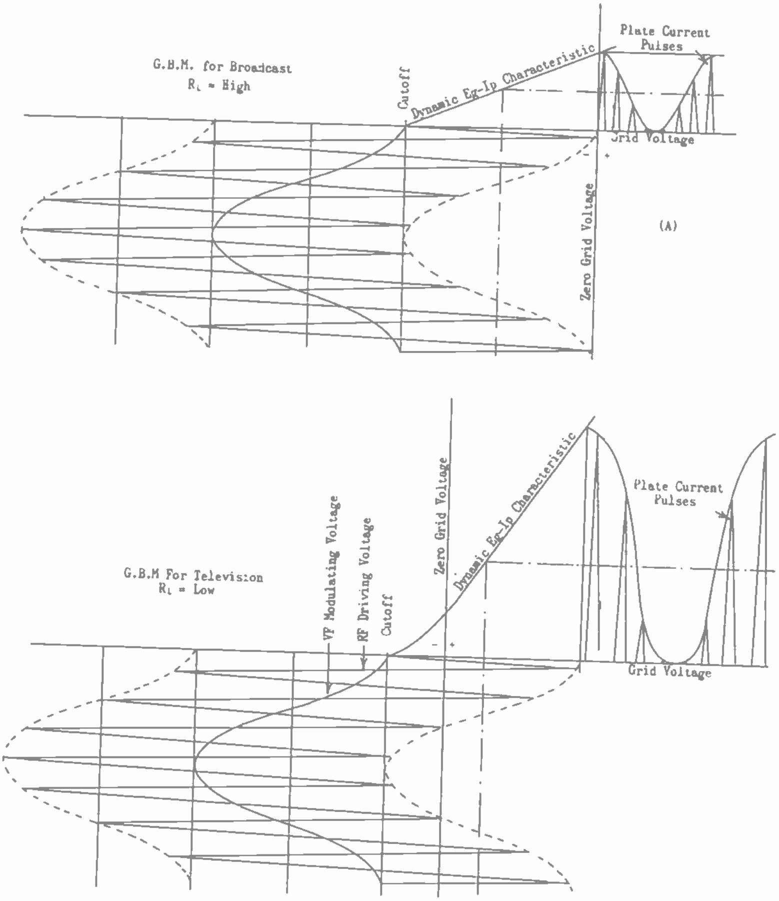

Fig.13 A general comparison between grid bias modulated transmitters for standard broadcast application and for television picture service.

Due to the fixed value of excitation required for maximum power output, and due to the fixed load impedance required for maximum bandwidth, it is not possible to adjust the grid modulated transmitter for television services to secure a perfect amplitude characteristic with respect to the modulating voltages. Curvature occurs during the negative modulation peaks and shows up in the transmitted picture as a slight reduction in contrast in the highlights or the white portions of the picture. This effect is apparent in Fig. 13.

6. TRANSMISSION LINE MODULATION. A bread-board setup of a transmission line modulated transmitter operating at a carrier frequency of 200 mc. is shown in Fig. 14. The schematic diagram of this transmitter is shown in Fig. 15.

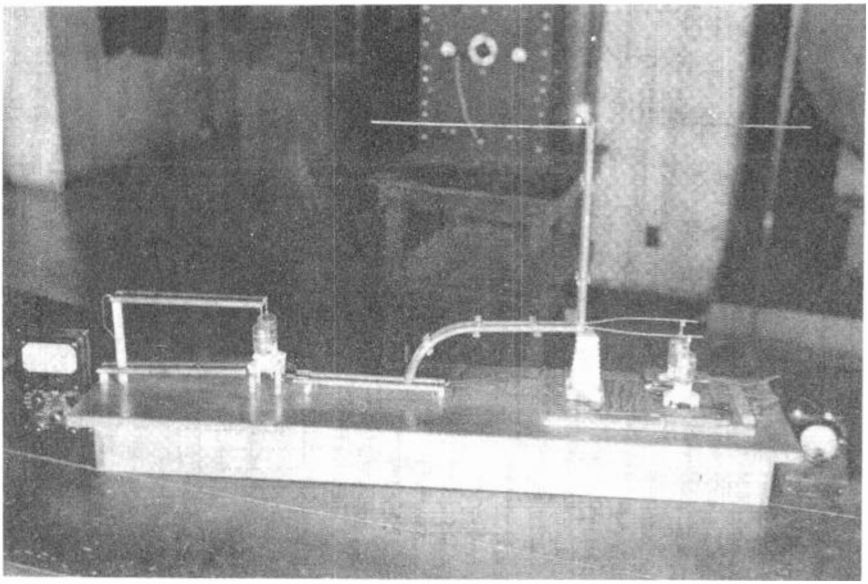

Fig.14 A laboratory 200 mc. picture transmitter employing transmission line modulation.

Before considering the mode of operation of a transmission line modulated transmitter, let us review from the preceding lesson, one of the characteristics of a quarter-wave section of transmission line. You will recall that a quarter-wave section of line operated at its resonant frequency has the property of an impedance inverter; that is, the characteristic impedance of the section of line is the mean between the sending end and the receiving end impedance according to the following relation:

$$\frac{Z_{in}}{Z_o} = \frac{Z_o}{Z_{out}} \quad \text{or:} \quad Z_{in} = \frac{Z_o^2}{Z_{out}} \qquad (6)$$

Where Z_o is the characteristic impedance of a quarter-wave line section.

In the transmitter of Fig. 15 are shown three such impedance inverters; L_t, L_m, L_a, all joined together at the junction J. In addition to these three transmission lines, three others are used to form the tuned circuits in the grid, plate, and cathode of the push-pull oscillator. When operating at lower frequencies, only a grid and plate tank would normally be used, but at 200 megacycles or higher, the additional tank in the cathode circuit serves to obtain a better control over the phase and amplitude relations of the

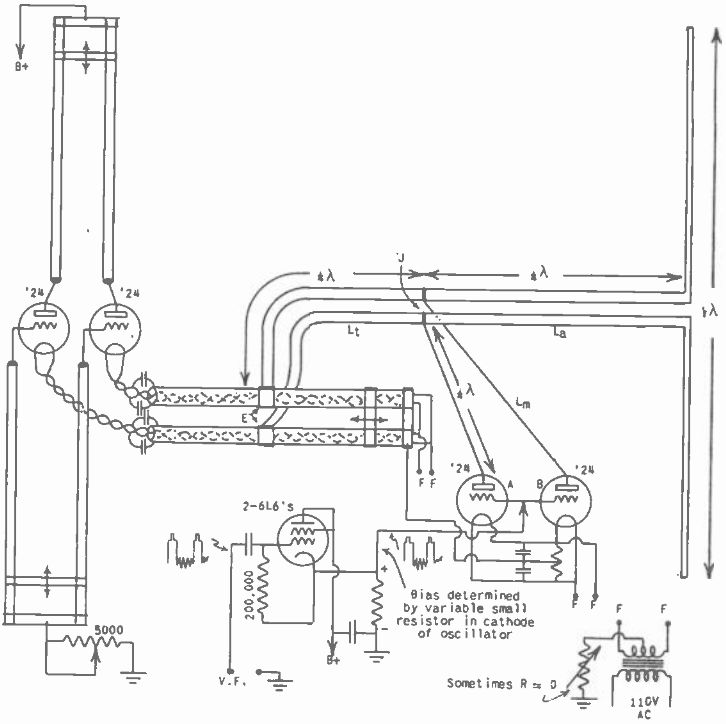

Fig.15 Circuit employed in laboratory 200 mc. T.L.M. picture transmitter.

exciting voltage and the output voltage. It would be quite practical to obtain the output from either the grid tank or the plate tank with preference given to the cathode tank which was used in the case of this particular transmitter.

Let us now consider the operation of the transmitter as a whole. If we were to place a short circuit across point J, no power would be transmitted to the antenna load, nor would any power be absorbed from the RF oscillator, because this short circuit at point J would be reflected as an open circuit at the oscillator tank, due to the quarter wave transformer that joins these two points. If point J were open circuited, the antenna would be connected to the RF oscillator via the transmission line, and the coupling could be adjusted until the maximum output obtainable from the oscillator was delivered to the antenna. This would correspond to peak conditions of modulation. Now, if the characteristic impedance of line La was so adjusted that the antenna load was reflected to point J as 40 ohms, then if we were to place an additional 40 ohms in parallel with the antenna transmission line at point J, half the power appearing at point J would be dissipated in this additional resistor, and half would go into the antenna.

The power appearing at J, under this latter condition, would be only half the power which appeared before the additional 40-ohm resistor was added. The reason for this is as follows. Suppose the transmission line L_t had a characteristic impedance of 40 ohms. (This is for the sake of simplicity in explanation. In practice, the actual value of the line L_t is unimportant). Then, under peak conditions, the antenna load of 40 ohms at point J would be reflected to the oscillator as 40 ohms, and a certain power, $E^2 \div 40$, would be delivered to the antenna. When the additional 40 ohms is placed at point J, the total impedance at J becomes 20 ohms which is reflected to the oscillator tank as 80 ohms, due to the impedance inverting characteristic of the line L_t. Since the voltage generated at the oscillator tank remains unchanged, the total power delivered to point J is now $E^2 \div 80$, or only half of the total power delivered before this additional resistor was added. Of this reduced power, only half of this actually reaches the antenna, so the antenna power is now only one-quarter of the peak output, corresponding of course, to carrier conditions. Therefore, at carrier conditions, it is necessary to dissipate in the control resistor at J, a power equal to the power delivered to the antenna.

Let us consider the function of the line L_m. The line L_m is adjusted, in conjunction with the tube capacities of the modulators, to be electrically one-quarter wave long. Under these conditions, L_m becomes an impedance inverter. Suppose it has a characteristic impedance of 300 ohms. If we place a resistance of 2250 ohms across the plates of the modulator tubes, this will be reflected to point J as 40 ohms, corresponding to carrier conditions.

$$(6) \qquad \frac{Z_{in}}{Z_o} = \frac{Z_o}{Z_{out}} \qquad \text{Or:} \qquad \frac{2250}{300} = \frac{300}{40}$$

If the tubes were open circuited, point J would be shorted out, due to the inverting action of L_m. This would correspond to zero output (zero antenna power), or the valley of modulation. If a very low resistance were placed across the modulator tubes, say for instance, 100 ohms, this would be reflected as 900 ohms, across J. Since J is already terminated in 40 ohms, which is reflected by the antenna transmission line, the additional 900 ohms added by the modulator line is of little consequence, and practically full output is delivered to the antenna.

It can be seen that if we could obtain a variable resistance across the end of the line L_m, we could modulate the power delivered to the antenna load. This is accomplished by making use of the variable plate resistance of the modulator tubes M. No DC plate voltage is applied to these tubes, although an RF voltage of course exists. A small RF voltage also occurs between the grids and filaments of the modulator tubes due to the voltage drop across the lead AB in Fig. 15 which interconnects the modulator grids. Since this lead is carrying a heavy RF current, due to the grid-plate capacity of the modulator tubes, a push-pull voltage is developed between opposite grids. In addition to this small push-pull RF voltage, a video voltage is applied between the parallelled modulator cathodes and the parallelled modulator grids. The video voltage controls the plate resistance of the modulator tubes, and in turn, controls the power delivered by the RF oscillator to the antenna load. When the modulators are driven to cutoff, they are open circuited and dissipating no power; nor is any power delivered to the antenna. When the grids of the modulators are driven highly positive, the voltage drops across the tubes are quite low, and again, very little power is dissipated in the modulator. Under this condition, maximum power is delivered to the antenna load.

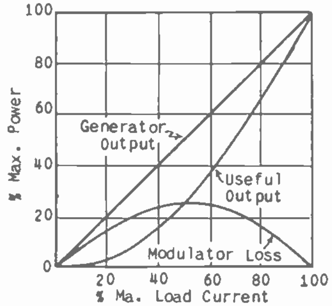

Fig.16 Relation between dissipated power and useful power in a T.L.M. transmitter.

For other values of modulator bias between zero and peak output, the modulator assumes various plate dissipations. At a bias conforming to carrier conditions, the modulator loss is exactly equal to the power delivered to the antenna. The dissipation curve over the modulation cycle is a portion of a circle, as shown in Fig. 16.

For carrier conditions, the RF oscillator, or power amplifier as the case may be, must be capable of supplying 2 kw. output for 1 kw. in the antenna. It must also have a peak capability of 4 kw. Under optimum conditions, this could probably be achieved with a dissipation of between 1 and 2 kw. The modulator must be capable of dissipating a power equivalent to the antenna power at carrier conditions, and therefore need have a dissipation of only 1 kw. In practice, however, it is desirable to have sufficient power capabilities in the oscillator or power amplifier to maintain, for short intervals of time, the peak output of the transmitter as this greatly facilitates adjustment. Consequently, the Philco 1 kw. transmission line modulated television transmitter uses a pair of 846's as a water-cooled oscillator in the RF generator. These tubes have a combined dissipation of 5 kw.

In order to drive the modulators to low values of plate resistance to obtain high peak output from the transmitter, the grids are driven highly positive during the peaks of modulation. Terrific grid currents result, and it is necessary to choose a tube designed to carry large grid currents and withstand considerable grid dissipation. The tube must also have a steep diode line, which of course means a low minimum plate resistance. For these two reasons, it is not generally possible to use tubes having a total dissipation equal to the power output of the transmitter. In the case of the Philco 1 kw. transmitter, four type 450T's are used as modulators.

This type of modulation is theoretically quite linear with respect to the video voltage impressed upon the grids of the modulators. Due to the large grid currents taken by the modulators, however, considerable non-linearity will result in the driver ahead of the modulators unless it has a low internal impedance. A cathode loaded tube (Fig. 17) would serve excellently for this purpose, since it degenerates at all frequencies, and the feedback in the cathode circuit removes the non-linearity which would otherwise be introduced. A cathode loaded tube in contrast with a plate loaded tube requires a great deal more grid swing. Where one normally expects to obtain a voltage gain in a video stage ahead of the modulator, a cathode loaded driver would yield a voltage loss, and the stage ahead of the video driver would then have to develop a great deal more voltage, and consequently, be a very much larger tube than in the case of a plate loaded driver. It would not necessarily have to furnish any grid current to the cathode loaded stage, and the cathode loaded stage would maintain the same picture polarity as shown in Fig. 17.

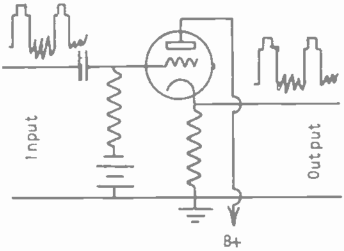

Fig.17 Cathode loaded video amplifier having low driving impedance.

Since this ingenious modulation means removes the limitation of bandwidth from the RF stage, it can operate with much higher efficiency. While the impedance inverter line Lm does have some discrimination against sideband frequencies, still, the overall result is a transmitter of greater bandwidth for a given number of tubes employed. Furthermore, the possibility of linear modulation under optimum conditions is an improvement over the inherent non-linearity of grid bias modulation. The video voltage swing required for the transmission line modulator is less than the modulation output voltage necessary for grid bias modulation. The practical difficulties are the unusual modulator grid requirements, particularly with regard to the heavy grid currents, the difficulty of maintaining linearity in the driver and the difficulty of maintaining adjustment of the various transmission line functions.

In order to calculate the performance of a particular tube used as a modulator in a transmission line modulated system, it is necessary to re-plot the plate characteristics in terms of RF plate voltage versus RF plate current in order to draw a load line on the characteristics, since no DC voltage is applied to the plates of the tubes. The power output then derived from the load line would be the sideband power generated by the modulator. This method is rather tedious since each point on this new family of curves must be worked out from the static characteristic. Rather than analyze the theoretical performance of a 1 kw. picture transmitter, using this method of modulation, it is better for our purpose to inspect a typical case and study its performance. The 1 kw. Philco transmitter previously mentioned was very excellently designed, engineered, and described by the inventor of transmission line modulation—W. N. Parker.[1] The overall modulation characteristic of this transmitter is shown in Fig. 18. Fig. 18A shows the effect of varying the impedance at the junction J, while Fig. 18B shows the effect of varying the inductance of the connecting strap between the grids of the push-pull modulators, thereby allowing small values of RF voltage

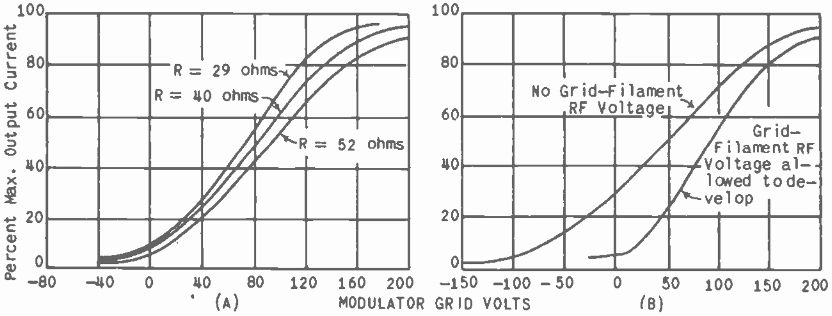

Fig. 18 Typical modulation characteristics of a transmitter employing transmission line modulation.

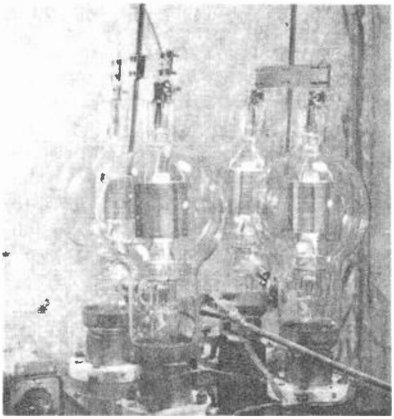

Fig. 19 Four type 450T's as a shielded modulator in a T.L.M. television transmitter.

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[1] A Unique Method of Modulation for High-Frequency Television Transmitters, William N. Parker, Aug. '38, I.R.E.

to appear between them. It will be noted that increasing the inductance of this connecting strap results in a smaller amount of required swing from the driver, but a swing which appears almost entirely in the positive grid region (as seen in Fig. 18), resulting, as previously mentioned, in very heavy grid current. The curvature at the upper part of the performance curve is due to overload from the RF source; in this case, an oscillator, and could be reduced by the use of the same RF generator tubes as a power amplifier, in order to obtain greater output and consequent lower RF driving impedance. The curvature at the lower part of the modulation characteristic is largely due to losses in the modulator line, which must consist of small conductors widely spaced in order to obtain the correct characteristic impedance. Radiation resistance from this line is materially reduced by enclosing the line, modulator tubes and all, in a metal box as shown in Fig. 19.

7. LINEAR AMPLIFIERS. While it is true that successive similar stages employed in a chain of linear amplifiers greatly reduce the bandwidth of the transmitter compared to a high-level system of modulation where only one tuned circuit is involved, nevertheless, satisfactory results can be obtained, due principally to two factors. First, the standardization of single sideband transmission has reduced, by a factor of approximately 1.55 to 1, the range of frequencies which must be handled in the RF amplifier, as will be seen by referring to Fig. 22. For a top modulation frequency of 4.5 mc., the transmitter would have to be 9 mc. wide in order to handle the two sidebands which would be generated in the case of double sideband transmission. With the advent of single sidebands, this bandwidth for an equivalent video frequency has been reduced to 5.75 mc. Actually, the bandwidth may be somewhat smaller than this, since the video frequencies must be completely attenuated at the extreme edges of the band just mentioned. The bandpass coupling circuits in the linear amplifier stages can then be adjusted in order to assist in the attenuation of the outermost sideband frequencies. This serves the double purpose of assisting the filter action of the single sideband filter and, at the same time, increases the impedance into which the RF tubes must work because of the increased bandwidth.

The second factor which is responsible for the satisfactory performance of a chain of linear amplifiers is the ability to combine successive stages in such a way that the attenuation of certain frequencies in one stage is offset by a peak in the response at the same frequencies in the following stage. When this can be achieved with satisfactory overall phase characteristic, the transmitter will behave as if each tuned circuit were essentially flat over the entire band. This can be best understood by considering Fig. 20. It was necessary to choose a logarithmic scale on which to plot the gain of the successive stages, since the overall transmission of the band of frequencies is the *product* of the individual stage gains. Since multiplying the gains of the individual stages gives the overall gain of a transmitter, we can achieve the same result graphically by plotting the gains on log paper and adding the total amplitude.

The solid line of Fig. 20 represents the gain of the linear amplifier stage of Fig. 21, while the dashed curve of Fig. 20 represents the gain, or the response, of the grid modulated stage which drives the linear amplifier. The upper curve, which is the sum of the other two, is the overall performance of the transmitter. A satisfactory bandpass, such as the solid curve of Fig. 20, can be achieved in the final amplifier, since it works into a constant load—the antenna. The antenna reflects the damping on the overcoupled circuit necessary to give the required low effective Q. The bandwidth of the grid modulated stage is adjusted by changing the load resistance R of Fig. 21, into which the stage must work.

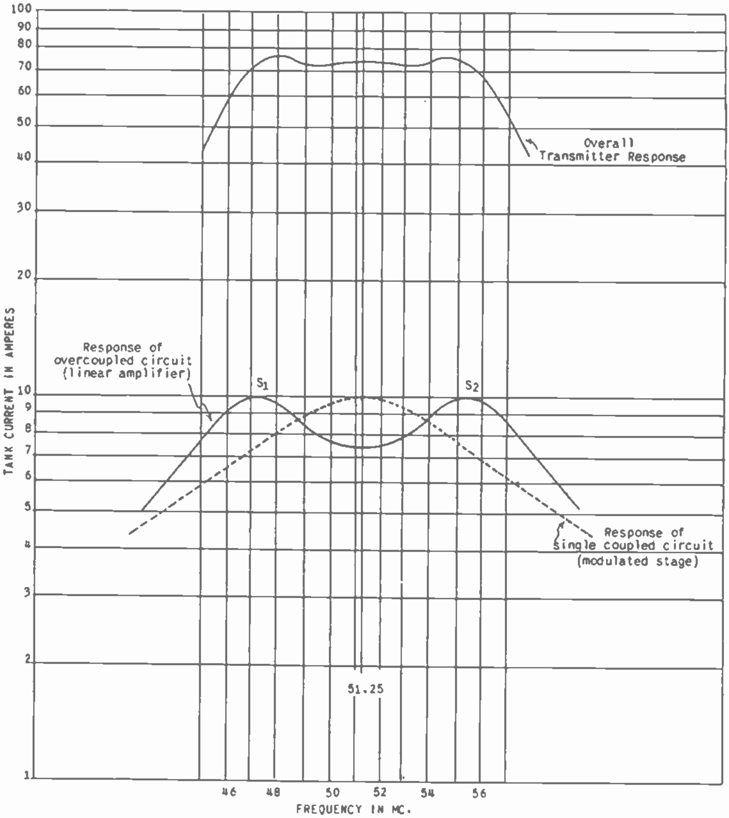


Fig. 20 Response of individual stages of a transmitter employing linear amplifiers.

The grid current drawn by the linear amplifier at the peaks of modulation increases the damping on the tuned circuit during maximum modulation peaks. This affects the shape of the dashed curve in Fig. 20 in a manner depending upon the instantaneous value of modulation. This effect becomes more serious when the tuned circuits of Fig. 21 are reversed; that is, when the overcoupled circuit is placed in the plate of the grid modulated stage and the single-tuned circuit becomes the tank of the linear amplifier. Under these conditions, the bandwidth of the overcoupled circuit and the distance between the peaks S_1 and S_2 (Fig. 20) changes with the varying amplitude of modulation peaks. Fortunately, this effect occurs chiefly during the blanking out and synchronizing impulses since they produce maximum transmitter output. These pulses are composed essentially of the lower frequencies and are not affected by the varying bandwidth of the transmitter. Likewise, amplitude distortion introduced during this interval is not so important as it would be if it occurred in the picture signal below the blanking axis.

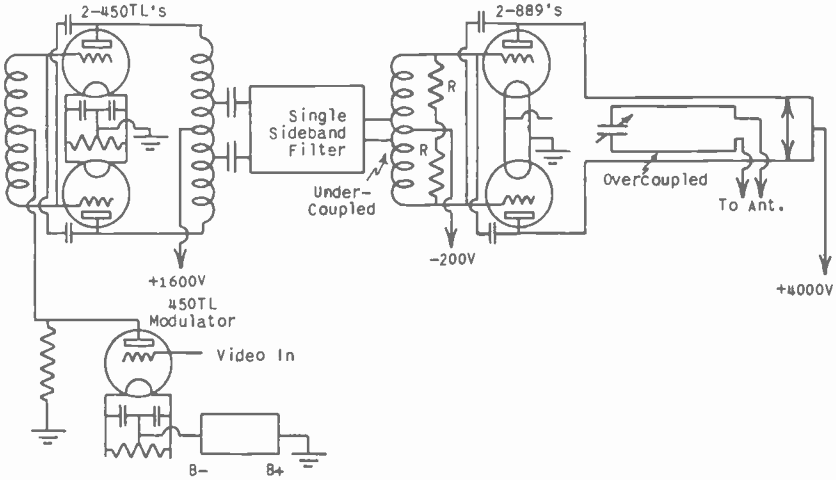


Fig.21 Schematic of one form of television transmitter employing low level modulation.

If several stages of linear amplification are incorporated, it is desirable to maintain an equal number of double-tuned and single-tuned circuits. The student should realize that while only simple overcoupled tuned circuits have been mentioned, numerous other types of bandpass circuits would be applicable for interstage coupling. Each stage must be able to furnish the peak driving power required by the following stage, and this peak power must be furnished with good RF regulation. It must furnish this amount of power over a wide band of frequencies, and therefore, only tubes which will produce considerable power gain with low-impedance load circuits are satisfactory.

We have calculated the performance of a pair of type 889's and 899's as grid modulated amplifiers. Using these calculations as a basis, it is easy to determine the performance of the same tubes as linear amplifiers.

At the peak of modulation, the Class B linear and the grid modulated amplifier perform in exactly the same manner. If we use tank circuits of identical Q in the transmitter shown in Fig. 21 as were used in the calculations for the grid modulated transmitter, the former calculations will apply. Thus, the 889 stage of Fig. 21 will develop 1 kw. output with a peak RF driving voltage of 800 to 900 volts, depending upon the plate voltage employed. The grid tank must be heavily loaded to secure the required bandwidth, and under these conditions will present a load impedance of 580 ohms as follows:

TYPE 889:	2 × Cgp	35.6 mmfd.
	Cgf	18.8
	Wiring Strays	10.0
	Total	64.4 mmfd. per tube.

Since the tubes are in push-pull, the total tank capacity becomes:

$$\frac{64.4}{2} = 32.2 \text{ mmfd.}$$

$$X_C = \frac{1}{2\pi fc} = \frac{10^6}{2\pi \times 51.25 \times 32.2} = 96.5 \text{ ohms.}$$

$$R_L = Q \times X_C = 6 \times 96.5 = 580 \text{ ohms.}$$

Choosing the highest probable grid swing, but neglecting the small driving power actually consumed in the amplifier tube, the power dissipated in the grid tank, which must be supplied by the driving stage, is roughly 180 watts for an output carrier of 1 kw. This represents a power gain of over five in the linear stage, which is quite good.

$$W = \frac{E^2}{R} \qquad E = \text{RMS grid swing} = \frac{\text{Peak grid volts}}{1.414} = \frac{900}{1.414}$$

$$W_P = \frac{(900 \times .707)^2}{580} = 718 \text{ watts at peak of modulation.}$$

$$W_C = \frac{718}{4} = 180 \text{ watts driving power at carrier conditions.}$$

Similar calculations for the 899 indicate a power gain of eight, which is most remarkable, considering that this figure is for double sideband transmission, with a top modulation frequency of 4.25 mc.

If the linear stage is to follow a single sideband filter, the bandwidth would be decreased by a factor of 1.55, as previously mentioned. The power output and the impedance of the grid tank could be increased in the same proportion, giving a carrier of over one and one-half kilowatts, with a driving power of 170 watts, or a power gain of nine to one. At a sacrifice of plate circuit efficiency, a power gain of ten to one or better could have been achieved. The grid driving voltage is increased slightly over the double sideband case, but the increase in impedance of the grid tank more than offsets the change in driving voltage.

$$\text{Driving power} = \frac{E^2}{R} = \frac{(950 \times .707)^2}{564 \times 1.55} = 515 \text{ watts.}$$

Thus, it requires 515 watts at peak of modulation to supply tank loss. The grid power necessary to supply the tube loss = 170 peak, and the total driving power = 685 peak or 170 watt carrier.

If the station engineer is not daunted by the complexity of circuit adjustment in a series of wide band linear amplifiers, the system offers definite advantages over high - level modulation when single sideband is employed.

8. SINGLE SIDEBAND TRANSMISSION. Since television receivers are adjusted to accept only the upper sideband of the radio transmitter and a very small portion of the lower sideband, the use of double sideband transmission becomes a waste of valuable space in the RF spectrum. Therefore, television picture transmitters are adjusted to eliminate most of the lower sideband in the manner shown in Fig. 22. This is R.M.A. standard T-115. This standard sets a limitation not only upon transmitter performance, but upon receiver performance as well. In order to understand the design of a transmitter filter, it is necessary to first consider the receiver performance.

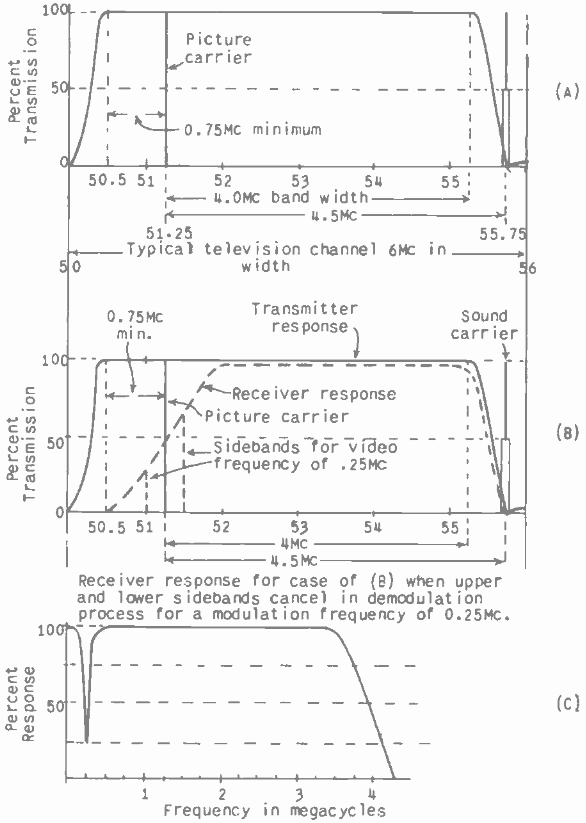


Fig.22 The characteristics of a television system designed for single sideband transmission and reception.

If we had an ideal bandpass at the receiver, such as shown in Fig. 24A, then we could locate the carrier at f_1 and we would accept only one sideband. Under these conditions, the transmitter could transmit either double sideband pictures or single sideband pictures, or could cut off any place in between on the lower sideband without affecting the receiver in any way whatsoever. Under these conditions, however, the receiver reduces the effective modulation percentage of the transmitter by 50%. Furthermore, the result is a combination of amplitude and phase modulation. This can be shown by referring to Fig. 23.

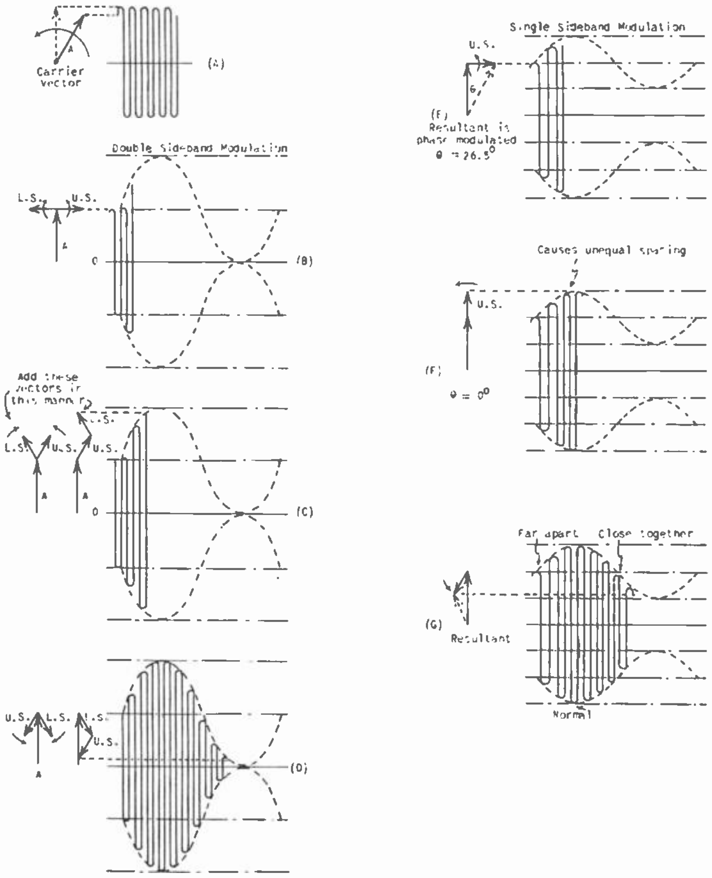


Fig.23 The effect of single sideband suppression upon the characteristics of modulated waves.

In Fig. 23 we have represented the carrier as a vector which rotates with constant velocity, and having constant magnitude. The projection of this rotating vector is the familiar sine wave of carrier voltage. When modulation occurs, there are, in addition to the carrier vectors, two other smaller vectors representing the two sidebands, one rotating faster and one slower than the carrier vector. In order to visualize the modulation process, we can let the carrier vector remain still, as in Fig. 23B, in which case the sideband vectors will rotate with a direction and velocity equal to the frequency difference between the carrier and the sideband frequency. These sideband vectors will add to, and subtract from, the carrier in such a way as to produce amplitude modulation, as shown in Figs. 23B, C, and D. The result is pure amplitude modulation. If one of these sideband vectors is removed, we can see from Figs. 23E, F, and G, that the result is a modulated carrier which is not only modulated in amplitude, but in phase as well. Furthermore, the modulation percentage is reduced. It has been shown that although this type of phase modulation introduces some distortion, for all practical purposes, it has a negligible effect upon the picture quality. For the rest of the discussion then, we can eliminate the effect of this particular type of phase distortion and continue with the consideration of the effect of a single sideband upon the amplitude of the wave.

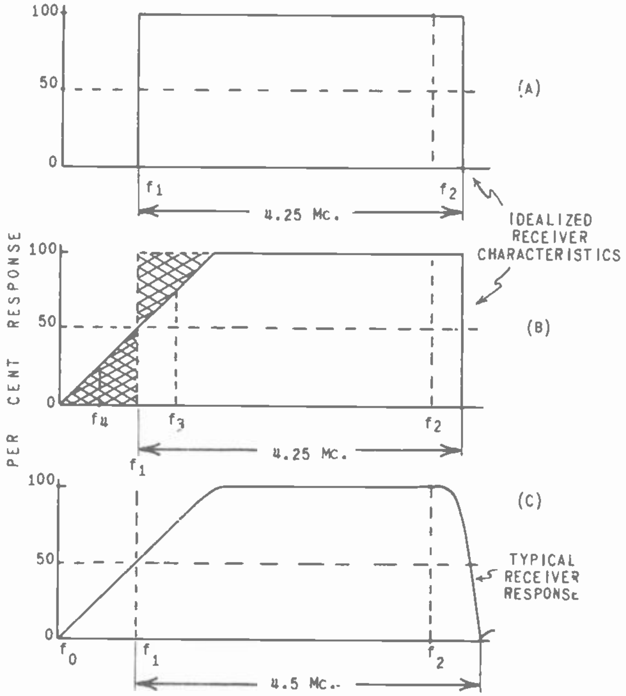


Fig. 24 Frequency characteristic of a single sideband receiver.

From the foregoing discussion, it appears that even if the characteristic of Fig. 24A were obtainable, which of course is not the case, it would not necessarily be the most desirable response characteristic of the receiver.

Consider a receiver having an amplitude response such as Fig. 24B. If the carrier is placed at f_1, then for very low audio frequencies, both sidebands are received and modulate the carrier in the usual manner of double sideband reception. At very high frequencies, such as f_2, only one sideband is received, but its amplitude is twice as great as the amplitude of the low audio frequencies and, consequently this sideband frequency modulates the carrier to exactly the same percentage as in the double sideband low frequency case. For intermediate frequencies (f_3) which are not quite sufficient in amplitude to completely modulate the carrier, a vestige of the lower sideband, such as f_4, is transmitted to make up the difference. In effect, the residual of the low frequency sideband adds to the high frequency sideband to make up the deficiency as has been shown by the shaded areas in Fig. 24B. As far as amplitude response is concerned, the receiver behaves exactly as in the double sideband case.

This condition is approximated in practice by operating the carrier down on the slope of the response characteristic of the receiver, as pictured in Fig. 24C. At frequency f_o of Fig. 24C, we will consider that the response of the receiver is sufficiently reduced that any frequency lower than f_o would not appreciably assist the upper sideband in the demodulation process. In other words, frequency f_o should be 10 to 100 times lower in amplitude than f_1 or carrier. With this in mind, we can now refer to Fig. 22 and consider the limitations imposed upon the transmitter.

In Fig. 22B, we have repeated the transmitter signal sideband standardization, but have included in the same curve a typical receiver response characteristic when the receiver is tuned to a carrier of 51.25 mc. The requirement placed upon the transmitter is that it must have a good frequency response as well as a good phase characteristic over the range from 50.5 to 55.25 mc. Since this is the only range of signals accepted by the typical receiver, any distortion of the transmitted signal between 50 mc. and 50.5 mc. should not be interpreted by the receiver, but any distortion occurring between 50.5 and 51.25 mc. would cause distortion in the received picture. For instance, if at a frequency of 51 mc., the sideband was delayed sufficiently to cause a 180° phase shift, the lower sideband would subtract from, instead of adding to, the upper sideband at 51.50 mc., and the resulting video characteristic would be distorted as shown in Fig. 22C. A phase shift at 50.25 mc. would not affect the receiver in this manner, since it is outside of the reception band of the receiver. Although the transmitter must preserve full output with no phase distortion from 55.25 mc. to 50.5 mc., it must cut off very rapidly between 50.5 and 50 mc., regardless of the phase conditions in this region. The output must be down to a very low value at 50 mc. in order that it will not produce sideband energy in the channel occupied by another station which, of course, extends to 50 mc. A small amount of output could be tolerated at 50 mc., but at 49.75 mc., the output should be very much less than

1% (60 db below carrier being a good figure), in order that it will not interfere with the sound carrier of the adjacent channel.

The single sideband filter can be inserted between the modulated amplifiers and a chain of linear amplifiers in the radio transmitter, or the filtering can be done at high level and the filter inserted between the output of the transmitter and the antenna terminals.

The performance characteristics of single sideband filters have been set forth. The physical arrangement for accomplishing the desired results will be considered in the next lesson on television antennas, since filters are composed chiefly of transmission line sections and bear more relation to the general problem of antennas and RF transmission, than to the problem of transmitters and RF generation.

9. MODULATOR AND VIDEO AMPLIFIERS. From the preceding sections on transmitter design for the various modulating systems, the power requirements and tube complements of the modulator were covered. In this section, we will consider the video amplifier which must feed these modulators, and we will discuss the circuit requirements which are necessary to effect a direct coupling between the modulator and the modulated amplifier to maintain the proper bias on this, as well as the lower power stages.

In studying video power amplifier requirements, at least three points must be considered. First, we must have the correct number of stages to maintain the proper polarity of signal at the modulator grid; secondly, we must have sufficient gain in the stages to supply the modulator with a 100% modulation capability; lastly, the number of tubes that must be operated in parallel for each stage must be determined. Another consideration is the coupling arrangement between the successive video stages in order to obtain an effective DC coupling, especially when grid current is required.

In general, transmitters are designed to accommodate a video signal of such polarity that the synchronizing pulses and the black of the picture are in a negative direction at the video input terminal of the transmitter. In the case of grid modulation, this means that an even number of video stages must be employed in order that the polarity of modulation will cause the synchronizing pulses to be represented by an increase in carrier, according to the R.M.A. standards. The polarity of waveform can be followed through in the block diagram of Fig. 12, which shows the 1 kw. RCA transmitter. In the case of transmission line modulation, an odd number of video stages must be employed in order to make the synchronizing pulses highly positive at the grid of the modulator, which of course, corresponds to maximum output. An even number of video stages must be employed in the case of plate modulation for the same reason an in the case of grid modulation.

The second and third points mentioned under this section heading, namely the overall gain and number of tubes per stage, are best explained by considering a specific example, for which we have chosen the transmitter outlined in Fig. 12. This is the same transmitter that was used in illustrating the performance of grid bias modulation. In previous calculations we found that the 813 modulators must deliver a peak-to-peak video voltage of 800 volts across a load resistor of 250 ohms. From the tube characteristics we find that with this value of load resistor, the modulators will require a peak-to-peak swing of 75 volts. If the transmitter is to be supplied from a line which will furnish a peak-to-peak output of two volts; then the amplifier ahead of the modulator must have an overall gain of 37.5.

We have learned that an even number of video stages must be employed. If we are to reduce the number of stages to two, in preference to four, then the gain per stage must average better than six. The last stage must work into a capacity of approximately 173 mmfd., and so cannot have a load resistance of over 216 ohms, even if the capacity of the video amplifier itself can be effectively eliminated. Under these circumstances, at least four type 807's would be required to furnish a gain of five.

10 - 813's MODULATORS:
Cin.........16.3 mmfd.
10 tubes... 10
163
Strays..... 10
Total...... 173 mmfd.

$$R_L = X_C = \frac{1}{2\pi fc} = \frac{10^6}{2\pi \times 4.25 \times 173} = 216$$

$$\text{Gain} = G_m \times R_L \qquad (7)$$

807 DRIVERS:
Gm........6000 mmhos.
Cout..... 7 mmfd.
Cin....... 11 mmfd.

Gain per 807 (neglecting their output capacity) = $6000 \times 10^{-6} \times 216 = 1.3$.
Gain for 4 tubes = $4 \times 1.3 = 5.2$

The first stage must have a gain of 7.2 and operate into a load resistor of not greater than 680 ohms as follows:

$$\text{Required gain of first stage} = \frac{\text{overall gain}}{\text{gain of last stage}} = \frac{37.5}{5.2} = 7.2$$

807 Input capacity...11 mmfd.
4 tubes... 4
44 mmfd.
Strays...10
54 mmfd.

$$R_L = X_C = \frac{10^6}{2\pi \times 4.25 \times 54} = 680 \text{ ohms}$$

(7) Gain per tube = $6000 \times 680 = 4.1$

It may be seen that two 807's will supply ample gain for the first stage.

Gain for two 807's = $2 \times 4.1 = 8.2$ (neglecting their output capacity)

On several occasions it was mentioned that in the case of cascaded video stages, or when a modulator drives a capacitative load, the capacity of either the output tubes or the load (whichever is the smaller) could be neglected. This can ordinarily be done provided special coupling circuits are used.

In Fig. 25A, the ordinary shunt compensated amplifier is shown. In this case the total of the input capacity of the second stage, the output capacity of the first stage, and the stray wiring capacity, all appear across the load resistor.

If a pi network, such as shown in Fig. 25B is used, this capacity is split into an output and input capacity of a filter network and a considerable increase in load resistance is possible with an attendant increase in gain. This is sometimes called "series peaking" or "series compensation". In general, it is possible to obtain a more uniform response over the band if a combination of both of the above methods is used.

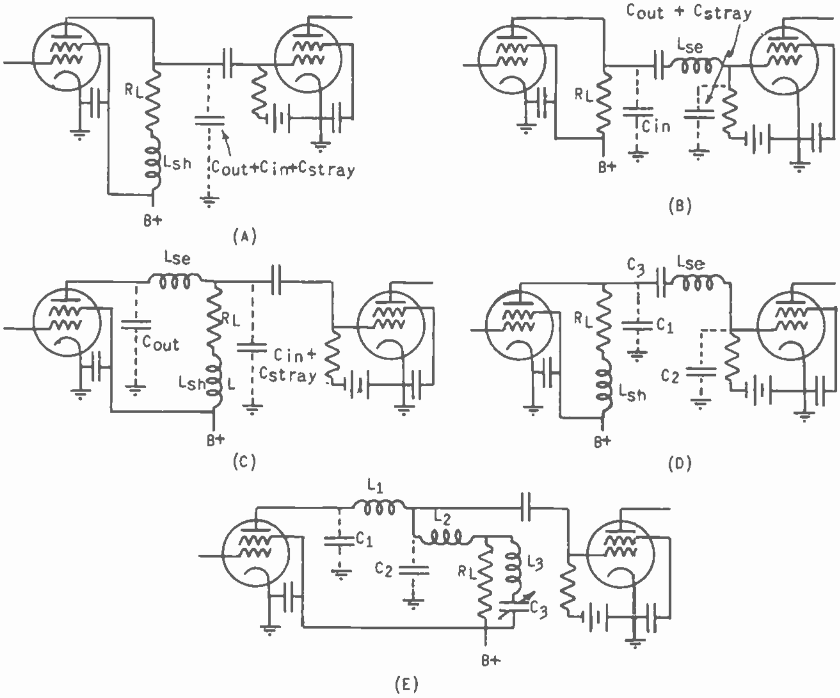


Fig.25 Wide band coupling networks for video amplifiers.

Combination compensation, such as shown in Figs. 25C and 25D, operates with maximum benefit when the capacities on the input and output of the pi network are in the ratio of two to one, or one to two, and the load resistor is connected to the terminal of the network having the smaller capacity. In the case of Fig. 25D, capacity C_2 would be twice the value of C_1. If C_1 and C_2 tend to be approximately equal, the coupling condenser C_3 can be moved to the opposite side of the series coil L_{se}. The capacity of this condenser to ground will then upset the equal capacity of C_1 and C_2. This optimum capacity balance (capacity ratio of 2:1) also holds for the case of straight series peaking.

Under the above conditions, combination peaking will afford a gain of 1.5 to 1.8 times the gain obtainable from shunt peaking, and since the smaller capacity is just 50% of the larger capacity, it follows that the gain obtainable from this type of compensation is the same as if the amplifier had been designed for shunt peaking, neglecting the smaller capacity.

When it is impossible to obtain a capacity ratio of two to one, but equal capacities can be achieved, then the network of Fig. 25E can be employed with an even "flatter" response characteristic than in the case of combination peaking. This is a two-section filter, terminated in a load resistor equal to the capacitive reactance of C_2. Across the load resistor is placed a series resonant circuit, tuned to a video frequency slightly higher than the top frequency that is desired to be preserved. The network is then said to be terminated in an "M derived" section. Since C_1 is balanced to be equal to C_2, the gain that can be achieved is just twice the gain obtainable by the use of shunt peaking with the same total capacity, or conversely, the gain is equal to that obtainable with ordinary shunt compensation when one of the capacities has been neglected.

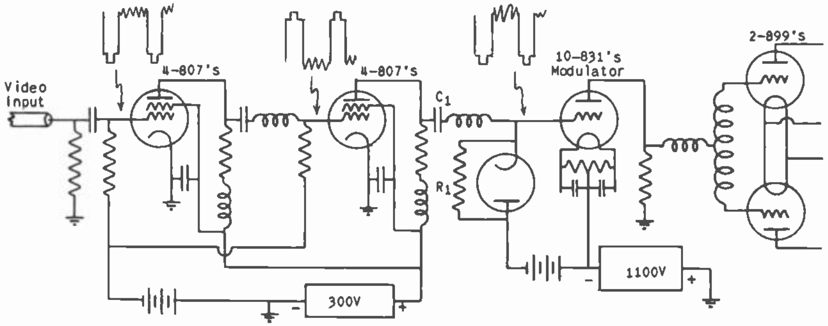


Fig.26 The use of DC restoration circuits in television transmitters.

The fourth consideration in the design of video amplifiers was stated as the problem of maintaining an effective DC coupling in the transmitter. The first and most obvious method of achieving this is to level the signal at the grid of the modulator (that is, restore the DC component with a diode). This is shown in Fig. 26. There are two disadvantages to this method, the first of which is minor and not ordinarily objectionable in practice. The student should refer to Lesson 9, Unit 5, and review the process of DC restoration before continuing.

It will be recalled that the operation of the leveling diode depends upon charging condenser C_1 (of Fig. 26) through the impedance of the diode, and discharging it through R_1. In this manner, a sawtooth voltage is generated and added to the picture signal at the grid of the modulator. If this effect becomes prominent, it not only adds a shading signal to the picture, but may decrease the quality of the synchronizing. In practice, it is seldom bad enough to cause much trouble.

If the modulator must be swung into the positive grid region, which is quite often the case, the leveling problem becomes a different story. Notice in Fig. 26, that the polarity of the signal is such that two types of leveling occur. The diode levels on the synchronizing tips, but the grid current of the modulators cause these tubes to level on the highlights of the picture signal. The diode and the modulator grids will thus vie for control of the DC

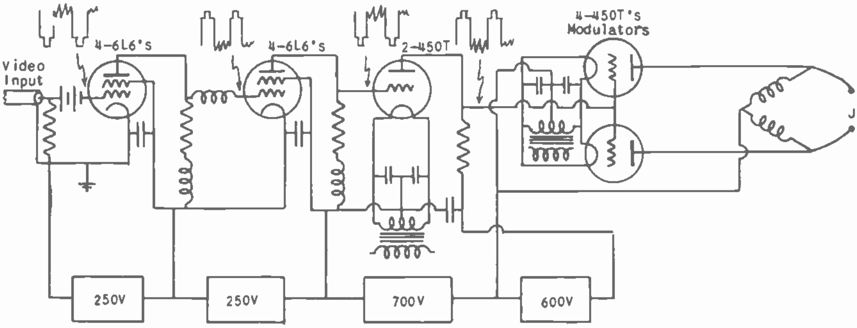


Fig.27 Direct coupling of video amplifiers

level of the picture. In this case, the level of the blanking pulses will be prone to shift in accordance with the amount of picture signal present.

If the whole video amplifier is DC-coupled, as shown in Fig. 27 (typical video amplifier for transmission line modulated transmitter) no trouble will be experienced with lack of DC insertion

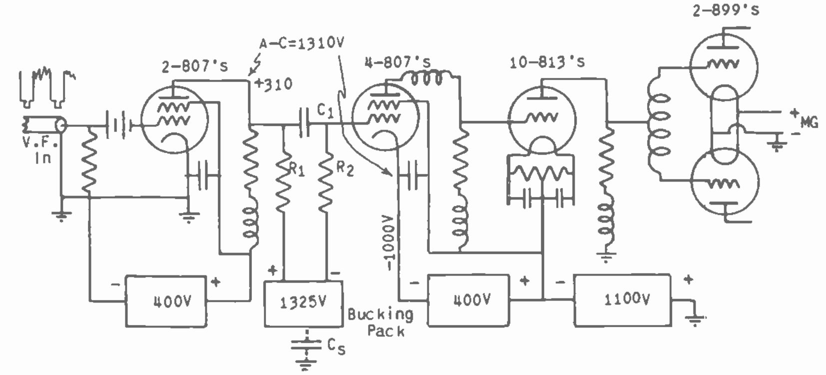


Fig.28 The use of bucking packs to effect a direct coupling between video stages.

when the stages are run into grid current since the DC component is carried all the way through. The difficulty with this method is that the individual power supplies, as well as the cathodes and filaments of the various stages, will be at a potential of perhaps several thousand volts above ground. Care must then be exercised in mounting small parts (particularly metal cased electrolytic condensers) upon the grounded video chassis.

A compromise between the two above methods could be achieved by direct-coupling between the driver and modulator, and employing leveling at the grid of the driver.

Still another method of direct-coupling between successive video stages is illustrated in Fig. 28. In this method, the plate voltage of one stage is offset by a bucking pack connected directly between the plate of one stage and the grid of the next, and adjusted to give the correct bias on the following tube. In order that the capacity of the pack to ground will not result in a loss of high frequency response, it is isolated by two resistors (R_1 and R_2). The high frequencies are coupled between the stages by the capacitor C_1. At low frequencies, corresponding to slow variations in background, the capacity C_1 is not effective, but neither is the stray capacity C_s, so no drop occurs across the coupling resistors (R_1 and R_2) and the pack "floats" up and down with respect to ground.

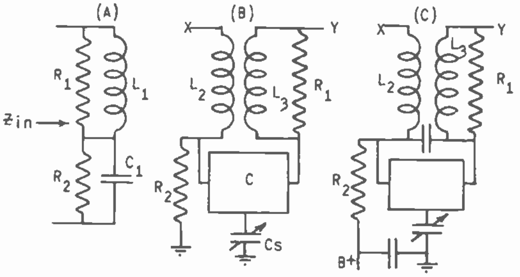


Fig.29 A constant resistance network.

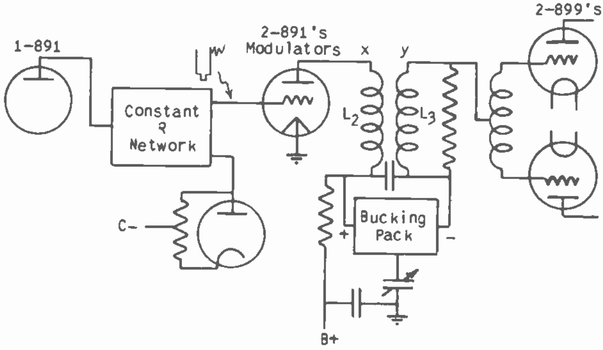


Fig.30 The application of constant resistance networks as inter-stage couplers in video amplifiers.

Due to the decoupling resistors (R_1 and R_2) in Fig. 28, this method cannot be employed where any grid current is apt to flow in the driven stage. A method somewhat similar in principal can be used in this case, based upon the principles of constant resistance networks. The circuit in its final form is shown in Fig. 30, whereas the development of the principle is illustrated by Fig. 29.

It can be shown that the network illustrated in Fig. 29A exhibits the characteristics of a pure resistance whose magnitude is constant at all frequencies when the various components of the circuit are proportioned in a very definite manner as follows:

$$Z_{in} = R_0 = R_1 = R_2 = \sqrt{\frac{L_1}{C_1}}$$

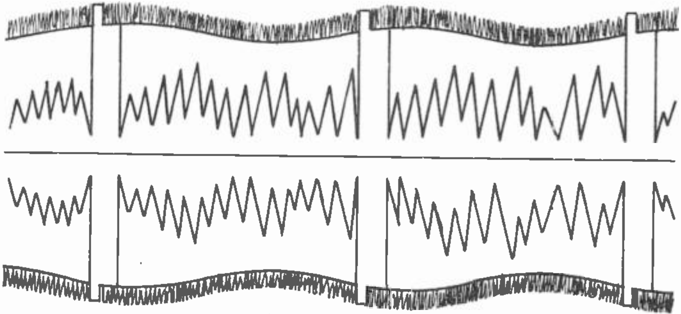
When this is done, the network as a whole can be substituted in a video amplifier for the load resistance in the plate circuit of the tube. This would have no particular advantage over an ordinary resistor except for the development shown in Fig. 29B. Here, the coil L_1 has been split into two parts, L_2 and L_3, each having the same inductance as L_1 and so tightly coupled that they act as a single coil. The capacity C_1 of Fig. 29A appears as C_s in Fig, 29B. The network still behaves as a pure resistance, and the voltage at x also appears at y, due to the tight coupling of L_2 and L_3 which are by-passed to each other at the lower end. The stray capacity of a coupling pack to ground, in conjunction with a small trimmer shown in Fig. 29B and C, and in Fig. 30, replaces the capacity C_1 of Fig. 29A.

The function of the coupling pack in this case is the same as in the case of Fig. 28, except that now the circuit will function in spite of grid currents drawn by the video stages. This is so because of the low resistance of the coils L_2 and L_3. The application shown in Fig. 30 is typical, and could be applied to the transmitter of Fig. 6.

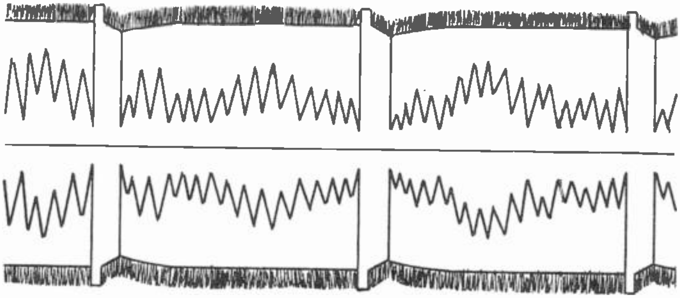
10. MISCELLANEOUS TRANSMITTER REQUIREMENTS. The power supply requirements of television transmitters are stringent. The power supply must not only be free from hum, but must have a very low internal impedance. This is very difficult to achieve in a high voltage power supply. The low impedance loads into which television transmitter tubes must operate, cause the tubes to deliver their optimum performance with a lower than normal plate voltage. This allows us to use a great deal more filtering than we would otherwise be able to do for a given power and a restricted cost. In broadcast transmitters, any residual hum, not filtered out of the power supply, can be degenerated in the transmitter. It is not practical to do this in television service because of the extremely wide video band that must be handled. All high voltage supplies should be as heavy duty as possible with good regulation. Three-phase power supplies are highly desirable as are oversized choke coils. The output filter should consist of at least 10 mfd. capacity.

Any hum on the carrier modulates the blanking and synchronizing level of the transmitter signal as shown in Fig. 31A. This results in a signal which is not only hard to synchronize at the receiver due to the varying level, but is apt to cause waviness in the edge of the picture. Bad offenders in this matter are filaments heated by alternating current, such as the modulator or modulated amplifier. In general, unless the filament is a polyphase one, it is desirable to operate the high power tubes from a DC generator. Any residual hum can be balanced out by creating an equal hum of opposite phase and inserting this at low level in the video amplifier in the same manner that shading signals are inserted in control room amplifiers. This subterfuge is to be used as a last resort, since such circuits are always awkward and apt to get out of adjustment.

Regulated power supplies should be used in all the low-level video stages, and if possible, each stage should have a separate supply. This facilitates trouble elimination, which in video amplifiers quite often occurs as signal components across the power supply leads, due to power supply resonance or poor regulation.



(A) Hum On Carrier



(B) Poor Power Supply Regulation

Fig.31 The effect of hum and poor power supply regulation upon the performance of picture transmitters.

It is not economically practical to use a regulated high voltage power supply to supply the modulator and the modulated amplifier. The best possible unregulated supply is not always good enough, and the effective internal impedance of the power supply causes some phase shift at the lower frequencies, due to the small condenser (10 mfd. or so) in the output stage of the filter. Such a condition in the supply for the modulator would cause an increase in the low frequencies, while the same condition in the plate supply to the modulated amplifier would cause a negative plate modulation to occur, which would result in a deficiency of low frequencies. The latter effect is generally more prevalent and is shown in Fig. 31B. This particular effect can be compensated for by properly adjusting the decoupling circuits in the low-level video stages in exactly the same manner as in correcting the poor low frequency response due to coupling condensers in receiver circuits.

While the above type of distortion can be corrected, another effect which arises from the same condition is not so easily overcome, and for this reason, mediocre power supplies should not be constructed with the idea in mind of correcting the distortion which they introduce. Poor regulation in the power supply will cause the output voltage to drop as the current increases. The lower voltage causes the overall output of the transmitter to be reduced, and the effect is that the transmitter has a smaller response to DC variation than to AC variation. Dissimilar response to AC and DC causes the blanking level of the transmitted picture signal to shift as the type of picture is changed. If this effect is present to any great extent, the blanking level might shift enough to cause the video signal to exceed the modulation capabilities of the transmitter and saturate the synchronizing impulses. The effect is exactly the same as a compromise between AC and DC coupling in the video amplifier and the effect of Fig. 1 occurs, but to a lesser degree. This defect is quite apparent when the picture is removed and the background level adjusted to the blanking line. Under such extreme conditions, poor regulation in the power supply will be evidenced by shifts in the blanking level of the RF envelope, and probably by a slight change in the magnitude of the synchronizing pulses.

When video stages are to be directly coupled (this is sometimes desirable when it is necessary to supply grid current) it becomes necessary to operate several power supplies in series (See Fig. 27), each power supply furnishing its individual video stage. Under these conditions, ordinary power supplies are not satisfactory. In order to prevent coupling between stages, regulated power supplies are a necessity.

When it is desired to operate direct-coupled amplifiers in the manner shown in Fig. 28, the bias voltage on the second stage is the difference between the voltage A-C and the potential of the bucking pack. Since this differential voltage is a small percentage of the total output of the pack (15 volts out of 1300), any variation in one of these packs will be applied directly as a signal on the grid of the video stage. If the output of the bucking pack changed only 1 volt (a small fraction of one per cent), this signal through the video amplifier would modulate the transmitter several per cent. Obviously, only the very best types of regulated packs are suitable for this service.

Since the video amplifiers and modulators carry frequency components as high as 4 mc., they must be placed close together and close to the modulated stage in order to reduce the stray capacities of the connecting leads. Under these conditions, the video amplifier is close to a strong RF field. Extreme care must be taken to shield the video amplifier and correctly isolate the power supply leads. Otherwise, RF pickup will drive the grids of the video amplifier into the positive region, causing them to "level" on the peaks of the RF wave and thus shift the operating bias. This bias shift gives rise to amplitude distortion by operating the tubes on the wrong portion of their characteristic curve.

If a leveling tube is used such as shown in Fig. 26, RF pickup may cause this tube to level on the RF signal rather than on the tips of the synchronizing, and this would make it impossible to restore the DC component of the video signal by leveling.

If the sound transmitter and picture transmitter are operated close together, another type of trouble which might arise is pickup of the output of one transmitter in the circuits of the other. If the output of the sound transmitter is picked up by the video amplifier, a certain amount of rectification will take place, due to the unavoidable non-linearity in some of the stages. This causes cross-modulation between the audio signal and the video signal, and the background of the received picture will vary at audio modulation frequencies. A similar but more curious trouble has occurred by rectification, in the control tubes of the regulated power supplies, of the RF voltage picked up in these supplies. Such pickup and consequent rectification will modulate the supply voltages at audio frequencies, and in turn inject this signal into the video amplifier with exactly the same result as in the previous case.

Excellent shielding is necessary in television picture transmitters. All of the RF stages, and particularly the driver stage, should be shielded. Any radiation from the driver which is picked up at the receiver will result in a steady signal which is unaffected by picture modulation. The effect is exactly the same as if the modulation *capabilities* of the transmitter had been reduced in the negative direction; that is, the direction toward minimum carrier. The amount of radiation that escapes from an unshielded stage is astounding. In some cases, this radiation might be as high as 10% of the maximum radiation possible to produce from the same stage feeding a correctly adjusted antenna.

If the driver has poor RF regulation, there will appear in its tank circuit, a modulated RF signal, having opposite polarity to the signal which appears in the tank of the final modulated amplifier. If any stray radiation exists from the driver in this case, the receiver will pick up both a positive and a negative signal. Since a positive signal is transmitted via the feed line and antenna system, its path length to the receiver is different from that of the stray signal escaping directly from the driver. In this case, two images will be received, having opposite polarity and displaced on the face of the Kinescope. This effect is particularly important in the case of transmission line modulation, where the RF source has a relatively high amount of circulating energy, compared to the amount that is actually delivered to the antenna. In this case, any stray radiations from the RF source (that is, the oscillator or power amplifier which feeds the transmission line modulator) will be an appreciable percentage of the power radiated by the antenna.

In addition to the necessity for good shielding, plus the necessity for low capacity in the modulator and video amplifier, and the reduction to an absolute minimum of all stray capacities across the wide band tank circuit, all of the construction details discussed in the lesson on television sound transmitters should be applied in the construction of television picture transmitters. At the conclusion of this lesson, the student should review the previous one to refresh himself on the principles outlined therein.

11. ADJUSTMENT TECHNIQUE AND MONITORING PROBLEMS. Picture transmitters are adjusted by a cut and try process, but it is important to have a thorough understanding of the principles discussed in this lesson; particularly the considerations in the design of the various types of transmitters. With this understanding, the effects of changes can be anticipated, and the results of the changes can be better understood and analyzed.

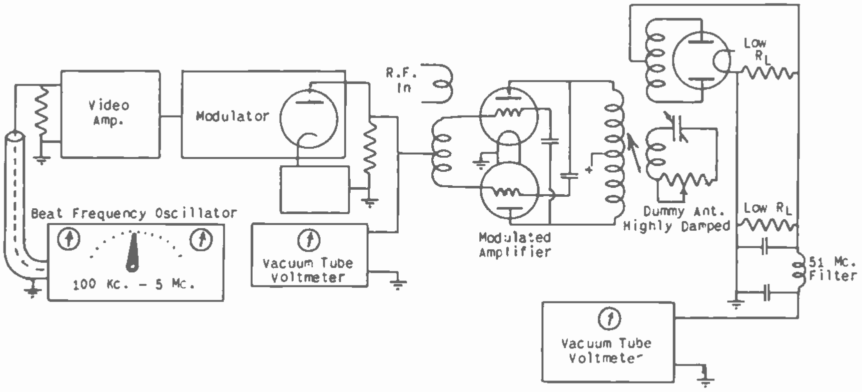


Fig.32 The setup for adjusting the bandwidth of the radio frequency stages of a picture transmitter.

In the case of a plate modulated transmitter which is to be operated at a previously calculated plate voltage, the load on the tank circuit is adjusted by observing the bandwidth of the transmitter in the following manner. A load is obtained which is as nearly a pure resistance as possible. This can be either a previously adjusted antenna system or a dummy load. The coupling to the transmitter is then varied in definite steps, and for each step, a frequency response of the transmitter is measured from the output of the modulator to the RF load. A vacuum tube voltmeter is suitable for measuring the output of the modulator, and a heavily loaded diode detector plus a peak-reading voltmeter can be coupled to the antenna circuit for measuring the rectified output of the transmitter. (see Fig. 32) The coupling is adjusted until the required bandwidth is achieved, after which it is fixed and not touched thereafter. The excitation and bias are then varied to obtain maximum linearity at maximum output, observed by an oscilloscope when a low frequency audio generator is used to furnish input to the modulator. With the exception of the RF load adjustment, all other adjustments follow standard broadcast practices.

The grid modulated transmitter is adjusted in exactly the same manner except that the excitation and bias adjustments are more critical if we are to obtain maximum output at maximum linearity without exceeding the ratings of the tube.

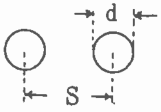
The linear amplifiers are adjusted in the same manner except that individual stages are adjusted one at a time, starting at the output stage and working backward. In this case, the response of the linear amplifier must actually be plotted over a wide band of frequencies, since in the case of an overcoupled circuit, we are not after a flat response, but are after a particular shaped response, as shown from previous calculations.

If the transmitter has been adjusted with the actual antenna connected in the circuit, then no further adjustments are necessary. If the transmitter was adjusted by the use of a dummy load, then all the voltage and current readings of the transmitter must be recorded. The real antenna is then substituted for the dummy antenna, or dummy load, and with all other adjustments of the transmitter remaining the same as previously, the coupling to the antenna is increased until the plate current of the output amplifier is exactly the same as the value recorded from the adjustment with the dummy load.

The adjustment of a transmission line modulated system is radically different. Consider the circuit of Fig. 15. The spacing of line L_m is adjusted so that the line will have a surge impedance of approximately 250 ohms. Line L_a is adjusted until the antenna reflects approximately 40 ohms at junction J. This impedance is not measured, but it is assumed that the antenna load is known approximately, and the spacing of the line L_a is adjusted to transform this impedance back to a 40-ohm junction impedance, according to the following relations which we have used several times before:

$$\frac{Z_{in}}{Z_o} = \frac{Z_o}{Z_{out}}$$

$$\text{Where } Z_o = 276.5 \log_{10} \frac{2s}{d}$$



The next step in adjustment is to short-circuit the plates of the modulator tubes and adjust the coupling from the tank circuit or the spacing of the line L_t until maximum output is obtained from the transmitter as indicated by oscilloscope or antenna ammeter. The short-circuit is then removed from the modulator tubes and the length of L_m is adjusted, with the modulators biased off, until the antenna output falls to zero. The short-circuit should then be replaced and the first step repeated, this time with the plates of an oscilloscope connected directly across the antenna feed line, or excited from a tuned circuit which is link-coupled to the antenna feed line. The deflection of the oscilloscope for maximum transmitter output is recorded.

Next, an AF signal is inserted on the grids of the modulator and increased until maximum allowable grid current on the modulators is reached. The audio signal should also be applied to the horizontal amplifier of the oscilloscope if the vertical plates are connected to the RF output of the transmitter. The trapezoid then traced on the screen of the tube should have an envelope similar to the curve in Fig. 18A. If a good minimum carrier is reached on the negative peaks of modulation, but if the positive peaks of modulation do not drive the output to within 90% of the maximum output recorded on the oscilloscope for modulator-short condition, then the impedance at the junction should be lowered by decreasing the line spacing of transmission line La. If the reverse is true; that is, if it is possible to secure a fairly good maximum output from the transmitter, but the negative peaks of modulation do not reduce the transmitter output to nearly zero, then the impedance at junction J should be increased by increasing the spacing of the transmission line L.

When the transmitter is properly adjusted for AF response, no further adjustments are necessary except in the video amplifier, since the transmission line modulation method does not normally limit the band width below that required for 4.5 mc. top video frequency.

It is highly desirable to have at the transmitter a monitor fed by the radiated output, but this is extremely difficult to do. Stray fields from the transmitter are prone to cause more pickup in the RF monitor and the feed line to the monitor antenna, than is picked up directly in the monitor antenna itself. Extreme care in shielding must be used and the monitoring antenna must be located in the clear and fairly close to the transmitting antenna, say, within 100 yards. The feed line to the monitor should be of the low-loss type in order to produce a large signal from the antenna itself in comparison to the stray fields delivered to the monitor. The monitor is preferably a one or two-stage RF amplifier and diode detector, rather than the conventional superheterodyne receiver. If the antenna response is known to be flat, the simplest way of monitoring is to connect a diode and a pickup loop to the feed line between the transmitter and the antenna. The video signal output of this diode detector is then fed to a video amplifier and applied to the screen of the picture monitor. The picture monitor should be arranged to observe not only the rectified output of the transmitter, but also the signal on the line input to the video amplifier, and on the grids of the modulator with a three-position switch, in order that a direct comparison may be had.

We should also have an oscilloscope which constantly monitors the RF waveform of the transmitter. The plates of the oscilloscope should preferably be connected directly across the antenna transmission line or through a half-wave section of line to the antenna transmission line, but can be, for second choice, fed by a tuned circuit which is link-coupled to the antenna transmission line. The disadvantage of this latter method is that the bandwidth of the tuned circuit is small and the waveform is not a perfect replica of the transmitter performance. If the tuned circuit were sufficiently damped to pass a wide band of frequencies, then it would have to dissipate considerable power in order to develop the several hundred volts necessary for complete deflection of the oscilloscope. This is wasteful of energy, and is to be avoided if possible.

The requirements for frequency monitoring have been covered in the previous lesson, and will not be reviewed here. A convenient means of checking the overall bandwidth of the transmitter is to apply a sweep frequency video generator to the input of the video amplifier or the line input to the transmitter. If the oscilloscope plates are connected directly to the antenna transmission line, the bandwidth can be observed by noting the percentage of modulation over the various portions of the sweep cycle. Alternatively, a detector could be applied to the video amplifier of the monitor to rectify the sweeping video frequency so that the envelope may be applied to an ordinary oscilloscope (see Fig. 33). The trace of this envelope over the sweep cycle is a plot of the overall frequency characteristic of the transmitter and monitor.

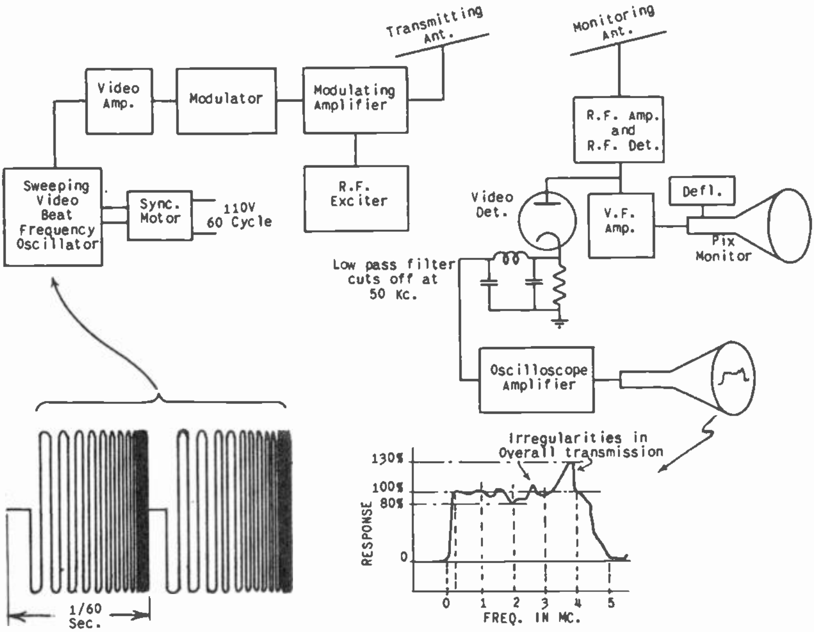


Fig.33 The monitoring means for observing the overall frequency characteristic of the transmitter.

The real proof of the transmitter performance, however, lies in a visual inspection of the received picture, carefully observing its overall picture quality as compared with the same monitor when fed by the input to the transmitter.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. (A) Why is it desirable to transmit the DC component of the video signal? (B) What is the effect upon the transmitter if the video stages are AC-coupled?

2. Sketch the basic circuit for series and for parallel plate modulation of television picture transmitters. Indicate by sketch the polarity of the picture signal at the grid of the modulator.

3. Sketch the basic circuit of a grid modulated picture transmitter, and indicate by sketch, the polarity of the picture signal at the grid of the modulator.

4. Sketch the basic circuit of a transmission line modulated picture transmitter, and indicate by sketch, the polarity of the picture signal at the grid of the modulator.

5. What method or methods are employed to maintain the wide band characteristics of linear amplifiers in television transmitters?

6. What is meant by a single sideband filter? Where and why is it employed?

7. What are the comparative advantages and disadvantages of grid and plate modulation for television?

8. What are the chief differences in the mode of operation of grid modulated transmitters for standard broadcast stations and television service?

9. Sketch a video amplifier to couple between the line (from the control room amplifier) and the modulator of a grid modulated stage including the following: leveling (DC restoration) at the grid of the modulator; picture polarity at each stage in the chain; combination series and shunt peaking (high frequency compensation).

10. Sketch several methods of preserving the DC component when coupling between two video stages.

Notes

(These extra pages are provided for your use in taking special notes)

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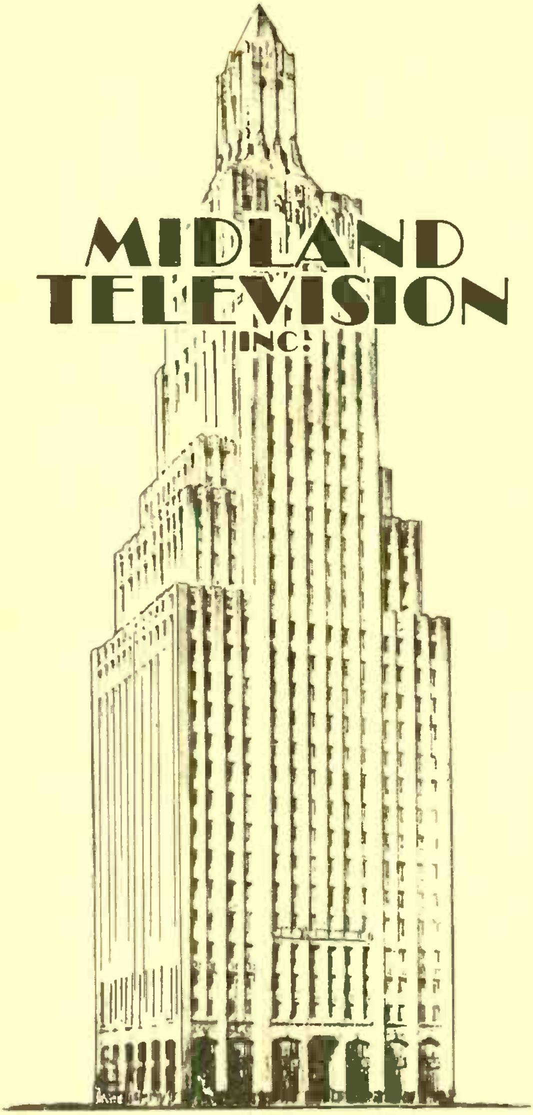


POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT
NO.
7

TELEVISION TRANSMITTING ANTENNAS

LESSON
NO.
6

FEAR

.....the shadow of ignorance.

Somewhere...some time we read the following words, written by a psychologist whose name has slipped into a corner of our brain. Some day it will 'pop' up without warning. Those words are.

"Fear is only another name for ignorance".

Men who fear the future, usually do so because they feel that they lack the knowledge necessary to hold the success they have achieved. And, if they have failed to achieve any degree of success, they fear the future because the very same lack of knowledge has prevented them from saving money to take care of their needs, as age, slowly and relentlessly creeps upon them.

Just as fear is the shadow of ignorance, self confidence is the shadow of knowledge.

A short time previous to the writing of this little story, a student approached one of our engineers who was patiently constructing an elaborate piece of television testing equipment. On the bench before this engineer, was an amazing conglomeration of wires and parts. For weeks he had been striving to attain certain results, but so far had not been successful.

Gazing at the array of wires and parts, the student asked, 'Do you think you will ever get the right combination?' The answer was, 'Of course I will. It just takes time'. The engineer was CONFIDENT because he possessed knowledge. And today that piece of equipment is functioning perfectly in our laboratory.

While you may not realize it, each bit of knowledge that you glean from the lessons you are mastering, IS ADDING MATERIALLY TO YOUR SELF-CONFIDENCE. You should have no need to fear....FEAR. And the greater the extent of your knowledge, the more heroic the shadow of SELF CONFIDENCE that you cast before you, will be.

Lesson Six

TELEVISION TRANSMITTING ANTENNAS

"Regardless of how perfect or efficient a television station's pickup and transmitting equipment might be, good results cannot be secured unless the transmitter delivers its energy into a properly designed antenna.

"Because of the importance played by antennas and filters, it is highly important that you study this lesson carefully. I am sure that you will spend the necessary time to master it completely."

1. ULTRA-HIGH FREQUENCY PROPAGATION. In a previous lesson, the student learned that broadcast coverage could be divided into three areas: First, the area served by the ground wave, second, the area in which the ground wave and sky wave combined to produce a strong but unreliable signal subject to considerable fading, and third, an area served by the sky wave, which enjoyed fairly reliable signals at night, but which could not be reached in the day time.

The characteristics of ultra-high frequency signals must be classified differently. Since the antennas are elevated to great heights, for reasons which we shall presently consider, there is a local area served by a combination of a direct ray and a ray reflected from the ground. The second area which an ultra-high frequency signal may serve is the area immediately beyond the horizon, and the signal reaching this area is due to diffraction at the earth's surface. The power of television transmitters is such that this area does not ordinarily receive a reliable signal, but it must be considered, inasmuch as interference would always be caused with another transmitter in this area operating upon the same frequency. The third area into which ultra-high frequency signals may be propagated is any territory beyond approximately 200 miles, and in such territory the signal serves no useful purpose but may cause *occasional interference* to other transmitters operating upon the same frequency.

The transmissions into distant areas are due to two causes. The first cause is refraction of the wave in the atmosphere under certain conditions; and the second is transmission due to the sky wave under extreme conditions of ionization in the lower ionosphere.

In modern broadcast installations, the antenna is located close to or even directly connected to the ground. Under these conditions, the service area of a station located at the high frequency end of the broadcast band is only a small fraction of the area served by the ground wave from a station located at the low frequency end of the band. This difference is due to the increased absorption of the higher frequency by the ground, and by various objects in the path of propagation. As we go higher in frequency, this effect is magnified until it would be impossible to obtain satisfactory coverage from a low antenna, operating at a frequency of 44 to 104 megacycles. On these frequencies, it is necessary to elevate the antenna as high as possible in order to secure a direct path between the transmitting antenna and the receiving antenna as shown in Fig. 1.

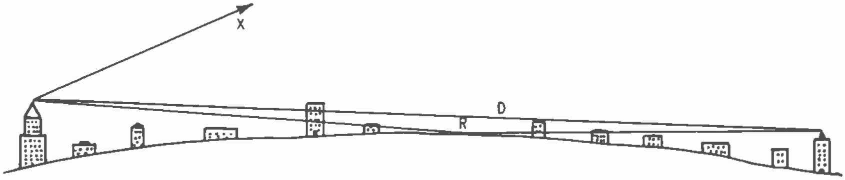


Fig.1 Normal transmission path between transmitter and receiver.

Unfortunately there exists, besides the direct path from the transmitter to the receiver, indicated by D, an additional path R, which is a ray reflected from the ground. High angle radiation such as X, which is detrimental at standard broadcast frequencies, is not important on the ultra-high frequencies, except that it is desirable to utilize this energy by suppressing it along the direct path D. It does not combine with the direct path to produce interference, since it is rarely returned to earth, and then only at great distances. The reflected ray R suffers a phase reversal upon reflection. If the path D and the path R were of the same length and no losses were encountered upon reflection, the direct and indirect rays would cancel at the receiving location. However, the path R is somewhat longer, and therefore, a slight phase difference exists between the direct ray and the reflected ray as given by the following formula:

$$\theta = \frac{4\pi ah}{\lambda r} \qquad (1)$$

where: θ represents phase difference between direct ray and reflected ray, h is the height of the transmitting antenna in meters, a is the height of the receiving antenna in meters, r is the distance between the transmitter and receiver in meters, and λ is the wavelength in meters.

The direct field from a half-wave dipole is: $E = 7(\sqrt{W}/r)$, where W is the watts radiated. Combining this equation with equation (1), we arrive at the following expression for the field at any distance from the transmitter.

$$E = \frac{7\sqrt{W}}{r} \times \frac{4\pi ah}{\lambda r} \quad \text{or,} \quad E = \frac{88\sqrt{W}\,ah}{\lambda r^2} \text{ volts per meter} \qquad (2)$$

Thus, if we assume perfect transmission and perfect reflection from level ground, the field at the receiving location will vary directly as the height of the transmitting antenna, directly as the height of the receiving antenna, directly as the square root of the power, inversely as the wavelength, and inversely as the square of the distance from the transmitter. If we double the height of either the receiving antenna or transmitting antenna, we double the received field. If we increase the transmitter power by four, we double the strength of the signal at the receiver. Theoretically, if we go to a higher frequency, the field strength increases. On the other hand, a receiver located at four miles from the transmitter would receive only one-quarter the field strength of a receiver located two miles from the transmitter.

In the practical case, a number of factors influence the accuracy of formula (2). Although the reflection from the ground is not perfect, the efficiency of reflection is quite high (if this were not so, a greater field strength would be delivered at the receiver than formula (2) indicates). Scattering and absorption in the open country tends to reduce the magnitude of the direct wave. This is particularly true if the direct wave must pass through any obstruction such as a building, some trees, or over the brow of a hill. For this reason, practical values of field strength often approach the value indicated by formula (2), but seldom exceed it. In general, observed values of field strength lie between 30% and 60% of the values indicated in formula (2). Variations do occur, however, between 10% and 100% of the values indicated in this formula. The chief reason for such variation is the great difference in quality of locations, such as; the height of the receiving position (depending upon whether it lies on a hill or in a valley), and the flatness of the ground and the number of obstructions that lie between the transmitter and receiver.

On the higher frequencies, the attenuation of the ground and intervening objects is markedly more than at the lower frequencies. Consequently, although formula (2) indicates that a higher frequency will yield a greater field at the receiver, this may or may not be true, depending upon the elevation of the transmitter and the openness of the country between the transmitter and the receiver. One survey indicated that 30 megacycles was considerably superior to 100 megacycles when the transmitting site was somewhat obstructed. In a later survey with the transmitter located at the top of the Empire State Building, a transmitter frequency of 120 megacycles gave an average field of 13% higher than the field from a transmitter operating at 81 megacycles. The actual magnitude of the difference is not very important. The significant fact is that the higher frequencies; that is, the higher channels of the television band are not as reliable and do not give as uniform a coverage as the low frequency channels.

Since elevation of the antennas above a level ground is an important factor in formula (2), it is interesting to plot a cross-section of a city in various directions to see how closely the assumption of level ground is justified. As an example, Kansas City has an average terrain with many rolling hills, but no real large ones. It can be seen from Fig. 2 that a great deal of variation of field strength may be expected from the nominal value calculated by formula (2), even if absorption in the obstructing buildings and trees is neglected. Although none of the hills shown in this figure completely obstruct the line of sight between the transmitter and receiver, nevertheless, it can be seen that the difference between the direct path and the ground reflection path does not follow the simple relation of formula (1), and variations on the order of 10 to 1 may be expected for individual locations.

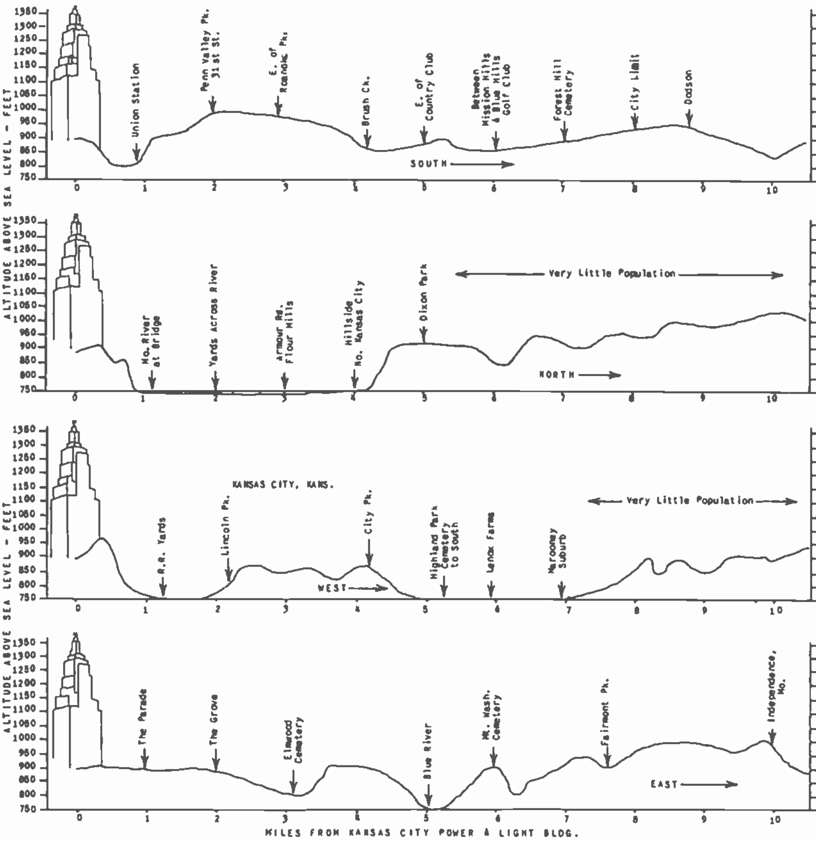


Fig.2 Elevation profile of Kansas City.

The inverse square law of formula (2) is applicable for all distances within the line of sight. The line of sight between the transmitter and receiver is limited to the optical horizon. The optical horizon is the distance AB of Fig. 3, and it is the maximum distance that an observer could see if he were standing at the exact location of the transmitting antenna. This distance is given by formula (3) below.

$$D = 1.22\sqrt{H} \qquad (3)$$

where: D is the distance in miles, and H is the height of the antenna in feet. If the receiving location is on the crest of a hill, or a sufficiently tall antenna structure can be erected, then the line of sight distance between the transmitter and receiver can be increased somewhat as shown by distance AC of Fig. 3. In this case, the distance to the optical horizon is calculated for both the transmitter and receiver by using formula (3) and the two distances are added together. A receiver located at point D also receives some signal, due to the diffraction of the wave in passing over the surface of the earth at point B. The traveling wave, after passing point B, is scattered and tends to fill the area XBD so that a receiver located at point D would receive some signal. This diffraction effect is more prominent on the low frequencies than on the high frequencies.

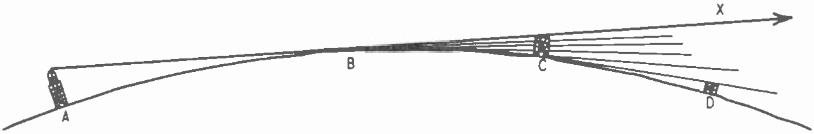


Fig.3 The optical horizon.

Buildings and other obstructions, as well as the curvature of the earth, do not cast an appreciable shadow for frequencies on the order of 1 to 20 megacycles, but at 100 megacycles, the shadowing effect is very much greater, while at frequencies corresponding to light, very little diffraction takes place and sharp shadows are cast by small objects.

For locations beyond the optical horizon, the field varies as the inverse 3.6 power, instead of the inverse square if the transmitter frequency is 41 megacycles. This means that the signal is attenuated much faster beyond the optical horizon than up to it. For frequencies around 92 megacycles, the signal strength falls off as the inverse fifth power beyond the horizon. Thus, the chances of receiving a television picture beyond the horizon are much greater for the case of a 40 megacycle transmitter than for a 100 megacycle transmitter. A comparison between the two is shown in Fig. 4.

Ordinarily, transmitters are designed to give a usable signal at the optical horizon. To produce a usable field beyond this point would require an amount of power out of proportion to the increased service area. If the service area is quite small, and a good transmitting location is available, there may not be sufficient population to warrant generating a usable signal at the horizon, and the amount of power in that case would be determined by the area which

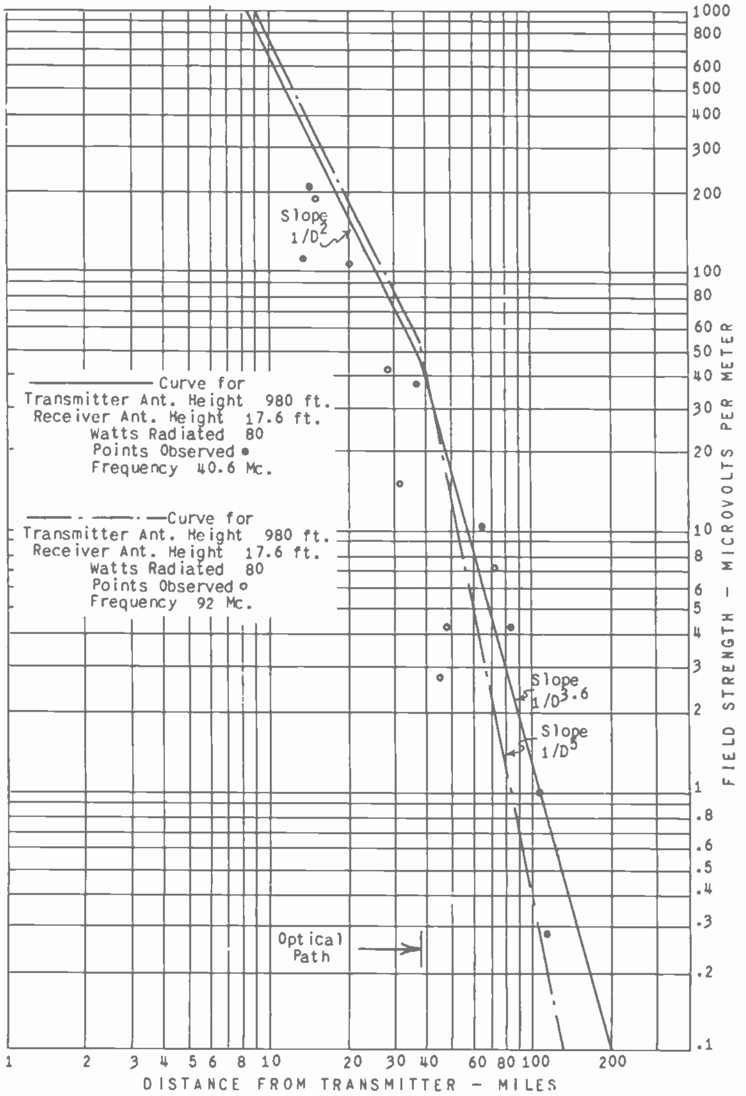


Fig.4 The effect of transmitter frequency upon field strength.

is considered the prime service area. An example of this was given in the preceding lesson, and the student should review at this time Section 2, Lesson 5, Unit 7.

2. HORIZONTAL VS. VERTICAL POLARIZATION. By "polarization of a wave" is meant the direction of the electrostatic lines of force in the field about an antenna. Thus, if an antenna is lying in the

horizontal plane and electrostatic lines of force exist between opposite ends of the wire, the lines of force also lie in the horizontal plane, and the waves are said to be horizontally polarized. The converse is true for vertical polarization. At broadcast frequencies, the polarization is not particularly important, since the average receiving antenna contains both a horizontal portion and a vertical down lead, so that the wire will pick up either horizontal or vertical polarization. Furthermore, reflections from the sky cause the wave to arrive at the receiver with random polarization. This is not the case for ultra-high frequency transmission. Receivers operated at ultra-high frequencies use an antenna which is small, and the lead-in is either carefully shielded, or arranged in such a way that it picks up a minimum amount of energy. Furthermore, there is less scattering and less rotation of the plane of polarization at television frequencies than at broadcast frequencies. For this reason, it is necessary to have the receiving antenna lying in the same plane at the transmitting antenna, barring any unusual circumstances of reflection.

Since the transmitting and receiving antenna must have the same plane of polarization, it is necessary that the transmitter polarization be standardized in order that a receiver may pick up more than one transmitter. The optimum polarization has been a strongly debated question. Witness of this is the fact that England and certain European countries have decided that vertical polarization is superior, while in the United States, horizontal polarization is the accepted standard. Besides horizontal and vertical polarization, circular polarization is also possible. Circular polarization is generated by feeding both a horizontal and a vertical antenna 90 degrees out of phase in such a way as to produce a wave whose plane of polarization rotates in the same way that the poles of a magnetic field rotate in a two-phase motor.[1]

The choice between horizontal, vertical, or circular polarization is influenced by four factors: First, is the relative field strength that may be produced by a simple antenna serving a typical metropolitan area. Second, is the signal-to-noise ratio created by either type of polarization, based upon equal fields. Third, is the indirect path propagation efficiency, which should be low compared to the direct path propagation for maximum signal quality. Fourth, is the ability of the wave to fill in the area behind an obstruction, and thus create the minimum amount of shadow. Of these four, the second and third considerations are by far the most important.

Before entering into a discussion of the comparative merits of horizontal, vertical, and circular polarization, it should be realized that any pickup in the receiver itself, or in the down lead, will nullify the effectiveness of the choice, since the random polarization of these components will pick up external noise, regardless of its polarization, or will pick up signals whose polarization has been changed due to reflection from some building or object. When the direct signal is reflected from an obstruction, it is not

[1] Refer to Lesson 8A, Unit 3.

always reflected with the same polarization as the incident wave. Thus, we find that on the average, a horizontally polarized transmission will be received with an average vertical component, or from 25% to 30% of the horizontal field strength. Likewise, a vertically polarized transmission is generally received with a horizontal component on the order of 20% of the vertical field strength.

Tests have shown that horizontally polarized waves are, in general, received about 20% stronger than vertically polarized waves throughout the television band. This is not a significant difference, especially when the difference in the complexity of the transmitting antenna is considered. In the next section we shall learn that in order to obtain uniform coverage in all directions from the antenna, a more complicated antenna system is required in the case of horizontal polarization than for the case of vertical polarization.

Although the subject has not been thoroughly investigated, at the present there appears to be an advantage in horizontal polarization over vertical polarization with respect to signal-to-noise ratio. Ignition interference and other sources of extraneous noise of the same general character are polarized chiefly in the vertical plane. For this reason, horizontal polarization of the television wave allows the receiving antenna to be polarized horizontally in order to discriminate against interference.

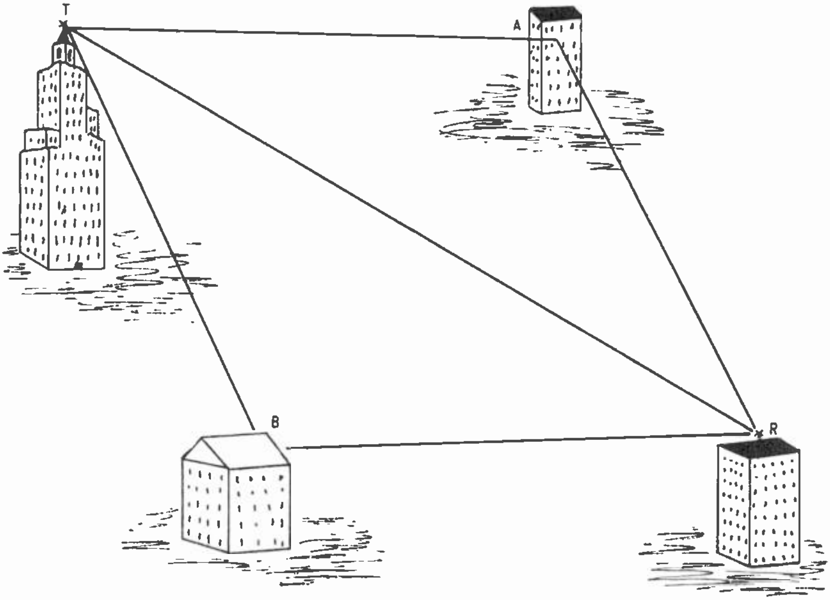


Fig.5 Undesirable multiple path propagation.

When a direct line of sight is available between the transmitter and the receiver, the direct wave between these two points is quite strong. In addition to this wave, however, there are numerous other paths that the signal may take. The reflection coefficient of stone masonry is quite high, and the signal may be reflected from numerous buildings and other objects and arrive at the receiver in phase or out of phase with the direct path wave. This is illustrated in Fig. 5. The indirect path may be TBR or TAR, or numerous others. Whether the indirect signals arrive in phase or out of phase with the direct signal depends not only upon the relative lengths of the transmission paths, but also upon the frequency of the transmitter. Since the television transmitter covers a wide band of frequencies, it generally happens that the direct path and indirect path add up for some of the sideband frequencies and cancel for others. The wideband response of a typical antenna installation at a residence fourteen miles from the transmitter is shown in Fig. 6. Notice that the variations are highly irregular, indicating several transmission paths. The transmission for the case of horizontal polarization is more uniform than the transmission for the

Fig. 6 Typical wideband transmission for horizontal and vertical polarization.

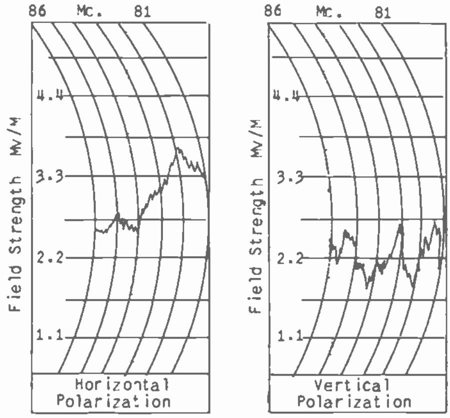
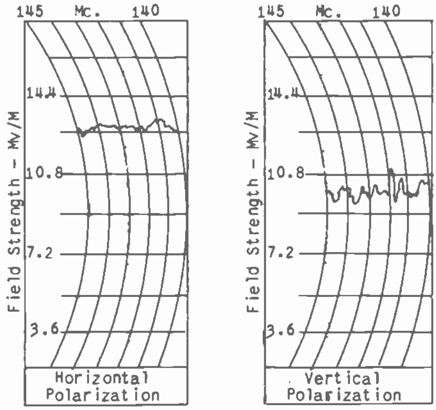


Fig. 7 Wideband response for an excellent receiver location.



case of vertical polarization at this particular installation. Even in the open field, perfectly free from obvious obstruction, considerable variations over the band take place as shown in Fig. 7. In this case, as in the previous one, horizontal polarization is superior to vertical polarization. In extremely poor locations, such as illustrated by the field strength graph of Fig. 8, neither horizontal or vertical polarization will furnish fields sufficiently free from extraneous signals to be usable.

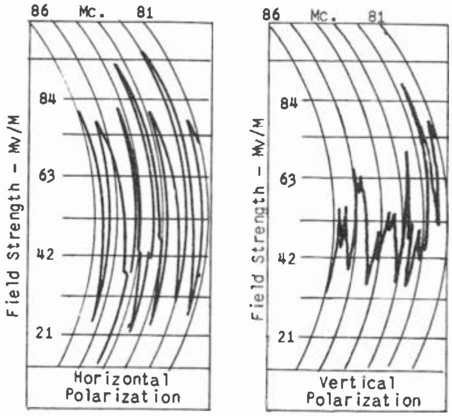


Fig.8 Wideband response when indirect path propagation predominates.

Suppose that in Fig. 8 the carrier is to be located at 81 megacycles, then the upper sideband contains amplitude variations exceeding 4 to 1. If a television picture is to be received from a transmitter supplying this much variation in field strength, the picture will be tremendously distorted, since after detection, these sideband frequencies become the video frequencies constituting the picture signal. Fig. 9 is reproduced from Lesson 1, Unit 6, to show the effect of multiple path propagation upon the received picture.

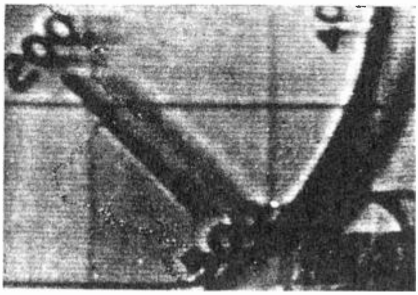


Fig.9 Effect of multiple path propagation upon the received picture.

It has been found that in most locations the indirect path propagation efficiency is better for vertical polarization than for horizontal polarization. Consequently, the indirect interfering signals are from 50% to 60% stronger for the case of vertical polarization than for horizontal polarization. It has also been found that the indirect interfering signals are from 10% to 20% stronger at 140 megacycles than at 80 megacycles.

The phenomena of diffraction is enhanced when the polarization is in the same plane as the edge of the object around which the signal must pass. Therefore, if tall buildings intervene between the transmitter and the receiver, diffraction takes place over the tops of the buildings in the case of horizontal polarization, and around the sides of the buildings in the case of vertical polarization. From this standpoint, there seems to be little choice between the two. However, in the case of intervening hills, the advantage lies with horizontal polarization in that it tends to fill in the valleys slightly better than does vertical polarization.

Although there is some disagreement between different investigators, horizontal polarization has become the accepted standard, because of its superiority in the four points just outlined.

3. ANTENNA DIRECTIVITY AND GAIN. Whenever possible, television antennas are so designed that they concentrate their energy in useful directions. Such design, however, must be carried hand in hand with the design of the wideband characteristics of the antenna system. Any directivity which is achieved at the expense of the antenna bandwidth is undesirable.

At present, antenna directivity is solely for the purpose of concentrating the energy toward the main service area, since this realizes a higher field strength at the receiver for a given amount of power at the transmitter. As television stations become more numerous, a different type of directivity may eventually be required. Specifically, it may some day be necessary to give protection to areas served by other transmitters operating on the same frequency in the manner which is being done by standard broadcast stations at present.

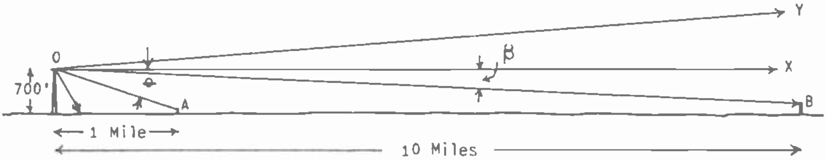


Fig.10 Vertical angle of radiation.

We shall first consider the problem of vertical directivity. In Fig. 10 is illustrated a transmitter whose antenna is located at the top of a tall tower; and two receivers, one at A, a distance of one mile, and one at B, a distance of ten miles. It is not possible to draw the vertical and horizontal dimensions of this figure to the same scale, as the building would be too small to be visible on the figure. Suppose that the tower is a tall one on a slight knoll so that the elevation of the antenna is approximately 700 feet above average level ground. It is obvious from this illustration that any radiation in the direction of X, parallel to the earth's surface serves no useful purpose, since it is not possible to elevate the receiving antenna to a height of 700 feet. Any transmission at elevations greater than X, say in the direction of Y would also have no useful purpose.

Even though radiation parallel to the earth's surface serves no useful purpose, oddly enough, television antennas must be designed to concentrate their energy in this direction. In Fig. 10, the receiving station at B located ten miles from the transmitter, receives its radiation from the transmitter at an angle β, which is three-quarters of a degree below the line parallel to the earth's surface. It is not practical to build an array which can discriminate between directions only three-quarters of a degree apart. Therefore, since maximum radiation is designed to follow the direction OX, maximum radiation will also occur along the line OB. A receiver located near the transmitter, such as A which is one mile distant, would receive radiation at an angle of 7½ degrees below the line OX parallel to the earth's surface. Even a directivity of 7½ degrees is not generally practicable in antennas designed for television. Stations closer to the transmitter than one mile would be in a region of such extremely high field strength that antenna directivity discriminating against reception in this region would not prevent these receivers from being served with a perfectly usable signal.

Consequently, television antennas should be designed to suppress as much as possible all high-angle radiation and to concentrate all of the energy in the form of a thin layer parallel to the surface of the earth. This desirable directivity is difficult to achieve in any type of high frequency service, but it is doubly so in the case of television, since the design of the directive properties must go hand in hand with a consideration of the bandwidth. Generally speaking, a complicated array is apt to have less bandwidth than a single dipole unless these two problems are coordinated in the design of the antenna system.

Horizontal directivity is less important than vertical directivity. In the usual case, the television transmitter is located in a tall building which, naturally, is in the center of the town, and therefore, often in the center of the populated area. In this case, no horizontal directivity would be desired and the antenna would be arranged to deliver a circular radiation pattern. If vertical polarization had been used, this would have been accomplished automatically. Since it is necessary to use horizontally polarized antennas, a circular pattern can be achieved by crossing two dipoles (half-wave antennas) and feeding them 90 degrees out of phase to produce a rotating field which is horizontally polarized. This arrangement is known as a single section *"turnstile"* antenna. The simplest way of feeding these antennas is to make the transmission line to one of them a quarter-wave longer than the line which feeds the other. If more than two antennas are used or if they are not crossed at their centers, it may be necessary to feed them with other phase relations. However, it is generally possible to produce a poly-phase rotating field with any number of symmetrically-disposed half-wave antennas.

The principle of producing a uniform field from a number of half-wave antennas is so important that it will now be discussed in detail. It is a familiar fact that a half-wave antenna produces a figure eight pattern in the plane of the antenna, as shown in Fig. 11A. At first, one would think that it would be possible to secure general coverage by placing two such antennas at right angles to each other and feeding them in parallel. Such is not the case.

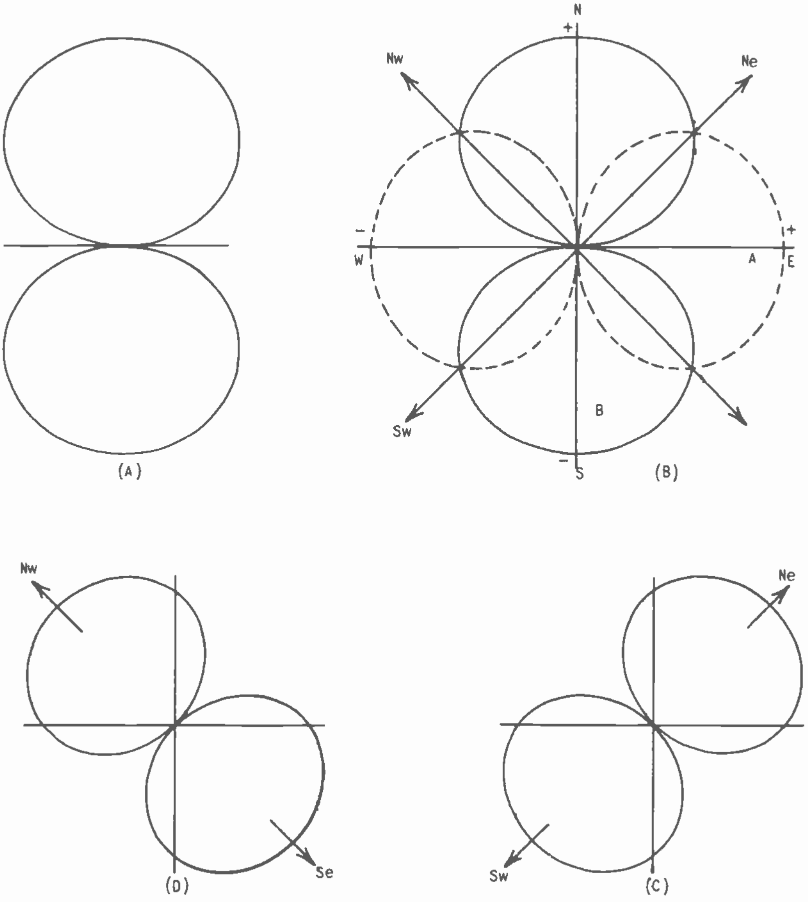


Fig.11 "Figure eight" pattern from two dipoles.

In Fig. 11B is shown the instantaneous fields from two antennas arranged in this manner. The polarities of the fields are indicated by the plus and minus marks upon the antenna wires. Of course, the radiation off the ends of either wire is zero; therefore, in the north, south, east, and west directions, one dipole or the other produces its maximum field and acts as if the other were not present. In the northeast and southwest directions, the fields from the two antennas add up, but in the northwest and southeast directions, the fields from the two antennas cancel. For example, assume any point in the southeast direction. It will lie equidistant from corresponding parts of dipole A and dipole B. Since at any instant the polarities of these two dipoles are opposite with respect to the assumed

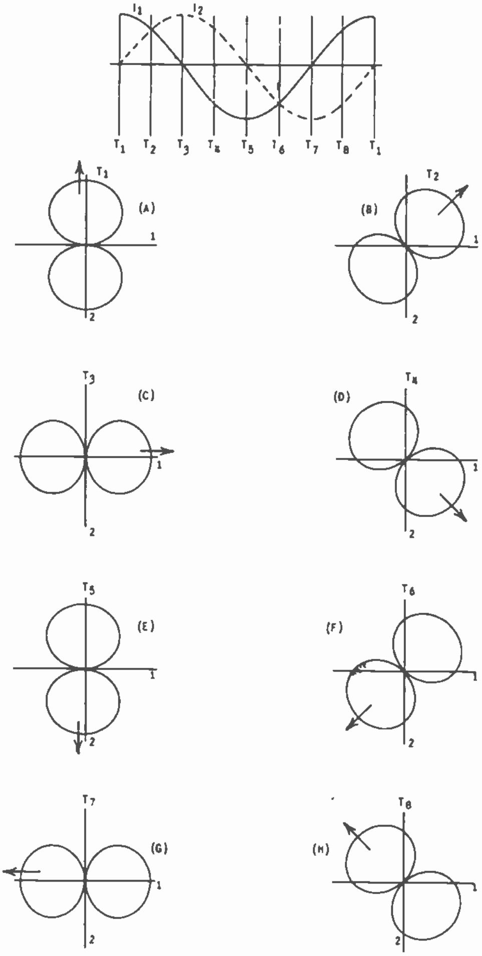
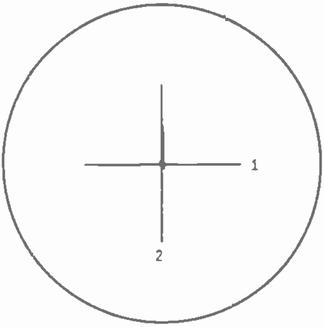


Fig.12 Turnstile connection produces a rotating field.

point, the field from one antenna will cancel the field from the other. The all-around result of these two super-imposed fields is shown in Fig. 11C, and it is seen that the field produced by two antennas is still a figure eight. If the leads to one of the antennas had been reversed so as to change the phase of one antenna by 180 degrees, the pattern of Fig. 11D would have resulted, which is apparently still a figure eight, but having the lobes in opposite directions.

When the crossed dipoles are fed with equal currents which are 90 degrees out of phase, as shown by the sine wave currents of Fig. 12, the figure eight pattern illustrated in Figs. 11A, 11C, and 11D will rotate at radio frequency speed so that all directions will be served with a uniform field strength. This can be understood by following through, in Fig. 12, the various instances throughout the radio frequency cycle and summing up the fields of the two antennas in all directions at each instant. At time T_1 the current in antenna 1 is positive and the current in antenna 2 is zero, resulting in a figure eight field in the direction indicated. A fraction of a cycle later (45 degrees later at time T_2) the currents are equal in the two antennas, but are somewhat smaller in magnitude than the current in antenna 1 at time T_1. The two fields add up to produce the figure eight pattern which is rotated in a clockwise direction from the field at time T_1. The production of this field is the same as the production of the field in Fig. 11. At time T_3, the current in antenna 1 has dropped to zero, but the current in antenna 2 has risen to a maximum, resulting in the figure eight in the direction as shown. At time T_4, the current in antenna 1 has changed direction, rotating the resultant field still further in a clockwise direction. Additional instances are shown until time T_8, when the rotating field has almost completed a cycle of 360 degrees.

Fig.13 Horizontal radiation pattern of turnstile antenna.



Actually, the field produced by the antenna is of constant magnitude at all instants of time, although its direction rotates at radio frequencies. However, to an observer in any direction, the effect is the same as if the antenna was producing an alternating field at radio frequency. Furthermore, the effect is the same for any direction about the antenna, or we can say that the antenna produces a circular pattern (general coverage) as shown in Fig. 13. One method of producing the 90-degree phase difference between the two antennas is shown in Fig. 14, and consists of feeding the two antennas through two terminated lines, one of which is electrically 90 degrees longer than the other.

At present, general coverage antennas are used almost exclusively. However, simple forms of horizontally directive antennas

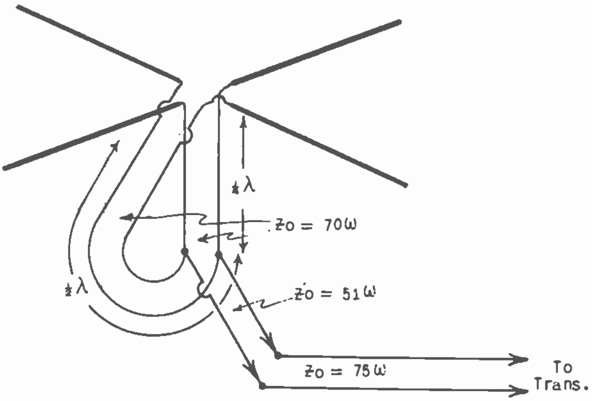


Fig.14 Turnstile connection of two dipoles.

are quite practical. For instance, if a transmitter is to be located between two towns which are close together, a simple horizontal antenna would give the figure eight pattern required to cover both towns. If the transmitter is to be located on the outskirts of a town or at the seashore where radiation in one particular direction would have no value, it can be suppressed by the use of a simple reflector of the parasitic type to give the cardiod radiation pattern shown in Fig. 15. In general, this type of array results in

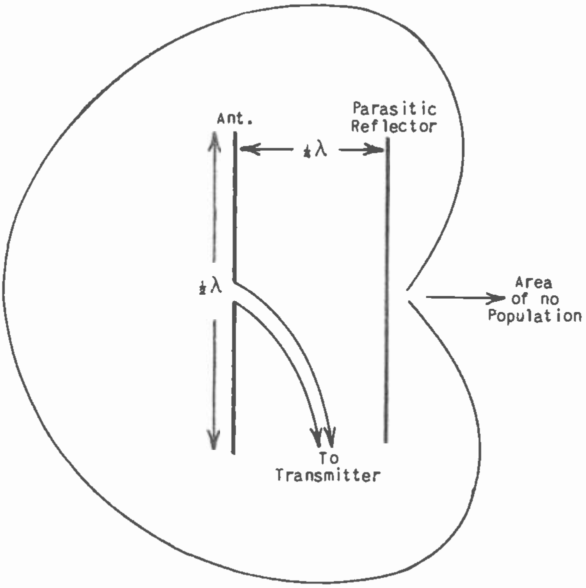


Fig.15 Horizontal directivity with simple parasitic reflector.

a restricted bandwidth, and for this reason is seldom used.

In discussing the problem of securing a circular radiation pattern for general coverage, or of securing a directional pattern for special cases, simple half-wave antennas were assumed; however, in practice this is not necessarily the case. As was pointed out at the beginning of this section, the narrowest possible directivity in the vertical plane is desirable; therefore, it should be realized that each of the crossed dipoles mentioned in the last paragraph may actually be a whole array of dipoles arranged to suppress vertical radiation. In this case, two similar arrays may be used, one mounted above the other, placed mutually at right angles, and arranged so that the arrays are fed with equal currents and a phase difference of 90 degrees. In simpler arrays, it is the usual practice to supplement *each element* of the array with a similar element placed at a 90-degree angle to the first and fed with a current displaced by 90 degrees as shown in Fig. 16.

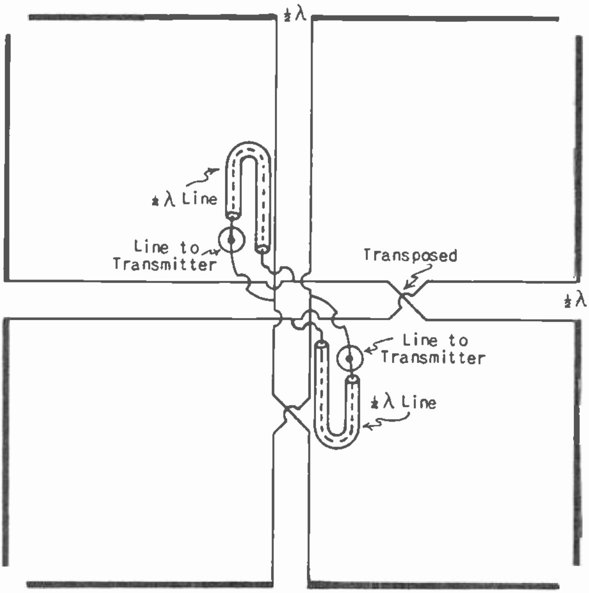


Fig.16 Method of connecting four dipoles to produce a circular field.

The specific nature of the array for suppressing vertical radiation will now be considered. Since television antennas must be kept fairly simple in order to facilitate the design of wideband characteristics, there are but two fundamental principles employed to secure vertical directivity. These may be used one at a time, or both together. One of these has already been shown in Fig. 16; the other general type is shown in Fig. 17, and consists of a two section "turnstile". These two systems are shown in simplified form in Fig. 18A and 18B respectively. When the two antennas are mounted in a plane parallel to the earth's surface, and fed 180 degrees out of phase, the radiation from the two antennas cancel in a vertical direction, since the earth immediately below the antenna (or the sky immediately above the antenna) is equidistant from the two elements which are carrying out of phase currents. Along the line parallel to the earth's surface, the radiation from the two antennas add up, because the field from antenna No. 1 requires 180 degrees longer to arrive at point X than does the field from antenna No. 2. Since the fields were originally started out of phase, they arrive at point X in phase. The result is the familiar figure eight pattern in the vertical plane.

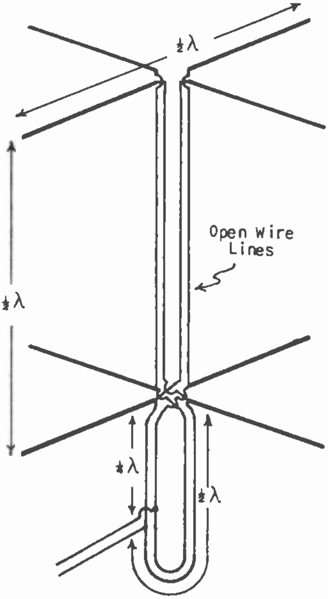


Fig.17 Two-section turnstile.

The same result occurs when two antennas are mounted one above the other and fed in phase. In this case, the fields from the two antennas add up in the direction X, but cancel in a vertical direction since the fields from the two antennas are out of phase at any point directly above or directly below the array, due to the half-wave spacing of the elements. The vertical patterns developed in Fig. 18 apply for the slightly more complicated structures of Figs. 16 and 17 which were arranged to give uniform coverage in all *horizontal* directions.

4. BANDWIDTH REQUIREMENTS. An ideal television transmitting antenna would radiate a constant field at any frequency throughout the television channel. It would also present a constant load impedance to the antenna feeder over the entire channel. The last of

these two characteristics is the most important; and if this condition is fulfilled, then the "flatness" of the antenna characteristic with respect to radiation of various frequencies will be satisfactory. Even if the radiation of all frequencies is not exactly the same, it can be compensated within limits by correcting circuits at video frequencies in the transmitter itself. Therefore, insufficient bandwidth of the radiation characteristic is not very important compared to the problem of maintaining a flat input impedance characteristic.

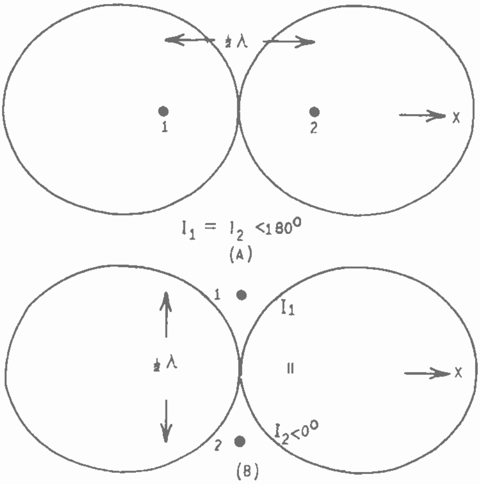


Fig.18 Basic methods of producing vertical directivity.

If the antenna is not properly matched to the transmission line, part of the energy traveling up the transmission line will be reflected from the antenna and will travel down the line in the opposite direction. Upon reaching the transmitter, it is reflected a second time and returned to the antenna. At this point, part of the energy will be radiated by the antenna and the remainder will be reflected again to the transmitter and back to the antenna. Thus, for each impulse generated at the transmitter, a series of impulses of decreasing magnitude will be radiated by the antenna. The effect at the receiver is to cause multiple images.

Let us take a typical case and examine the magnitude of this effect. Assume that a transmitter operating on a frequency of 51.25 megacycles feeds a television antenna through a transmission line which is 250 feet long. Assume that the antenna presents a load of 90 ohms to a 70 ohm transmission line. In this particular case, 78% of the energy would be radiated by the antenna and the remaining 22% would be returned to the transmitter. The wave returning to the transmitter would encounter almost perfect reflection at the transmitter since the tank circuits are sharply tuned and consist of pure reactive elements. The tube itself presents a negative resistance to the tank circuit. Upon the return to the antenna, 78% of the reflected wave would be radiated some time after the initial radiation had occurred. A signal strength comparison between the initial wave and the reflected wave, would be given as follows:

Generated Signal = 100%
1st Radiated Signal = 78%
2nd Radiated Signal = 22% of 78% = 17%

$$\text{Ratio: } \frac{\text{2nd Signal}}{\text{1st Signal}} = \frac{17}{78} = 22\%$$

Thus, it is seen that the antenna behaves as though two impulses instead of one were fed to it, the second impulse being 22% as strong as the first. Successive reflections would each be 22% of the preceding one, or in other words, the third reflection (fourth radiated signal) would be approximately 1% of the original signal, and thus indistinguishable at the receiver.

The distance between the transmitter and the receiver is greater for the case of the reflected wave than for the initial wave, because it must travel an additional path down the transmission line and back again, a distance of 500 feet. If we assume a velocity of propagation equal to .97 that of light, we find that the second impulse radiated by the antenna is delayed from the first by slightly over half a microsecond.

$$t = \frac{D}{V} = \frac{500' \times .3048\ (\text{meter/ft.})}{.97 \times 300{,}000{,}000} = .525 \times 10^{-6}\ \text{sec.}$$

On the face of a 12-inch cathode ray tube at the receiver, a series of images would be formed of decreasing intensity and spaced somewhat over a sixteenth of an inch apart. Assume a 7½" × 10" picture with 10% blanking time.

$$\text{Time to scan 10'' line} = \frac{.9}{13{,}230}\ \text{second} = 68\ \text{microseconds.}$$

$$\text{Displacement of multiple images} = 10'' \times \frac{.525}{68} = .077\ \text{inch.}$$

$$.077'' = \text{approx. } 5/64\ \text{inch.}$$

When the antenna does not have a constant input impedance over the television channel, it can generally be arranged to match the transmission line at the carrier frequency, but all of the sideband

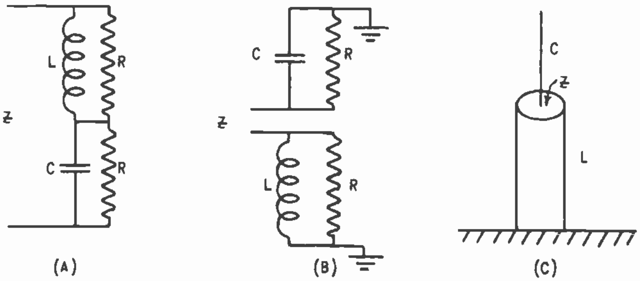


Fig.19 Simulating a constant-resistance network with a simple antenna.

frequencies will be mis-matched in varying degrees. When this condition exists, a multiple image is still formed at the receiver, but since the sideband frequencies suffer more reflection than the carrier frequency, the multiple images would have an absence of low video frequency components. Thus, the multiple images would lack contrast and have, in general, a very ghostly appearance compared to the main picture signal. Television antennas should have an input impedance which should vary preferably not more than 5%, and at most not more than 10%, plus or minus, over the television channel. Several methods of obtaining this desirably flat characteristic will now be considered.

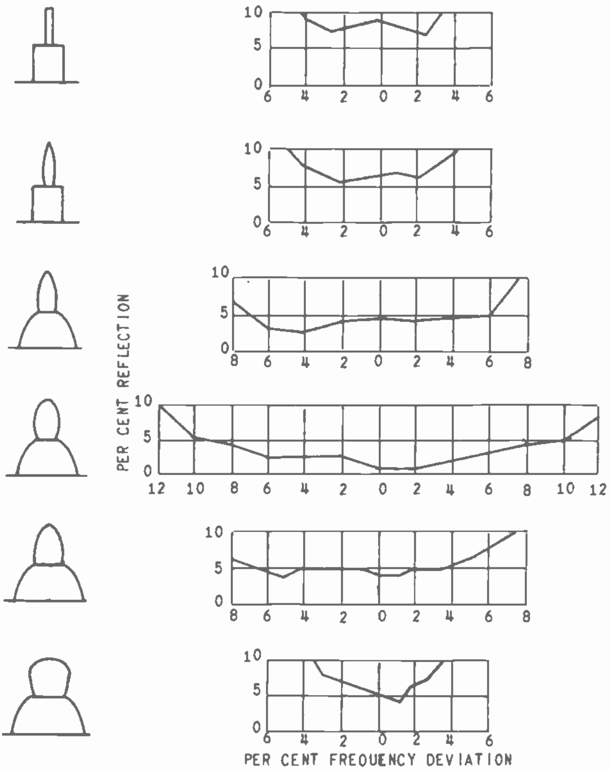


Fig. 20 Effect of radiator shape upon bandwidth.

The student has already learned that a network of the form shown in Fig. 19A exhibits a constant and pure resistance at all frequencies. If the center of the network is grounded, it can be arranged as shown in Fig. 19B. At one particular frequency, this network could be simulated by a short antenna fed with an extended transmission line as shown in Fig. 19C. An antenna shorter than a quarter-wavelength behaves as a capacity which is loaded with radiation resistance. Thus the antenna c of Fig. 19C simulates the resistance-loaded capacity of Fig. 19B. Since the outside portion of the transmission line is not grounded at its end, current must flow along the outside of the outer conductor back to ground. The line is chosen so that the path to ground is less than a quarter-wavelength, and in this case the outer surface of the transmission line presents an inductive reactance to the termination Z. The current flowing on the outside of this transmission line causes some radiation, and therefore, the inductive transmission line is loaded with radiation resistance, fulfilling the conditions set up under Fig. 19B. If the antenna and transmission line of Fig. 19C behave as pure capacity and inductance, the system will reflect a constant resistance to the terminus of the transmission line. However, both of these elements have distributed inductance and capacity; and therefore, the effective capacity (in the case of the antenna) and the effective inductance (in the case of the short transmission line) are not constant over a wide band of frequencies, due to the fact that at different frequencies, these elements are different fractions of a quarter-wavelength. There is, however, a way of making these elements approximate pure capacitance and pure inductance.

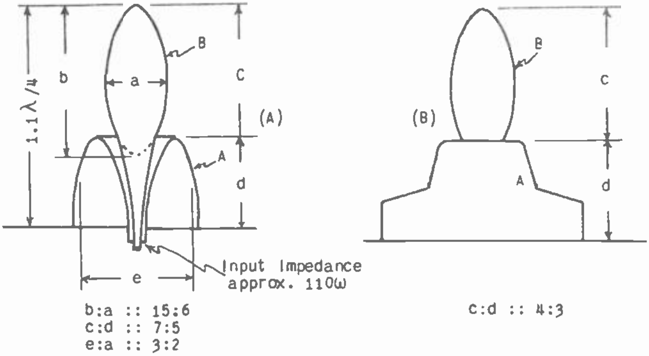


Fig.21 Optimum mechanical proportions for a wideband radiator.

The ratio of distributed capacity to distributed inductance along an antenna, or along a section of transmission line, varies with the diameter of the line. Thus, by altering the proportions of the line, it is possible to offset the change in electrical length over the television band by an equal and opposite effect of the varying ratio of capacity to inductance which the current distribution encounters. The effect of varying the shape of the antenna elements is shown in Fig. 20, and Fig. 21 shows the optimum cross-section of a quarter-wave element fed by a 110-ohm concentric line. A view of the RCA installation at the Empire State Building employing the principles just outlined is shown in Fig. 22, while the wide-band characteristic of the antenna is shown in Fig. 23. You will notice that the bandwidth of this antenna exceeds by a considerable margin the best characteristic obtained under Fig. 20. The reason for this is that the two pairs of antennas are fed 90 degrees out of phase in a turnstile fashion. The effect of the turnstile feeder connection upon the bandwidth will be discussed in following paragraphs.

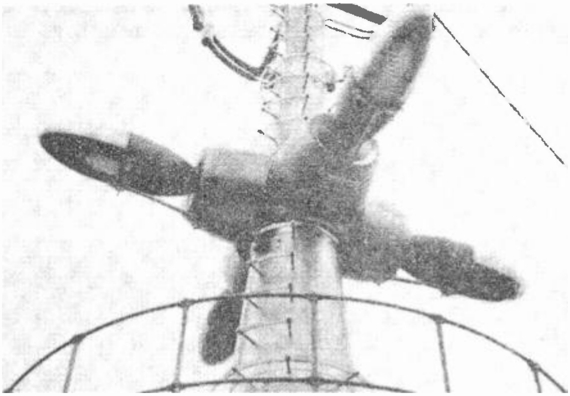


Fig.22 Empire State-NBC wideband turnstile.

The RCA installation in the Empire State Building is a splendid piece of engineering. However, the antenna itself is physically complicated and difficult to fabricate. Furthermore, if many of these complicated dipoles (pairs of "Indian Club" quarter-wave antennas) were used in an antenna array, the array would be expensive to build and would have considerable bulk and wind resistance, making it difficult to support atop a tall tower. For this reason, many modern installations still use systems of ordinary dipoles arranged to achieve broad band characteristics by one or more of the principles which will now be outlined.

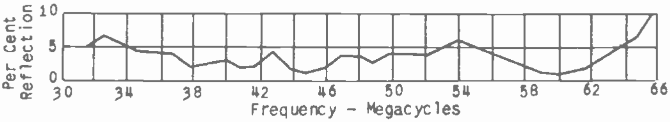


Fig.23 Bandwidth of Empire State antenna system.

One of the simplest methods of increasing the bandwidth of the antenna is to increase the diameters of the conductors used in the construction of the antenna. The larger the diameter of the antenna conductor, the smaller is the reactance or impedance change over the television band. This method is limited in its ability to flatten the impedance of the antenna system and a point is soon reached where an increase in conductor size is not justified by the small increase in bandwidth that is obtained. As the conductor diameters become appreciable fractions of a wavelength, it becomes difficult to arrange a connection from the antenna to the feeder.

To preserve uniform current distribution around the antenna conductor when it is of large diameter, the end of the conductor should be tapered to a point and a feeder attached to this point. Large diameter conductors have considerable weight and wind resistance so that the mechanical support is difficult to arrange. However, it is possible to use a cage system of wires to simulate, to a close degree, the performance of a large diameter conductor.

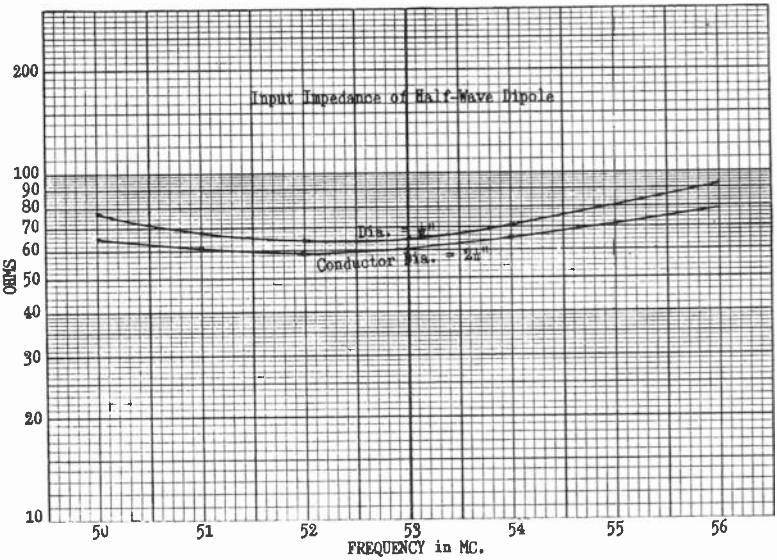


Fig.24 Effect of conductor diameter upon bandwidth.

A comparison of bandwidths for antennas of ½" diameter and 2¼" diameter, each operating at a frequency of 51.25 megacycles, is shown in Fig. 24. Conductors having a diameter of 2¼" are easy to support and even slightly larger ones are quite practical. For example, the General Electric antenna, located in the Helderberg mountains outside of Albany, New York, consists of eight dipoles, each about seven feet long and about 4" in diameter. The elements of this antenna will have a "flatter" input impedance than the best curve of Fig. 24. A picture of this installation is shown in Fig. 25.

Another way to increase the bandwidth of an antenna system is to apply correction circuits across the transmission line close to the antenna, but between the antenna and the transmitter. Fundamentally, an antenna is nothing more than an extended transmission line which is loaded by radiation resistance. Over one television band this is approximated to a fairly close degree by a series resonant circuit of low Q as shown in Fig. 26. At resonance, such a circuit is a pure resistance, but on either side of resonance, the circuit also presents a reactive component which is capacitive on

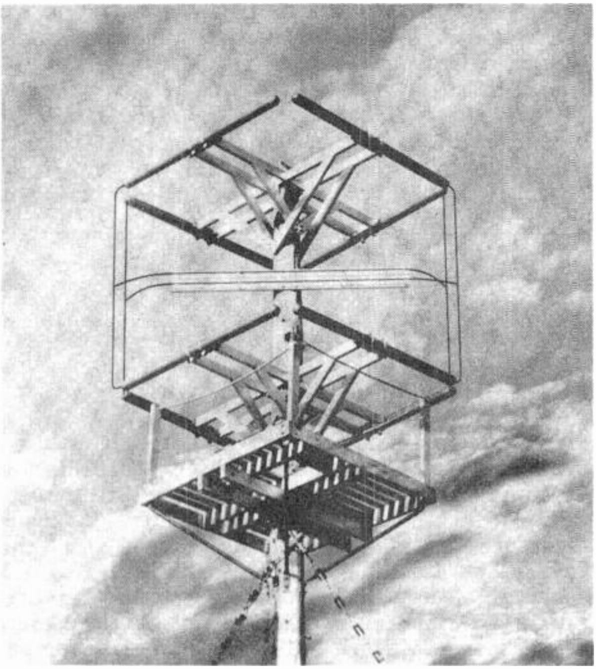
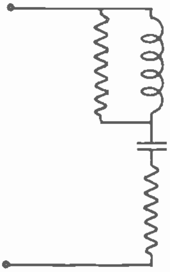


Fig.25 General Electric wideband antenna at Albany N.Y. (Photo, courtesy General Electric Co.)

the low frequency side, and inductive on the high frequency side. A parallel resonant circuit exhibits an opposite reactance. On the high frequency side of resonance, a parallel circuit is capacitive and on the low frequency side of resonance, it is inductive. Thus,

Fig.26 Equivalent circuit of a simple antenna.



by combining the two as shown in Fig. 27, it is possible to make the reactance change of the parallel circuit partially neutralize the reactance change of the series circuit. At resonance, the parallel circuit has a pure and very high resistance and thus does not influence the performance of the series resonant circuit.

The application of this principle is to place a high Q, parallel resonant circuit of low characteristic impedance across the terminals of each dipole in the antenna system. Such a tank circuit could be formed by the use of the Kolster "hat" studied in Lesson 4 of this unit. Although such a circuit would cancel the reactance variation of the antenna, the resultant resistance variation over the television band would still be very high; in fact, worse than before the tank circuit was added. However, by some refinements of this general principle, it is possible to secure an antenna system with very small impedance variations over the television band.

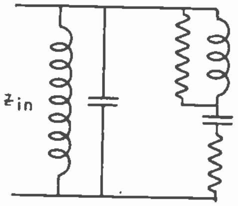


Fig.27 Compensating network plus simple antenna.

The curves of resistance and reactance, from which the impedance curves of Fig. 24 were plotted, are shown in Fig. 28. Resonance of the antenna occurs when the reactance is zero. Thus, it is seen that the larger the diameter of the conductor, the shorter it will be to produce resonance. Given a diameter of conductor and frequency of operation, it is possible to predict the resonant length of the antenna by applying Fig. 28, when the effect of surrounding objects is neglected. Cancelling the reactance of the antenna system by applying an additional parallel tank as discussed in the preceding paragraph causes the effective resistance of the antenna to rise. However, the radiation resistance of the antenna falls off somewhat below resonance. Consequently, on the low frequency side of resonance, the resistance rise due to the effect of the parallel tank circuit is offset by the decrease in radiation resistance. Provided the proper diameter of antenna elements is chosen, the result will be a nearly constant and pure resistance over one television channel.

In practice, the corrective circuits are not placed directly at the dipole, but are placed across the transmission line near the antenna. If the tank circuit is placed across the line, a half-wave behind the antenna, it will behave almost as if it were placed directly across the antenna.

The student has learned in the preceding lesson that a half-wave section of line reflects an impedance to its input terminals equal to the impedance placed across the output terminals. In the case of a wideband antenna system, the reflected impedance at the end of a half-wave section of line does not follow this simple rule inasmuch as a line is not an exact half-wave length at all frequencies over the television channel. However, the practical difference is not a disadvantage. The tank circuit placed across the line need not be limited to a simple form of tank circuit and is, in general, formed by the use of two sections of transmission line, one of which

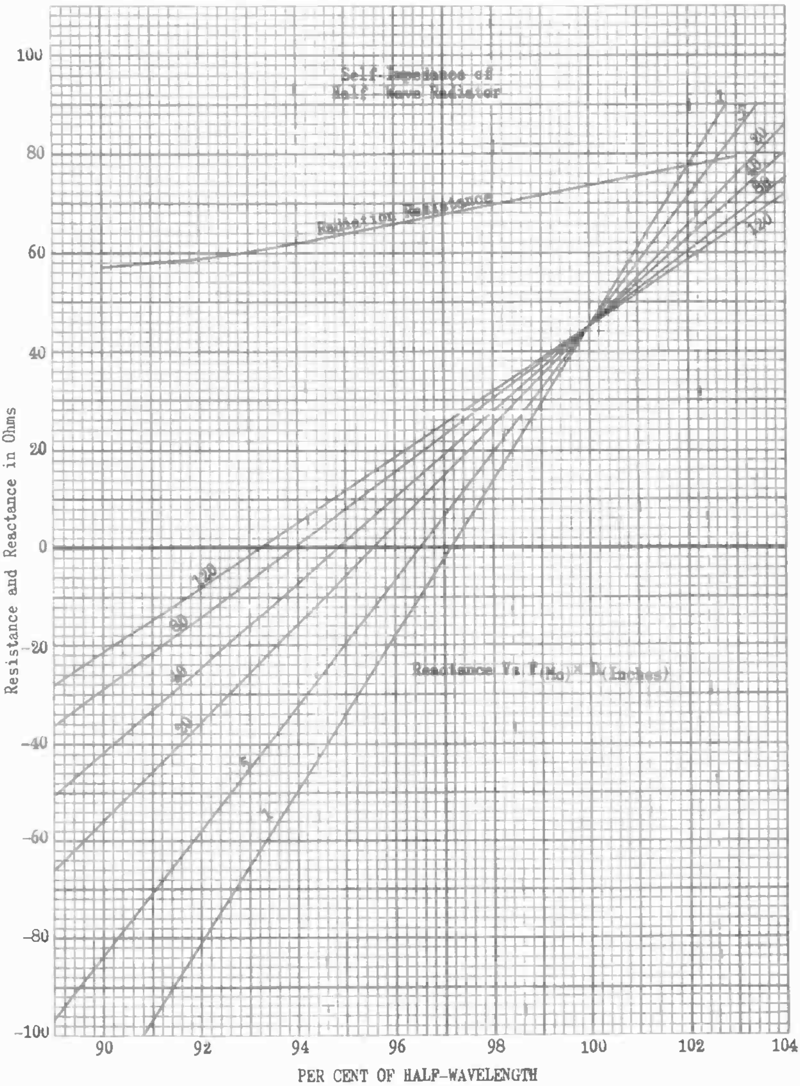


Fig.28 Self-impedance of half-wave radiator.

is tuned to a frequency above the television channel, and the other tuned to a frequency below the television channel. Thus, one of the lines will be inductive and the other capacitive, and the two together will simulate a peculiar type of parallel tuned circuit. The advantage of such a complicated circuit is the extreme flexibility of reactance characteristics that can be achieved.

In the case of a simple tuned circuit, the general shape of the characteristic follows a universal resonant curve as shown in Fig. 29A. In the case of the network just described, the circuit not only has a parallel resonant frequency, but two series resonant frequencies; one either side of the parallel resonant point. The difference between the two series resonant points can be altered by varying the lengths of the transmission lines, while the ratio of the slope on the low frequency side and on the high frequency side of resonance can be varied by altering the characteristic impedances of the lines. A general idea of the type of characteristics that can be formed in this manner is shown in Fig. 29B. This flexibility aids in the selection of a network which will best flatten the input impedance of the antenna system when placed across the transmission line. Not only can the choice of network be varied, but the position of its placement along the transmission line can be chosen for optimum flatness of input impedance. An application is shown in Fig. 30.

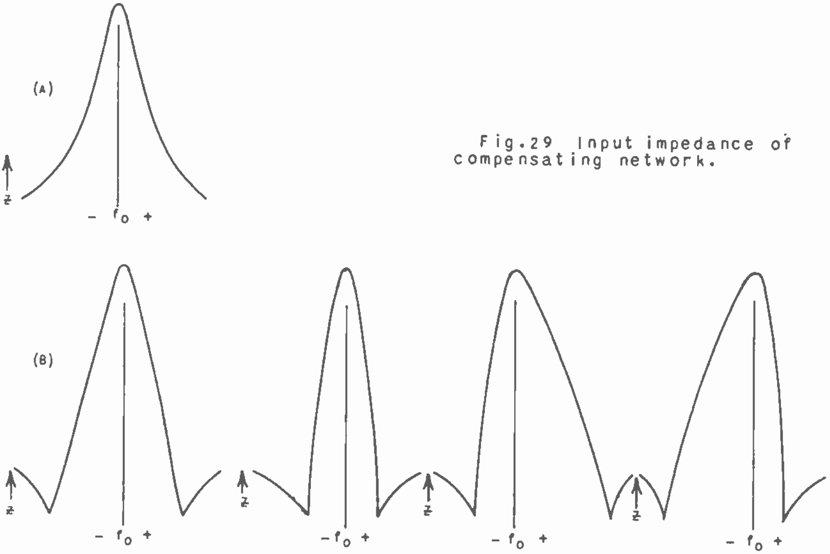


Fig.29 Input impedance of compensating network.

The television antenna of the Columbia Broadcasting Station atop the Chrysler Building employs this general principle as diagrammed in Fig. 31. The use of correction circuits can be thought of either as improving the input impedance characteristic of the antenna system, or as causing a reflection on the transmission line, which is 180 degrees out of phase to the reflection caused by the antenna itself. Whichever system of logic is employed, the action is the same in reducing the detrimental effect of mis-matching the feeder at the antenna.

A fourth way of reducing the impedance variations of an antenna system involves making use of the impedance inverting characteristics

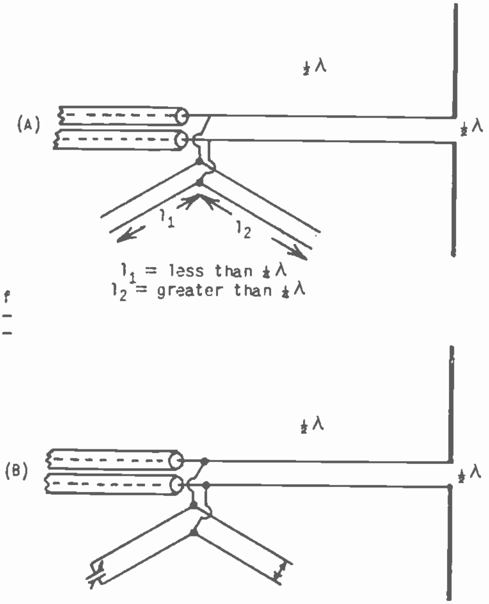


Fig.30 The application of transmission lines to wide-band antenna correction networks.

of a quarter-wave section of transmission line. If an antenna having the characteristics of Fig. 32A is inverted through a quarter-wave section of transmission line, which matches the antenna impedance at the carrier frequency, the input impedance to the quarter-wave transmission line is given by the curve of 32B. Now, if an additional antenna, similar to the first, is connected across the feeder to the first antenna, one-quarter wavelength away from the first, the combination will achieve a flat impedance characteristic over the entire band. Thus, the combination shown in Fig. 33 will exhibit a flat input impedance characteristic at point X.

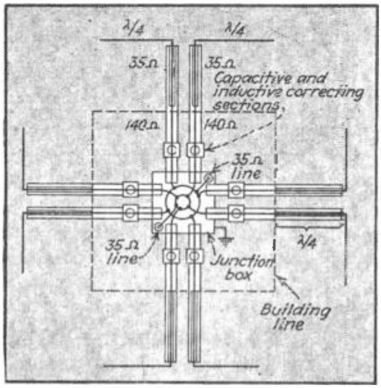


Fig.31 Columbia's Chrysler Building television antenna system.

Antenna A has its impedance inverted at point X as explained by Fig. 32. Not only is the impedance inverted by the following familiar expression:

$$Z_{in} = \frac{Z_0^2}{Z_{out}}$$

but the sign of the reactance is also inverted. Thus, an antenna at A (Fig. 33) which exhibits a capacitive reactance on the low frequency side of resonance will appear to point X as an inductive reactance, due to the inverting characteristics of the transformer T. This inductive reactance cancels the capacitive reactance of antenna C. The compensating effect of adding the reciprocal impedances of the two antennas at point X is very high provided that the mis-match of antenna A or antenna C is not excessive. Thus, if a 10% reflection is caused by antenna A and a 10% reflection is caused by antenna C, the resultant input impedance will be so flat as to cause a reflection of only 1%. However, if the antennas themselves would cause reflections on the order of 50%, the compensation effect of this type of connection is only 2:1.

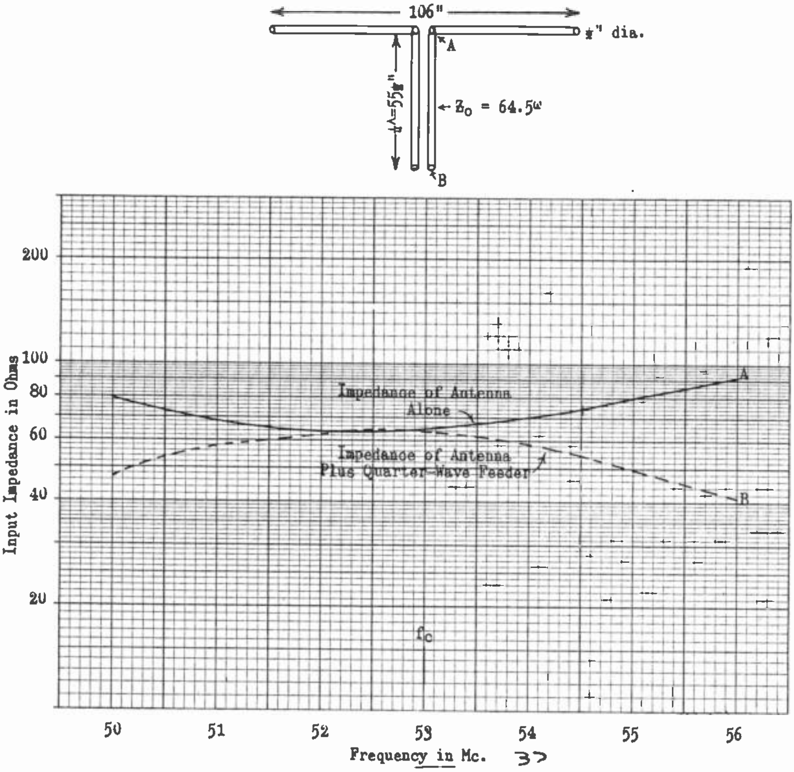


Fig.32 Impedance inverting properties of antenna matching quarter-wave line.

Fortunately, the use of the quarter-wave connecting transformer is highly desirable from another standpoint. We have already learned that a quarter-wave transformer must be inserted between two antennas to cause the 90 degree phase shift necessary to secure a circular radiation field from two antennas. Thus, the antennas of Fig. 33 need only be rotated at right angles to each other to function in the same manner as the antenna shown in Fig. 14. Either Fig. 14 or Fig. 33 (altered as mentioned) is known as a single section *turnstile*. It is the turnstile connection principle which accounts in a large measure for the extremely flat impedance characteristic of the NBC antenna installation whose performance was shown in Fig. 23.

In considering the directive properties of antennas, it was mentioned that the design of the directivity of the system must be carried hand in hand with considerations of bandwidth. The reason for this is that when two or more antennas are operated so that the space between them is comparable with the operating wavelength, considerable mutual coupling exists between the antennas. Thus, each antenna reflects an impedance, which has both a resistive and a reactive component, into the other antennas. The behavior of antennas is somewhat similar to the behavior of coupled coils (such as the primary and secondary of an I.F. transformer in the receiver) in that changes of loading and tuning in one winding are reflected as an impedance change in the other.

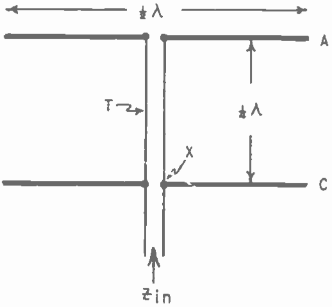


Fig.33 Combination of two antennas and quarter-wave line having wideband characteristics.

By varying the spacing between elements in an array, it is possible to make the mutual resistance and mutual reactance oppose the self-resistance and self-reactance of the elements and thus flatten the input impedance of the array as a whole. For instance, in the case of two out-of-phase elements which are located close together, the change of mutual impedance is such as to sharpen the input impedance of either, or both together, and decrease the bandwidth. However, if the same elements are placed slightly over a half-wavelength apart, the change of mutual impedance tends to nullify the change of self-impedance and thus improve the bandwidth. The improvement that can be achieved by this method is not very great, but some improvement is possible. On the other hand, if the array is designed without regard to the effect of mutual impedance between the antenna elements, the effect is generally to decrease the bandwidth.

5. SOUND ANTENNAS. Since the sound channel is very narrow, compared to the bandwidth of the television channel, mis-terminations of the transmission line are unimportant, except insofar as they increase the losses in the transmission line and decrease the power handling capability of the line. Consequently, if the television antenna is designed to have sufficient bandwidth to cover the picture channel, the same antenna will be a satisfactory radiator for the sound transmitter.

The two transmitters cannot be simply hooked up to the same antenna, since the power that would circulate between the two transmitters would cause cross-modulation to occur between the sound and the picture signals in the output stages of the transmitter. If the same antenna is to be used for both transmitters, it becomes necessary to employ a special coupling or combining network which will be discussed later in this lesson.

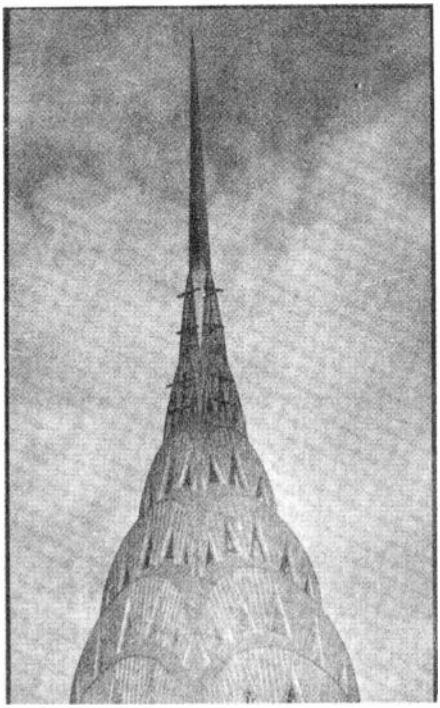


Fig.34 Television antenna installation of CBS.

The use of a coupling network can be avoided by employing two separate antenna systems, one for the sound transmitter and one for the picture transmitter. Even then the antennas must be located quite far apart, or be designed in such a way that a minimum mutual impedance exists between the two antennas. If this is not done, the mutual coupling between the two antennas causes the power radiated from one antenna system to be picked up by the other and propagated down the line to the output stage of the transmitter. If such is the case, cross-modulation will occur and the received picture will suffer from horizontal striations whose intensity varies with the character and intensity of the sound modulation. On the other hand, the sound quality at the receiver will suffer, due to a super-imposed, 60-cycle note, with many harmonics, which has been introduced into the sound by cross-modulation of the vertical blanking and synchronizing pulses of the picture transmitter.

If the reduction in mutual impedance between the two antenna systems is to be achieved simply by locating the antennas at a distance to each other, a great deal of space will be required, since considerable mutual impedance exists even with a spacing of one wavelength between antennas. Since at present most television antennas are located at the top of high buildings, it is difficult to mount one antenna system above the other and maintain a great deal of spacing between the sets of antennas. However, by employing antennas which have directivity in the vertical plane, and which radiate a minimum amount of energy straight up and straight down, it is possible to reduce considerably the mutual coupling between the sound and the picture antennas.

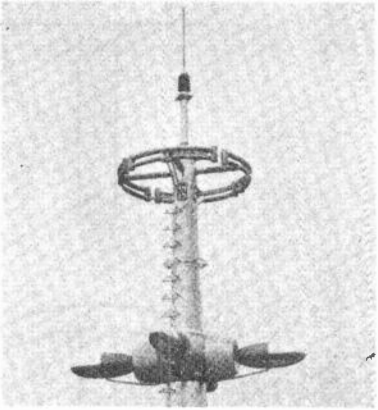


Fig.35 NBC television sound antenna.

The sound and picture antennas of the Columbia station in the Chrysler Building are pictured in Fig. 34 and employs the principle of Fig. 18B to eliminate radiation in the vertical direction. In addition to this, the antennas themselves are a considerable distance apart as can be seen in Fig. 34. This figure is not an actual photograph, since at the time of this writing, the installation of the antenna is not yet complete.

The top set of radiators which appear in the figure constitute the sound array, while the lower set of radiators constitute the picture array. Both the sound and the picture antenna consist of two tiers of dipoles, a half-wave apart vertically. Since the complementary dipoles are fed in phase, and since they are a half-wave apart in the vertical direction, the field from one dipole of the sound antenna arrives at the television antenna 180 degrees out of phase with the field from the complementary sound dipole. Therefore, any mutual coupling that exists between one of the sound dipoles and the television antenna is cancelled out by an equal and opposite mutual impedance existing between one of the other dipoles of the sound antenna and the picture antenna. The same situation exists with regard to any energy which might be transferred between the picture antenna and the sound antenna.

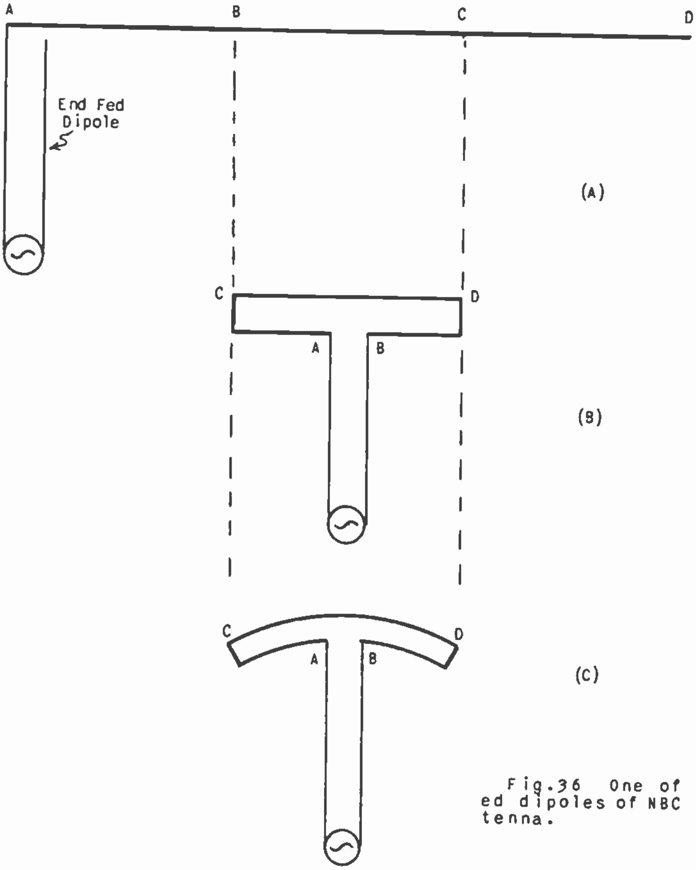


Fig.36 One of the folded dipoles of NBC sound antenna.

Mutual coupling between the sound antenna and the picture antenna can also be reduced in the case of turnstile antennas by connecting them so that the fields from the two antennas rotate in opposite directions. It was pointed out earlier in the lesson that to secure a uniform field with horizontal dipoles, a rotating field could be produced by connecting the crossed dipoles with a quarter-wave section of transmission line in order to delay the field produced by one dipole. If the connections to one dipole are reversed, the dipole which produced the leading field in the first case, produces the lagging field in the second case. The rotation of the resultant field will occur in the opposite direction. The two fields rotate at slightly different speeds, due to the difference in carrier frequency between the picture transmitter and the sound transmitter. For this reason, some coupling still exists between the two antenna systems.

Three other principles are available for reducing the coupling between the sound antenna and picture antenna. Since all three of these are incorporated in the NBC Empire State installation, they will be discussed in conjunction with this particular installation.

The sound antenna is shown in circular form mounted above the visual antenna in Fig. 35. If the sound antenna was a perfect circle, carrying uniform current, it would be at right angles to the turnstile antenna at all points, since the turnstile is lined up in such a way as to form the diameters of the circular sound antenna. Since the current in the sound antenna is at right angles to the current in the picture antenna, no coupling would exist between the two. In practice this is approximated by using four folded dipoles, arranged in a circle, for the sound antenna. The method of folding the dipole and attaching the feeder is shown progressively in Figs. 36A, B, and C.

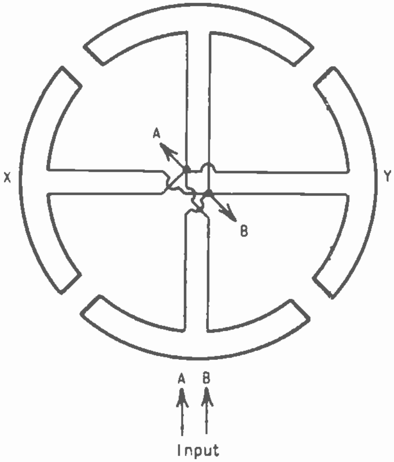


Fig.37 Connections of NBC circular sound antenna.

When the dipole is folded in this manner, the radiation resistance is lowered, due to the fact that radiations from part of the antenna are cancelled by radiations from other parts of the antenna. The lowered radiation resistance means a higher Q and, therefore, a more sharply tuned antenna. In fact, the antenna takes on the nature of a slightly loaded section of transmission line. The increased sharpness causes the antenna to be non-resonant over the television picture band and for this reason, as well as for the previous one, very little coupling exists between the sound antenna and the picture antenna.

The four dipoles are arranged in circular form and connected to a single feeder as shown in Fig. 37. Opposite dipoles are connected 180 degrees out of phase and since they are a half-wave apart in a horizontal direction, the radiation along any horizontal line adds up for the opposite dipoles. This connection, however, cancels out the effect of any voltages produced between the picture antenna and the sound antenna. The picture antenna induces equal voltages in opposite dipoles, such as X and Y in Fig. 37, but the induced voltages cancel at the connections of the feeder, A and B, since the branch feeders to the individual dipoles are reversed at this point. Thus on three points, namely, physical placement, sharpness of resonance, and reversing of feeder connections, the circular sound antenna reduces to a minimum the coupling to the picture turnstile antenna.

6. TRANSMISSION LINE REQUIREMENTS. The necessity for precise termination of the transmission line has already been considered. This section will be concerned chiefly with the mechanical nature of the line and the effect of the construction and application of a line upon the performance characteristics of the antenna system as a whole.

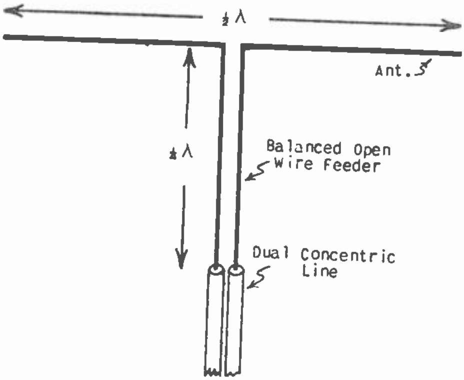


Fig.38 One use of open wire transmission line.

Transmission lines used for television purposes are almost exclusively of the concentric type. Balanced open wire lines of the type shown in Fig. 38, are sometimes used, in short sections of approximately one-quarter wavelength, as matching transformers between the main feeder and the antenna. They are occasionally used for longer distances because of the ease of installation when the transmission line is to be carried around corners, or where short lengths are used for interconnecting the dipoles of a complicated array. The open wire type is rarely used to form the complete transmission line, because it is necessary to prevent stray radiation from the line, and this is difficult to do with the open wire type. Since the line has to be carried up through buildings, or up the center of a tower, irregularities in the construction of the building or tower will tend to unbalance the line to ground, and will therefore cause stray radiation. Radiation from the transmission line affects the quality of the picture in much the same way as mis-terminating the line, since the path of the stray radiation to the receiver is different than the path from the transmitting antenna.

The unavoidable irregularity in the proximity of the tower or building acts to cause slight variations in the characteristic impedance of the line. This tends to cause reflections at various points along the transmission line. The detrimental effects of reflections have already been considered.

Another disadvantage of open wire lines is the difficulty of achieving a low value of characteristic impedance. You will be shown later that it is desirable to have the transmission line match the antenna load without the use of impedance-matching transformers. Because of the present methods of feeding the antennas for television, the impedance of the antenna system is generally lower than the value which is easily obtainable by the use of open wire lines.

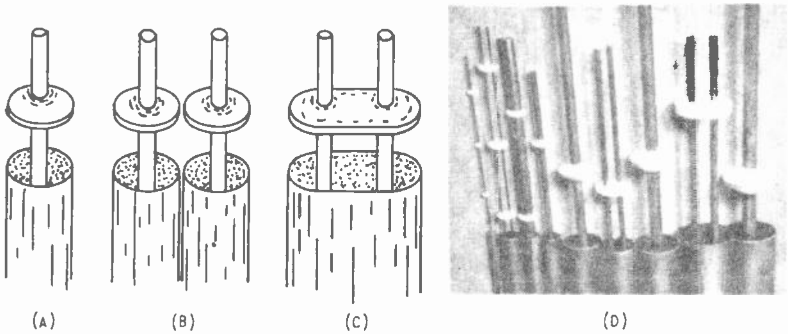


Fig.39 Forms of concentric transmission lines.

Concentric transmission lines are used in three general forms shown in Figs. 39A, B, and C. Fig. 39A pictures the unbalanced or ordinary coaxial transmission line. Fig. 39B shows how two of these lines may be used to feed a balanced load, while Fig. 39C shows a balanced type of concentric line.

The simple concentric line is the least expensive of the three types. Since the antenna load is always balanced to ground, it is difficult to arrange a feed system employing a single-ended transmission line. Another disadvantage of this type of line is that it is difficult to secure an absolute ground at both ends. Unless this is done, radio frequency currents will circulate on the outer shell and produce radiations with detrimental effects upon the picture quality.

There are two general methods for overcoming this latter disadvantage, which make it possible to feed balanced loads from this type of transmission line. Either of these two methods automatically aids in effecting a good ground for both ends of the transmission line.

The first of these two methods is the use of a half-wave section of line as a phase inverter, shown in Fig. 40. In this case, the line PO is made a half-wave long. Thus, the voltage delivered to point O is 180 degrees out of phase with the voltage appearing at point P. Furthermore, since a half-wave section of line reflects an impedance equal to its termination, the two halves of the antenna are effectively in parallel at point P, and equal power is delivered to both halves of the antenna. In other words, the voltage from P to ground is the same as the voltage from O to ground. This is the system employed in the British Broadcasting Company's antenna system which uses a large single feeder to the television antenna.

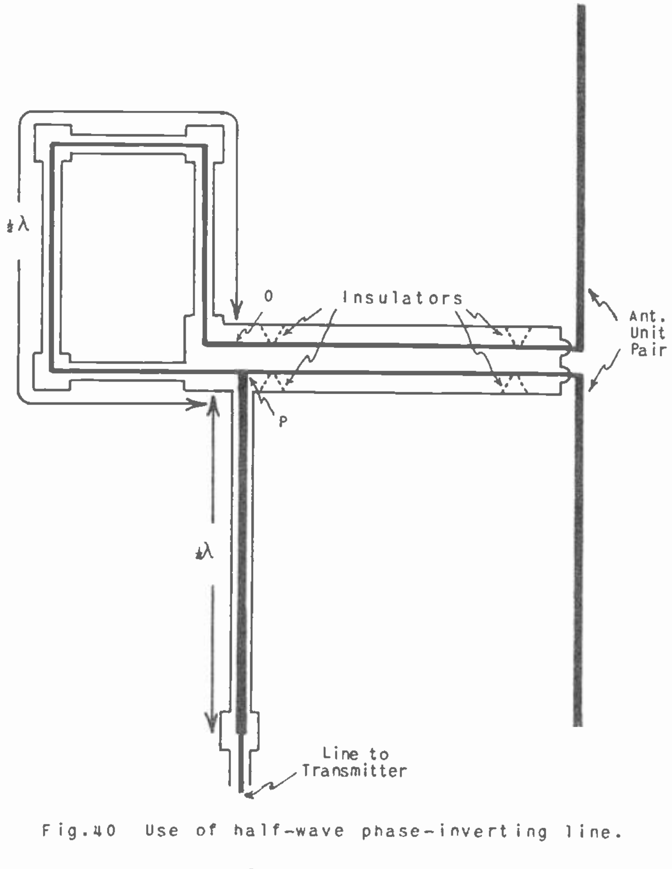


Fig.40 Use of half-wave phase-inverting line.

The second method employed in coupling between a balanced load and a single transmission line is the use of a *balance converter*, as shown in Fig. 41. In this method, the last quarter-wave of the transmission line is surrounded by a concentric sleeve which is connected to the transmission line at a point one-quarter wavelength from the end of the line, shown at point B. Thus, the outer conductor of the main transmission line is, in turn, the inner conductor of the balance converter. The balance converter is essentially a quarter-wave concentric line, short-circuited at one end (point B). The student already knows that a quarter-wave line shorted at one end exhibits an extremely high impedance at the other. Therefore, a high impedance exists at point Z between the outer conductor of the main transmission line and its surrounding sleeve. The sleeve is essentially at ground potential.

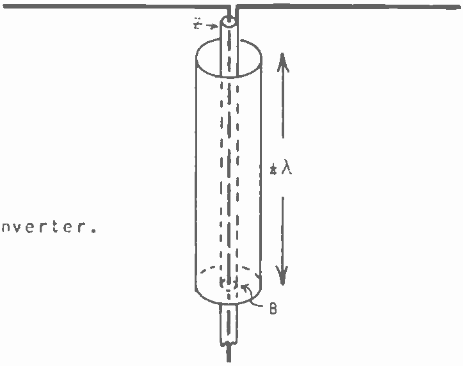


Fig. 41 Balance converter.

Since the outer conductor of the main feeder exhibits a high impedance to ground, it may be connected to one-half of the dipole in the same manner as the central conductor is connected to the other half of the dipole. The mutual coupling which exists between the two halves of the dipole, or antenna system, balances the voltages to ground. An equal voltage to ground appears on the inner and outer conductor of the main transmission line. On the other hand, the main transmission line has its outer shell effectively grounded at point B, because of the action of the balance converter, and from this point back to the transmitter, the transmission line behaves in a normal single-ended fashion.

This system of feeding a balanced load from a single concentric line is used in Midland's high frequency broadcast transmitter W9XER and also in the sound antenna installation for the NBC television transmitter atop the Empire State Building.

The picture channel of this transmitter utilizes the double barrel transmission line previously shown in Fig. 39B. This type of transmission line is also used by the Columbia Broadcasting System installation, and in fact by almost all of the present television installations. The balance achieved by the use of two lines does not possess the disadvantages of the open balanced type of line. These are overcome because of the shielding afforded by the outer conductor of the concentric type. In this case, the outer shells are bonded together at frequent intervals, and particularly, at both ends of the transmission line. They may either be bonded to the tower, or building, or entirely insulated from it; but in either case, they must be mechanically rigid. There is always some slight unbalance, causing some current to flow on the outer shell. If the shell currents are changed in any way, as would be the case should the bonds make intermittent contact, or with the lines not rigidly supported so that they would vibrate close to each other or to the tower in a high wind, the slight radiation from the lines would alter in character, causing changes in the received picture quality. Furthermore, a variation in input impedance of the transmission line would occur, and this change, although exceedingly small, would further affect the picture quality by causing the antenna to reflect a varying load to the transmitter. Of course, these effects do not actually occur in large transmitter installations, because the lines are large and rigid, and care is taken to secure stability in their installation.

The balanced and shielded type of line shown in Fig. 39C, is at present, rather special, and used only in smaller apparatus such as remote pick-ups or mobile transmitters. It has several advantages over any of the other types already considered, in that it is inherently balanced, and any unbalance occurring in the two lines has a smaller effect upon the transmitter and upon the transmitted picture quality. It has one slight disadvantage in that it is impractical to make the inner conductors special in shape, and when round conductors are used, the distribution of current on these conductors is not entirely uniform, and therefore, the efficiency from this type of line is slightly lower than the type shown in Fig. 38B. This however, should not be a very practical disadvantage, and it is to be anticipated that this type of line will become more prevalent in the future, particularly in the smaller stations of approximately 1 kilowatt power. These lines, as yet, are not commercially available as a standard product, but can be made to order at a very nominal charge.

The size of the transmission line is determined by the breakdown voltage of the line, the line loss, the power capability of the line, and the ease of mechanical construction. The breakdown voltage of a transmission line is rather erratic at the ultra-high frequencies involved, and is determined by the spacing of the line, and by the radius of curvature of the conductors. Thus, sharp points, caused by crimping the inner conductor to hold the isolantite spacers in place, reduce the radius of curvature, and materially lower the breakdown potential of the line. Metal filings, or dirt, work in a similar manner, and decrease the breakdown voltage tremendously.

The line loss is determined by the length of the conductor, the size of transmission line involved, and the frequency of the transmitter. In general, the conductors are chosen of sufficiently large diameter, in order to reduce transmission line losses, that the breakdown voltage comes within the necessary factor of safety. The transmission line used for a 1 kw. transmitter must consist of either one or two transmission lines of 7/8 inch to one inch in diameter, depending upon whether the system of 39A or 39B is chosen. In this case, the breakdown voltage would be on the order of 3000 volts at 50 megacycles, provided ordinary care was used to eliminate any sharp points, which would adversely affect the breakdown rating. Since the normal range of impedances is from about 50 to 150 ohms for this type of feeder, the maximum voltage encountered should not exceed 600 volts for the case of a line properly terminated as is necessary for television.

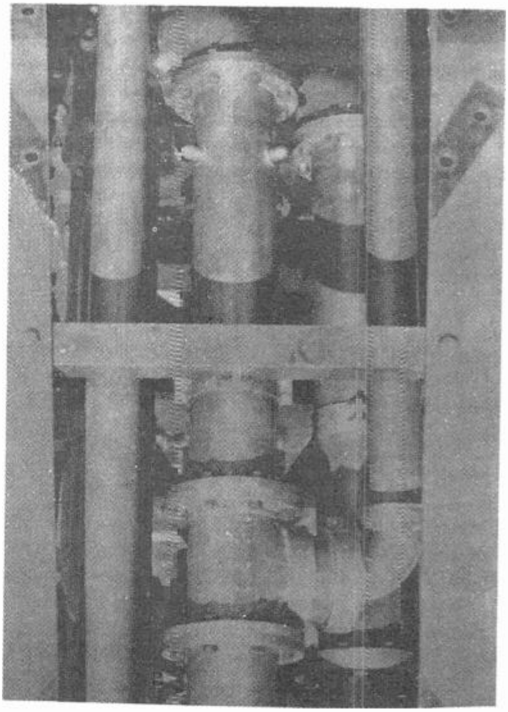


Fig.42 Transmission line of NBC 4 kw. television transmitter.

The power loss goes up as the square root of the frequency and directly as the length of the transmission line. For instance, a typical 1 kw. installation using a standard type of one-inch transmission line, would have a loss of .33 db for 1000 feet at one megacycle. Thus, if this type of line is used for a television transmitter on 51.25 megacycles, the loss involved for a 250-foot line, would be 130 watts as shown below.

$$\text{Loss} = \text{db at 1 megacycle} \times \text{length} \times \sqrt{f} = .33 \times .25\sqrt{51.25}$$

$$\text{Loss} = 130 \text{ watts for a 1000 watt carrier.}$$

It would not be good economy to use this size of transmission line for installations requiring a 250-foot feeder, since the loss encountered is appreciable and expensive to generate in the picture transmitter. Thus, the cost of a larger size of line must be balanced against the cost of producing the wasted power. It also follows that a short transmission line could be constructed using smaller conductors than in the case of the long transmission line. The 4½ kw. British Broadcasting Company transmitter in London uses transmission lines having an outer conductor diameter of five inches and an inner conductor diameter of one and three-eighths inches. The size of the transmission line used in the RCA 4 kw. installation in the Empire State Building can be seen in Fig. 42.

In the concentric type of transmission line, insulators are necessary to space the inner conductor in the exact center of the outer line. In commercial lines for broadcast transmitters, which have been adapted to medium power television transmitters, these insulators consist of isolantite washers spaced every foot or so along the inner conductor. The distance between insulators depends somewhat upon the manufacturer and upon the size of the line used. The larger line requires fewer insulators because of the greater rigidity of the inner conductor. The general construction of the coaxial line of the commercially available variety is shown in Fig. 39D. The insulators not only introduce considerable loss, since they are in the intense electrostatic field, but each insulator raises the capacity of the line at the point of its placement, because the dielectric constant of the insulating material is always greater than unity. The increased capacity causes a reduction in the surge impedance of the line, as can be seen in the following familiar equation.

$$Z_0 = \sqrt{\frac{L}{C}} \qquad (4)$$

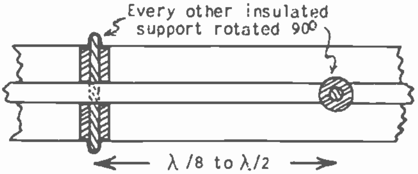
Each insulator causes a reflection of the signal, due to the change in characteristic impedance of the line at that point. The detrimental effect of the reflection is not as bad as would be expected at first thought. Because there are a great number of insulators, there is generally a reflection somewhere else on the line which is 180 degrees out of phase with the reflection from some particular point. The multiplicity of reflections from the various insulators tend to cancel out, and only a small amount of distortion is introduced into the transmitted signal.

If the insulators are widely spaced, and only a few of them are used, they should be constructed so as to have a very low capacity, and should be spaced at some particular fraction of the operating wavelength; either one-quarter wavelength apart, or one-eighth wavelength apart. When placed in this manner, they should be arranged in groups with long distances between groups, with the assurance that the reflection from one insulator will be cancelled by an equal and opposite reflection from another insulator.

In large transmission lines which are built especially for television, ordinary types of insulators are forsaken in favor of the type of insulator illustrated in Fig. 43. A thin rod is passed completely through the inner conductor into projecting cups on the outer conductor. The inner conductor is centrally located by means of small sleeves which slip over the rod insulator. The sleeves are made of the same insulating material as the rod. This type of support eliminates the necessity of crimping the inner conductor or using metallic parts to locate the insulator, thus averting irregularities of the conductor. Furthermore, the rod, because of its smaller size, does not increase the capacity between the inner and outer conductor as much as when the disc type of insulator is used. The type shown in Fig. 43 raises the capacity only .4 mmfd., which changes the characteristic impedance by only approximately .35 ohm for a 75-ohm transmission line.

In the case of the special transmission line illustrated in Fig. 42, the supporting rods for the inner conductors are of quartz, which is the best high frequency insulator available. Quartz is also used at other points in the RCA-NBC installation even to the use of quartz windows to seal the ends of the transmission lines at the throat of the antenna shown in Figs. 21 and 22.

Fig. 43 Insulator detail in large television transmission line.



Bends and junctions are to be avoided in the construction and layout of transmission lines since it is difficult to maintain constant impedance under such conditions. The feeders should be run in as straight a line as possible and wherever bends are to be used, right angles can be formed with the aid of junction boxes as shown in Fig. 44.

Since the impedance of a transmission line is given by formula (4), and since both the inductance and capacitance of the line changes at the junction box, the capacity at this point is made slightly lower and is then increased by the use of a small padder as shown. Thus, by adjusting this padder, the ratio of inductance to capacity can be made such as to maintain the proper characteristic impedance of the line at the junction. Notice also, the expansion joints provided to take care of differences in expansion between the inner and outer conductor.

At several points in this lesson, the use of a quarter-wavelength section of transmission line has been mentioned as an impedance transformer to match between a particular load impedance and the characteristic impedance of a transmission line. Such transformers are occasionally necessary, but should be avoided whenever possible. The use of these transformers *decreases the bandpass characteristics* of the antenna and the larger the mis-match that must be transformed by the quarter-wave section of line, the greater will be the sharpening effect by the use of such a transformer. Suppose the 2¼" diameter dipole of Fig. 24 is to be matched to a 70-ohm line, and to a 150-ohm line. For the two cases, the quarter-wave transformer would be chosen with a characteristic impedance of 65 ohms and 95 ohms. The input impedances to the two transformers are shown in Fig. 45. It is seen that the larger transformer ratio results in a sharper input impedance characteristic having less bandwidth. Thus it is seen that for optimum results, the antenna should be designed to match the feeder with a minimum transformer ratio.

7. MECHANICAL CONSIDERATIONS. Television antennas must be located in high places where wind and ice loads are severe. Consequently, the antenna structure must be physically strong, particularly in the case where it is located on the top of a high building. In this position, there must be absolutely no possibility of breaking, since the falling parts would not only cause property damage, but might endanger pedestrians. For this same reason, as well as for electrical characteristics, ice accumulation on the antenna must be avoided. Falling icicles from the height of a television antenna in a metropolitan area would be quite dangerous. In addition to strength and provision for the removal of ice accumulation, the antenna must be arranged for lightning protection and must conform to the architectural beauty of the building upon which it is erected.

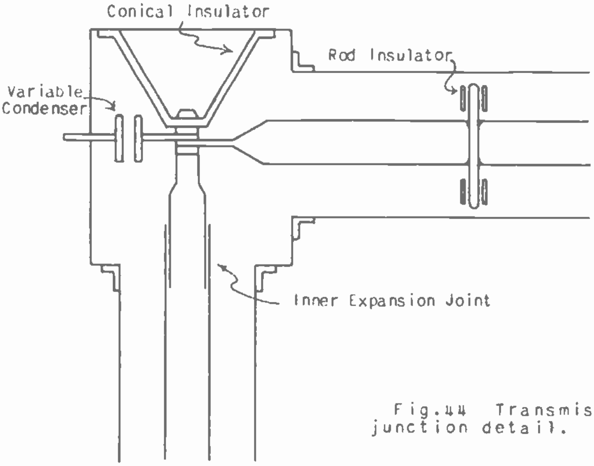


Fig.44 Transmission line junction detail.

The strength of the antenna will depend upon the area that is exposed to the wind and the amount of streamlining involved. The antenna should be capable of withstanding a wind velocity of 100 to 150 m.p.h., with a reasonable factor of safety under these conditions. The antenna illustrated in Fig. 22 was designed to withstand a wind velocity of 130 m.p.h., with a factor of safety of five. In addition to the ordinary strength of the antenna, care should be taken in the design to avoid mechanical resonance which might set up a vibration or an oscillation in the structure when high winds prevail. For example, an ordinary tubular antenna, such as a dipole, tends to develop a circular motion at the tip of the antenna. If the wind velocity is such that the rotational torque induced in the antenna hits resonance with its normal period of vibration, enough motion might be set up in the antenna to tear it loose from its support. This difficulty is overcome by making the antenna rigid as well as strong.

Lightning protection is best accomplished by grounding the center of the antenna, which is done either in the design shown in Fig. 22 or in Fig. 25. When the dipoles are fed in the manner shown in Fig. 25, the support not only affords lightning protection, but results in an extremely strong support for the antenna by avoiding the use of insulators between the antenna and the supporting arm.

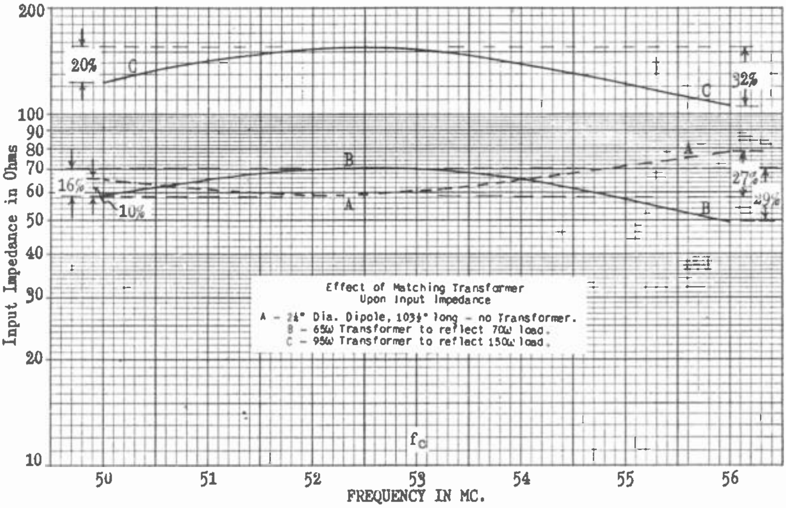


Fig.45 Effect of matching transformer upon input impedance.

When it is necessary to insulate the antenna as in the case of Fig. 22, the antenna element can be supported by a metallic insulator,[1] consisting of a quarter-wave metallic rod plainly visible in the figure. It is generally convenient to make the supporting arm shorter than a quarter-wavelength. In this case, the arm would exhibit an inductive reactance to the antenna. However, if the antenna is arranged to have a capacitive reactance at the carrier frequency, equal to the reactance of the supporting arm, the metallic support will then be resonated, and its bandwidth will be sufficient so that the antenna will not be effectively loaded over the television channel.

The accumulation of ice on an antenna changes its effective capacity and causes leakage between the various parts. Such a condition is particularly bad in the case of open wire feeders where the ice accumulation would upset the characteristic impedance, producing signal reflections and distorting the received picture. It is not always possible to design the antenna so that an accumulation of ice does not upset its characteristics, and if continuous service is to be had, some means must be provided for removing the ice from the antenna. The obvious method is to insert heating units in the antenna, which can be turned on either manually or automatically when the ice accumulates on the antenna. The heaters would melt any ice accumulation and prevent further accumulation of ice. The antenna of Fig. 22 is so equipped. The Calrod heating units and

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[1] Refer to Lesson 4, Unit 7.

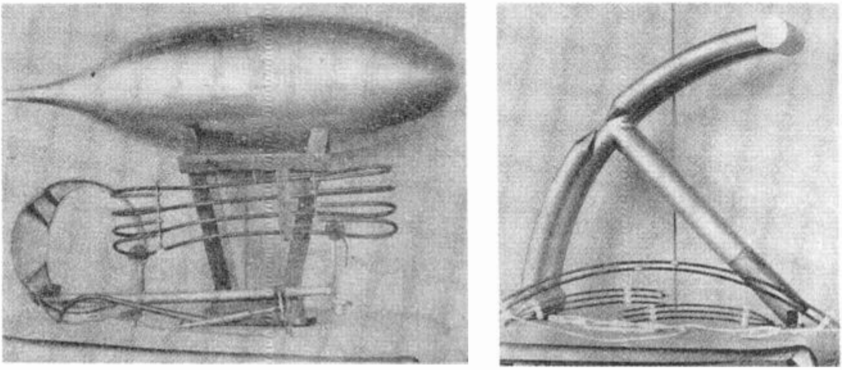

Fig.46 Calrod heating unit for television antenna.

associated parts are shown in Fig. 46, while ice accumulation on this antenna is shown in Fig. 47, just after the heat was applied. Obviously, a great deal of power will be required in severe weather to raise the temperature of such a bulky antenna in the icy wind to which it is subjected. Thus, this particular antenna is equipped with heating units arranged to consume a total power of 27 kw. under maximum conditions.

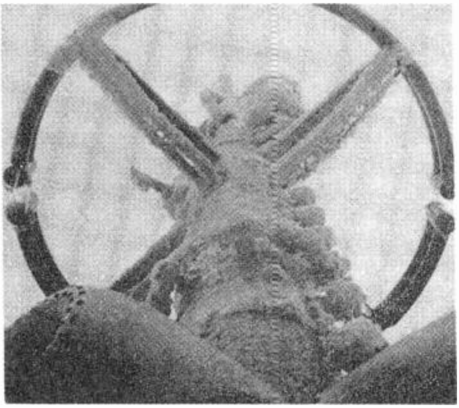

Fig.47 Ice accumulation upon television antenna.

8. ANTENNA ADJUSTMENT. Since any mis-match between the television antenna and the feed line seriously impairs the picture quality, it is necessary to accurately adjust the antenna after it has been erected. These adjustments are ordinarily rather small, since the antenna has already been designed to terminate the feeder. Formulas are available for accurately calculating the self-impedance of the antenna elements and the mutual impedances existing between them.[1] The properties of the transmission line and the transformer

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[1] Directional Antennas, by G. H. Brown, Proceedings of I.R.E., January '37; also High Frequency Models in Antenna Investigation, by G. H. Brown and Ronald King, Proceedings of I.R.E., April '34.

matching sections can also be accurately calculated so that the antenna adjustments after erection are refinements to take into account the effect of the structure which supports the antenna and the effect of any surrounding objects. It is difficult to calculate ahead of time the exact influence of small, non-resonant objects, placed in the field of the antenna.

In order to adjust the impedance of the antenna, three pieces of equipment are necessary: First, a source of variable frequency RF power; second, an accurate and flexible heterodyne type of frequency meter; third, an impedance meter.

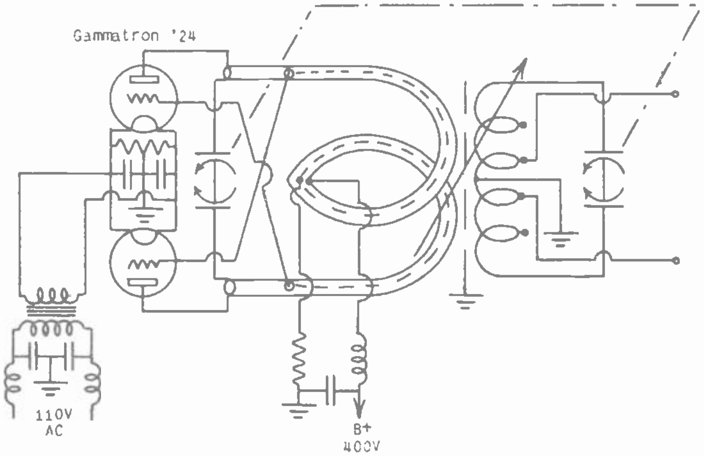


Fig.48 RF test oscillator for antenna adjustment.

The first (high frequency source of power) can be a stable self-excited oscillator which is capable of furnishing either a balanced or an unbalanced output (push-pull or single-ended to ground). The oscillator should be capable of furnishing from 2 to 50 watts, depending upon the type of impedance meter used, but in no case should it be loaded to near its maximum output, as this would decrease the stability of the oscillator. The oscillator should be well shielded to prevent any stray radiations from influencing the antenna performance or the impedance meter. The frequency control should preferably be a single dial, and the coupling to the output should be easily variable over a wide range without materially affecting the frequency control. A circuit diagram of a suitable oscillator is shown in Fig. 48. It employs a unity coupled circuit, with the grid coil wound inside of the copper tubing that forms the plate tank circuit. The output tank circuit is electrostatically shielded from the main tank, and variable coupling is provided. An arrangement of taps permits varying the output impedance, and extreme care is taken to maintain symmetry and balance to ground in the output tank circuit. If single-ended output is desired, a somewhat different output tank would have been necessary, but the balanced type of output is the most generally useful in antenna work.

A suitable frequency meter is the General Radio Type 620A, heterodyne frequency meter, which can be used to measure frequencies up to 300 megacycles with an accuracy of .01%. A picture of this unit is shown in Fig. 49.

Impedance meters useful at television frequencies may be divided into three general classes, each of which may be applied in several forms. The familiar radio frequency bridge, usually employed by broadcast stations, is not useful at television frequencies, since the stray capacities between the various arms of the bridge and ground, as well as the undesirable inductances of the interconnecting leads, complicates the circuit to an impossible degree.

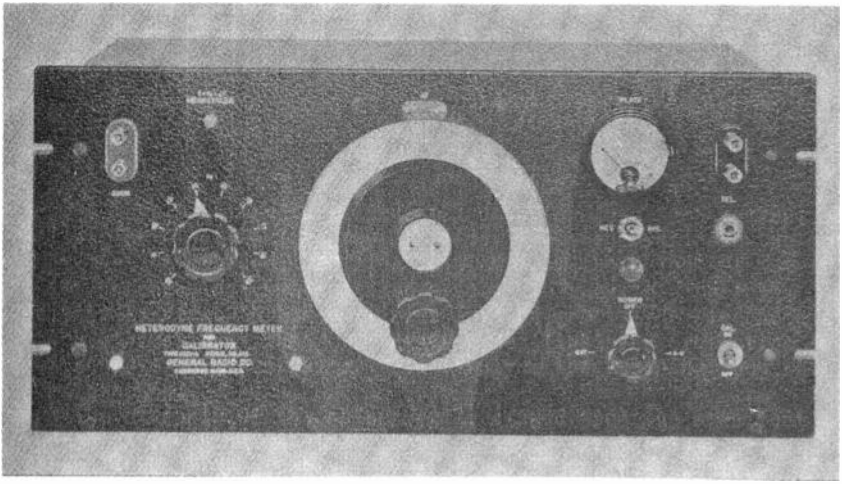


Fig.49 General Radio frequency meter for television services.

The first practical type of impedance meter to be discussed employs the volt—ammeter method. The circuit of such a meter is shown in balanced form in Fig. 50. The tank circuit is coupled to the RF oscillator previously described. The voltage developed across the tank circuit is applied to the load at terminals XX through the two thermo-milliammeters. If the load is perfectly balanced, the readings of the two milliammeters should be identical; but in case slight unbalances exist, the average current of the two milliammeters is taken as the load current. The voltage across the load is the same as the voltages across the tank at YY when the series resonant circuit L_2C_2 is tuned to the frequency of the input signal. The inductance L_2 is simply the unavoidable inductance in the connecting leads, which is generally sufficient to resonate with the condenser C_2. Suppose the load is a pure resistance. In this case, we can apply Ohm's law, and observe the value of the load resistance if the reading of the vacuum tube voltmeter is divided by the average reading of the milliammeters (R = E/I). Since some stray capacity exists across the output terminals indicated as C_4, and also across the vacuum tube voltmeter, this must be canceled by the inductance L_3. Otherwise, the milliammeters will read even when the output terminals are open-circuited. The small condenser C_3 is provided to keep the capacity current between the output terminals at a minimum at all frequencies for which the frequency meter will be used.

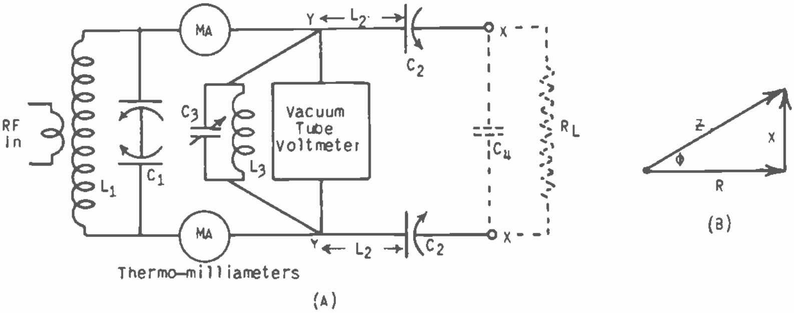


Fig.50 Volt-ammeter type of impedance meter.

Before applying the meter to a more complicated type of load, it is necessary to obtain a calibration of the setting of the condenser C_2. This is done in the following manner. A carbon resistor of known value is placed across the output terminals, and the condenser C_2 rotated to obtain minimum reading on the vacuum tube voltmeter. This will occur when the series resonant circuit L_2C_2 is tuned to the frequency of the RF generator. Under these conditions, minimum impedance is reflected across the terminals YY, and this impedance is exactly equal to the load resistance R_L. This is done for several frequencies through the television band, and thus, a setting of C_2 is obtained for any particular value of load resistance at any frequency. When this calibration curve is obtained, we can proceed to measure the value of any unknown load across the terminals XX.

With the terminals XX open-circuited, C_1 is tuned for maximum reading on the vacuum tube voltmeter, indicating resonance of the input circuit. Condenser C_3 is tuned for minimum reading on the thermo-milliammeter, indicating that the stray capacity across the output terminals has been balanced out. Next, the unknown load is placed across terminals XX, and condenser C_2 is rotated for minimum reading of the vacuum tube voltmeter. The load across YY in this case will be a pure resistance, because any reactance in the output circuit has been neutralized by adding or subtracting capacitive reactance with condenser C_2. This was automatically done when C_2 was tuned for minimum reading on the tube voltmeter. The resistive component of the load is found by dividing the reading of the vacuum tube voltmeter by the average milliammeter reading.

This is indicated by the vector R in Fig. 50B. From the calibration curve of C_2, the known frequency, and the resistance just measured, the correct setting of C_2 can be found to produce resonance of L_2C_2. If all other adjustments remain unaltered, it will be noticed that the reading of both the vacuum tube voltmeter and the milliammeter change as C_2 is rotated to its correct position. These new readings are taken, and the load impedance is given by: $Z = E/I$. Thus, we have found the impedance of the load and its resistive component. The reactance of the load is given by the following equation:

$$X = \sqrt{Z^2 - R^2} \quad \text{(see Fig. 50B).}$$

Whether the reactance of the load is capacitive or inductive will be determined by whether it was necessary to add or subtract capacity from condenser C_2 in order to balance out the reactance of the load. When obtaining the resistance component of the load, if the setting of C_2 was at a lower value than the value indicated by the calibration chart, then it can be deducted that the load circuit has inductive reactance. If the setting of C_2 was higher than normal, then capacitive reactance exists in the load circuit.

Space does not permit the complete description of all the available varieties of impedance meters, but the principles will be outlined.

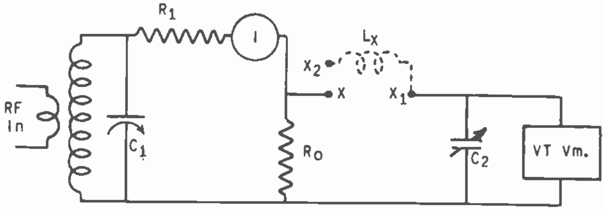


Fig.51 Q-meter.

One circuit suitable for measuring ultra-high frequency impedances is known as a Q-meter. The basic diagram is shown in Fig. 51. The principle consists of measuring the voltage rise in a tuned circuit when it is tuned to resonance. A tuned circuit supplies energy to a large resistance R_1 and a small resistance R_o. The latter is made quite small so that the voltage across it is constant, regardless of the tuning of C_2. The unknown load is connected at points XX_1, and resonated by the capacity C_2. The capacity of C_2 to produce resonance, is a measure of the inductive reactance of the load, while the vacuum tube voltmeter which measures the voltage rise across the condenser, in the series resonant circuit, can be calibrated in terms of the Q of the unknown load.

Condenser C_2 is of the low loss variety, usually employing quartz insulation. If the unknown load has capacitive reactance, it is necessary to insert a known inductance, such as L_x, in series with the load in order that it can be resonated by condenser C_2. Having determined *the reactance and the Q* of the unknown circuit, the quotient is the resistive component of the impedance ($R = X/Q$). The impedance of the unknown is given by $Z = \sqrt{R^2 + X^2}$.

The most widely used variety of high frequency impedance meter places the load either across a tuned circuit or in series with it. The value of the unknown load is indicated by the change of voltage or current in the tuned circuit. The principle is employed in several ways, but the basic diagram is shown in Figs. 52A and B. The series resonant circuit is used to measure low values of impedance up to approximately 300 ohms, while the shunt or parallel resonant circuit shown in Fig. 52B, is used to measure higher values of impedance. The circuit is first resonated with the output terminals short-circuited and the coupling adjusted for full scale reading of the current meter. The unknown impedance is then placed across the output terminals and a new current reading is observed. The current in each of these two cases will be inversely proportional to the total resistance in the circuit. In the first case, it was simply the unavoidable resistance of the tuned circuit, while in the second case, it was the unavoidable resistance plus the load resistance. Thus, it is possible to determine the resistive component of the unknown impedance by the drop in the current meter, if the unavoidable resistance in the tank circuit is known. The reactive component of the load can be determined by the amount of reactance which had to be added or subtracted by condenser C_1 to maintain resonance when the unknown impedance was inserted.

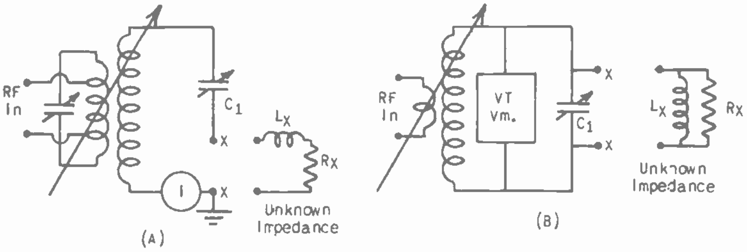


Fig.52 Impedance meter which measures tank loss.

The parallel resonant circuit is worked in the same manner, except that the voltage across the tank is observed instead of the current through it. In this case, as in the previous one, the change of capacitive reactance of the standard condenser is a measure of the unknown reactance, while the resistance of the unknown impedance is calculated by the ratio of the tank voltage before and after the unknown impedance was added.

Another way of determining the resistive component of the unknown impedance is to resonate the tuned circuit with the unknown impedance connected, and observe the current or the voltage reading, (depending upon whether Fig. 52A or Fig. 52B is employed). Next, the standard condenser C_1 is rotated until the current or voltage reading drops to .707 of its original value. With this tuning obtained, the change of reactance of the standard condenser from its original setting is equal to the *total* resistance of the circuit. The total resistance is the unknown load resistance plus the equivalent resistance of the tuned circuit.

Both of the preceding methods depend upon knowing the equivalent resistance of the tuned circuit. This is found by using the meter to measure a known resistance. The meter reading will be proportional to the total resistance in each case:

R = known resistance. R_c = circuit resistance.
M_1 and M_2 are the meter readings.

$$\frac{M_1}{M_2} = \frac{R_c}{R + R_c} \qquad \text{or} \qquad R_c = R\left(\frac{M_1}{M_2 - M_1}\right)$$

Thus, the circuit resistance can be found.

A more accurate way of determining the resistive component of the unknown impedance is to substitute known values of resistance until the same value of tank current or tank voltage is reached as with the unknown impedance connected. In this case, the substitution resistor will equal the resistance of the unknown impedance. This method is limited, inasmuch as it is difficult to obtain a sufficiently large number of fixed resistors for substitution. It is impossible to incorporate a variable resistance at this frequency because of the inductance of the switches and connecting leads.

All of the methods described for measuring impedances at ultrahigh frequencies are subject to some error, due to unavoidable stray inductances, capacitances, and resistance losses in the leads. If all of these were taken into account, the mathematical manipulation necessary to obtain the true resistance and reactance from the measurement would be excessive. Therefore, when precise measurements are necessary, the impedance meter should be calibrated with known values of capacitance, inductance, and resistance, and a series of curves drawn for various frequencies throughout the television band. Known values of capacitive reactance and inductive reactance can be obtained for calibration purposes by the use of sections of transmission line, the reactance of which can be very accurately calculated.

It is necessary to apply sections of transmission line as standards to calibrate the standard condenser of the impedance meter, since even a small air-type tuning condenser has enough inductance in the condenser plates and in the connections to the stacks of plates to cause the effective capacity of the condenser to vary widely with frequency.

Standard resistors for calibration purposes are fortunately easy to obtain. One-half watt or one watt resistors maintain their low frequency values of resistance even at television frequencies, for resistances of from 20 ohms to 10,000 ohms. Either the metallized coating on glass or the small carbon type of resistor is satisfactory, and the ceramic covered variety is preferable. Molded bakelite covering of resistors is not satisfactory, especially for values above 1000 ohms, since the losses in the bakelite approximate the losses in the resistor itself. Very small values of resistance are not satisfactory, since the inductance of the resistor becomes appreciable and skin effect comes into play. But, for practical sizes of resistors, that is, about 10 or 20 ohms, this is not very important. Resistors above about 10,000 ohms have a bad frequency characteristic because of the distributed capacity involved. These resistors can still be used as standards if they are previously calibrated by a method which will presently be discussed. Wire leads on the resistors should be made as small as possible, since their inductance affects the accuracy of the standard.

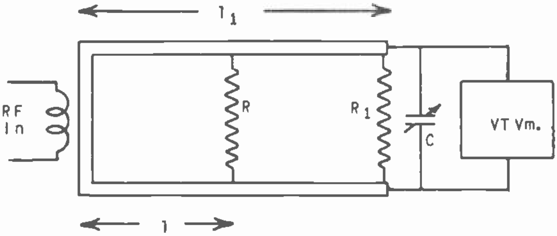


Fig.53 Calibration of standard resistors.

In measuring high values of resistance, it is possible to use low resistance standards by taking advantage of the voltage distribution along a transmission line. The set-up is shown in Fig. 53. In this case, the unknown resistor R_1 is placed across the end of the transmission line which is very much shorter than one-quarter wavelength and tuned to resonance by condenser C. The unknown resistor is then removed, and the known resistor slid along the line until the voltmeter reading is the same as the previous one. When the line is very short, the voltage distribution along the line varies directly as the distance from the short-circuited end. Therefore, the unknown resistor and the known resistor bear a very definite relation, depending upon the distance they are placed from the short-circuiting bar to achieve the same voltage developed across the tank. The unknown resistor is found as follows:

$$\frac{R_1}{R} = \frac{l_1}{l} \qquad R_1 = R\frac{l_1}{l}$$

We are now ready to see how the impedance meter can be applied in making adjustments to an antenna system after it has been designed and erected. The impedance meter can be carried up the antenna tower and connected between the end of the main feeder from the transmitter and the various antenna elements. If the elements are center-fed, the condition of zero reactance will occur at the resonant frequency of the antenna elements. Thus, by plotting the impedance curve, it can be determined whether a small amount should be added or subtracted from the antenna length to make its resonant frequency agree with the calculated value. The resonant frequency of a simple dipole is denoted as f_c in Fig. 32. If this had occurred at too low a frequency, a fraction of an inch would be removed from the ends of the dipole and the frequency run repeated.

In the antenna of Fig. 30 and Fig. 31, the impedance meter should be placed between the main feeder and the branch feeder, with the branch feeder containing the corrective network. Impedance measurements are taken over the television band and small alterations are made in the correction circuit. Thus, it is possible to observe the effects of the correction circuit in flattening the input impedance to the antenna, and accurate adjustments arrived at.

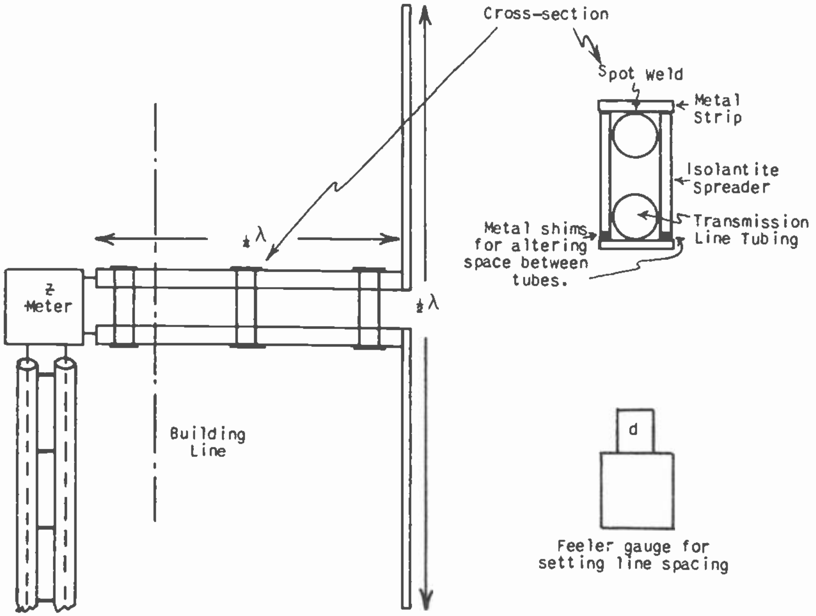


Fig.54 Quarter-wave matching section for antenna adjustment.

Another use for the impedance meter is in the adjustment of quarter-wave matching sections which match between the individual antenna elements and the transmission line. If the impedance meter is placed between the transmission line and the quarter-wave transformers, the spacing of these transformers can be varied a thirty-second of an inch at a time, until the antenna impedance reflected by the quarter-wave section of line is of the desired value. This application is shown in Fig. 54, which also shows the method of altering the spacing of the line as the adjustments are made.

The impedance meter is next used in measuring the characteristic impedance of the main transmission line, and adjusting it to remove irregularities. The irregularities occur at the bends in the line, and are compensated by the method shown in Fig. 44. The transmission line is terminated in its characteristic impedance (instead of the antenna) and the small trimmer condensers at the junction boxes of Fig. 44 are adjusted to remove impedance irregularities.

After the lengths of the antenna elements have been adjusted and the spacing of the quarter-wave matching section adjusted, as well as the wideband correction network, the impedance meter is removed from the upper end of the transmission line and connected to the transmitter end of the main feeder. A new impedance run is taken at this point, and further adjustments are made to the individual adjustments of the antenna system, carefully noting the changes in the character of the line input impedance. If the antenna is supported on an open structure, the presence of the meter and the observer in the field of the antenna may have upset the impedance characteristics. For this reason, additional adjustments are made with the impedance meter located at the transmitter.

The input impedance to a long transmission line depends not only upon the termination of the line, but upon the line length and the frequency at which the impedance is measured. If the line is an even number of quarter-waves long, the input impedance is the same as the terminal impedance; or if the line is an odd number of quarter-waves long, the input impedance is given as follows:

$$Z_{in} = \frac{Z_0^2}{Z_{out}}$$

Since the feed line is usually one hundred to several hundred feet long, it will be an even number of quarter-waves long and an odd number of quarter-waves long at several frequencies throughout a single television channel. Thus, if the transmission line is terminated in an impedance such as Fig. 55A, the input impedance to the transmission line will be equal to curve *a* at some frequencies, and equal to *b* at other frequencies. Thus, these two curves form the envelope of the actual impedance, which varies between these limits, as shown in Fig. 55B.

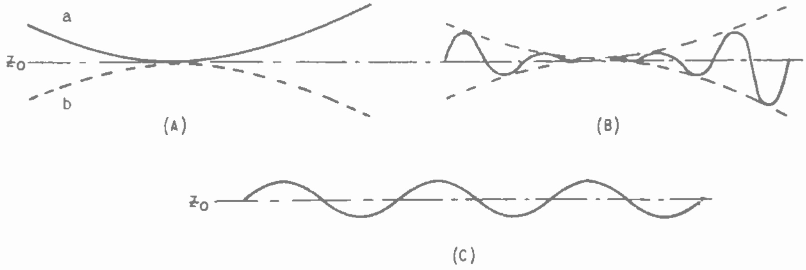


Fig.55 Impedance variation of input to transmission line.

By purposely mis-terminating the transmission line in a known resistance, and measuring the input impedance over the television band, as shown by Fig. 55C, the exact electrical length of the transmission line can be determined.

With the input impedance to the line measured, and the length and characteristic impedance of the line known, it is possible to calculate backward and determine the load impedance on the transmission line due to the antenna, and thus decide what adjustments are necessary to improve the overall impedance characteristics.

9. SIGNAL COMBINING NETWORKS. At the present time, the majority of television installations have separate antennas for the sound and television transmitters. Most of the larger companies have experimented with networks for operating the two transmitters on a single antenna, but only a few of them have been satisfied with the results. There is no standard type of network used for this application. Each company or station has developed a method of its own, which is subsequently witheld from all other companies pending action on their patent applications.

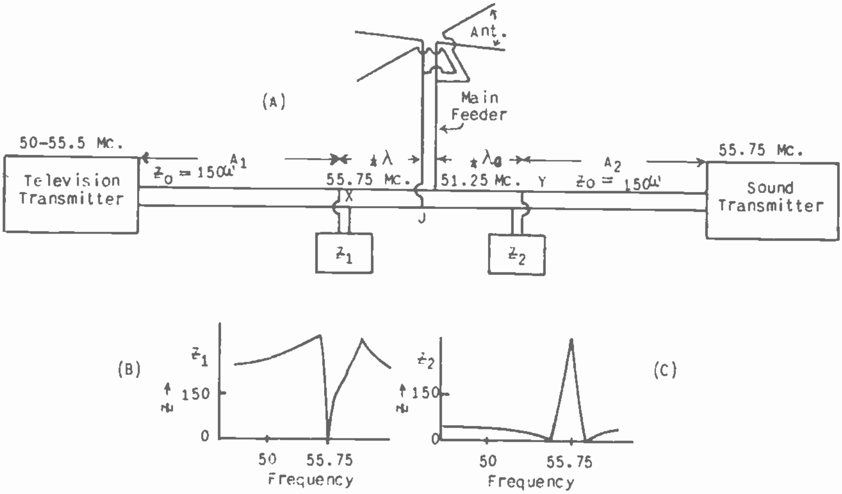


Fig.56 Network for combining sound and picture signals into a single antenna.

However, the general principle employed in all of these installations is the same, and will be described in conjunction with Fig. 56. The television transmitter and the sound transmitter are connected to the main antenna feeder by a balanced transmission line. For simplicity, they are shown as open wire lines, but in general, concentric types of transmission lines would be used. Across the branch transmission lines are placed two filter networks, consisting of from 4 to 8 sections of transmission line. The network Z_1 is placed, as shown, one-quarter wavelength away from the main feeder, but between the television transmitter and the main feeder. The impedance of this network is shown in Fig. 56B. It has a high impedance over the television band and series resonates in such a way as to reflect a short-circuit across the transmission line at the frequency of the sound transmitter. A short-circuit at the sound frequency prevents any energy from the sound transmitter from traveling back line A_1 to the picture transmitter.

The low impedance which results across point X at the frequency of the sound carrier, is reflected through the quarter-wave section of transmission line, tuned to the frequency of the sound transmitter, and appears as a very high impedance across the junction J. The main antenna feeder also appears across this point and will have an impedance equal to the characteristic impedance of the transmission line, since it is carefully terminated by the antenna. Thus, the sound transmitter sends its energy out along the branch feeder to the junction J, where two impedances appear in parallel. One is the relatively low impedance of the antenna system, (in the vicinity of 150 ohms) and the other is a very high impedance reflected by the section of transmission line XJ. Since this impedance is extremely high, all of the energy from the sound transmitter flows into the main antenna feeder, and practically no current flows to the television transmitter. What little current flows into this line is short-circuited at point X, and is dissipated as circulating energy in the short-circuiting network.

The result is that practically no energy passes from the sound transmitter to the television transmitter to cross-modulate with the television transmitter and produce objectionable interference between the two. Furthermore, neither the television transmitter nor the short-circuiting network Z_1 have any loading effect upon the sound transmitter. The sound transmitter feeds all of its energy into the antenna circuit and behaves exactly as if the television feeders were completely disconnected from point J.

The network Z_1 has a high impedance over the picture channel. Although it effectively shorts out the feeder as far as energy from the sound transmitter is concerned, it has nevertheless, very little loading effect upon the television transmitter, since it is a fairly easy matter to make the impedance over the picture channel equal to at least 10 times the characteristic impedance of the branch feeder A_1. Thus, only a small portion of the output from the television transmitter circulates in the network Z_1, and the television transmitter, as well as the sound transmitter, behaves almost as if no loading were placed upon it other than the antenna load reflected by the main feeder.

The same condition holds for any energy which would tend to circulate from the television transmitter toward the sound transmitter to cause undesirable cross-modulation in the output stages of the sound transmitter. However, a different type of network must be placed across the feeder to the sound transmitter. It will be seen that this network is exactly the inverse of the network which we previously considered. Therefore, it can either be made up as a complementary network, or the same network connected through a quarter-wave section of transmission line. The desirable characteristics of network Z_2 are shown in Fig. 56C. It is seen that we wish to have a low impedance over the television band and a very high impedance at the frequency of the sound transmitter. The low impedance over the picture channel short-circuits point Y, and keeps any energy from the television transmitter from entering the feeder A_2, which leads to the output stages of the sound transmitter.

The low impedance at point Y is inverted through a quarter-wave section of transmission line tuned to the television picture carrier, and appears across point J as a very high impedance. This high impedance is in parallel with the impedance of the main antenna feeder, and therefore, practically all of the energy from the television transmitter enters the antenna feeder, and only a small portion flows into the high impedance looking toward the sound transmitter. Of the energy that does enter line JY, most of this is dissipated as circulating energy in the network Z_2 because of the low impedance placed across point Y by this network. Therefore, the television transmitter behaves as if neither the sound transmitter nor the network Z_2 were present, with only the antenna connected to the output of the television transmitter. Furthermore, the parallel resonance of network Z_2 at the sound transmitter frequency places an extremely high impedance across the transmission line at point Y, and therefore, the sound transmitter functions as if the network Z_2 were not present.

Each transmitter works into the impedance of the antenna in a normal way and behaves as if the other transmitter, with its associated feeders and networks, was not present.

The networks Z_1 and Z_2 are formed of extremely low loss elements (sections of transmission line) and thus behave like pure reactances at frequencies slightly removed from the resonant frequencies of the networks. The output stages of the transmitter, plus the transmission line sections A_1 and A_2, also behave as low loss circuits for any energy feeding back toward the transmitter at a frequency other than that of the carrier. This condition gives rise to a possibility which may destroy the otherwise smooth operation of the combining network.

Consider point Y. This must have a low impedance over the entire television picture channel. However, the low impedance formed by network Z_2 is a pure reactance and if it happens that the transmission line A_2, in conjunction with the output stage of the sound transmitter, becomes an equal and opposite reactance, parallel resonance might occur at point Y, and thus nullify the effectiveness of the network Z_2. The same thing is true of Z_1 except this case is not quite so critical, since there is only one frequency, namely the sound carrier frequency, at which parallel resonance could cause any difficulty. It is not likely that the television transmitter plus feeder A_1 would happen to be a low impedance at the sound carrier frequency. Even if this condition did occur, slight alteration of the length of transmission line A_1 would clear up the difficulty.

The condition for point Y is not quite so simple. The network Z_2 presents a low impedance over a wide range of frequencies, and it is quite possible that this reactance will be neutralized by the line A_2 and the sound output tank, at some frequency within the picture channel. Fortunately, the reactance of network Z_2 is only on the order of a few ohms, not over 10 ohms at the most. If the reactance of line A_2, plus the output of the sound transmitter, can be made equal to at least 10 times the reactance of Z_2 over the television picture channel, no danger of parallel resonance will occur. As a matter of fact, if the reactance of the sound transmitter plus the line A_2 happens to have the same sign as the impedance Z_2, the performance of the network will be improved, rather than impaired. That is, if network Z_2 presents a capacitive reactance across position Y, the impedance across this point would be lowered rather than increased by a capacitive reactance of line A_2.

If the sound transmitter is placed fairly close to point Y, thus making the transmission line A_2 reasonably short, a length of transmission line A_2 can be chosen which will in no way impair the operation of the combining network.

10. SINGLE SIDEBAND FILTER. The performance characteristics of single sideband filters were set forth in Lesson 4 of this Unit. The physical arrangement for accomplishing the desired results will now be considered.

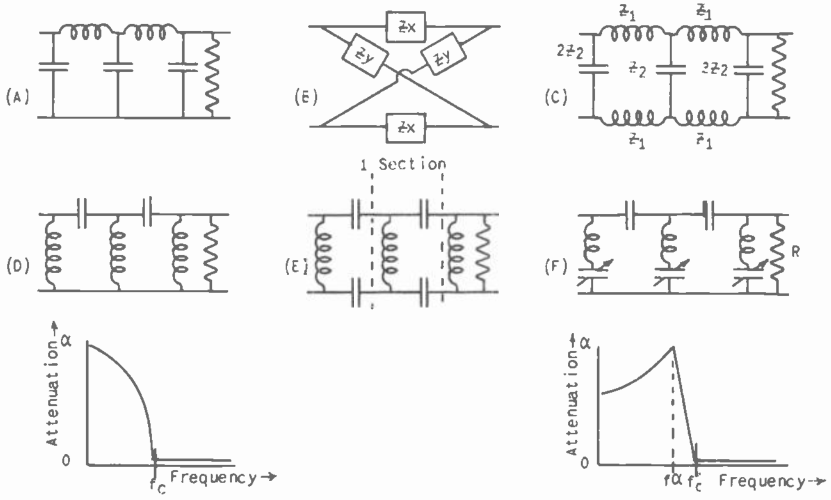


Fig.57 Basic filter network.

Consideration has been given to generating television signals at low levels and at lower frequencies where filters using ordinary coils and condensers can be developed to remove one of the television sidebands. In this case, amplifier tubes could be placed between various filter sections and thus make it possible to design a filter having the desired characteristics, especially in regard to a negligible phase distortion. The single sideband energy would be subsequently multiplied in frequency and in power to obtain the desired output. This system has not as yet been tried outside of the laboratories and will not be discussed here.

High power filters employed at ultra-high frequencies, whether placed between the modulated stage and linear amplifier, or between the linear amplifier and the antenna, are of the same general type. These always consist of sections of concentric transmission lines rather than lumped constants, although in some cases, air dielectric condensers are used in conjunction with transmission lines to form the desired network.

The student is already familiar with "low-pass ladder type" filters, even though the term applied may not be familiar. This is the ordinary type of filter used in power supplies, the basic diagram of which is shown in Fig. 57A. As we progress through the study of filter circuits, many new terms will arise with which the student should familiarize himself in order that future outside reading will have more meaning.

In power supplies we started out with a pulsating DC current which contained a DC component, plus a 60-cycle wave, plus harmonics of 60 cycles. It was desired to remove everything above 20 or 30 cycles so that only the DC component would be passed by the filter. This explains the use of the term *low-pass filter*. The term, *ladder type* of filter, is derived from the physical appearance of the network as distinguished from the lattice type of structure of Fig. 57B. The low-pass ladder filter, as used in a power supply is of the unbalanced form; that is, one side is grounded. It can also be used in the balanced form as shown in Fig. 57C. In either case only the low frequencies are passed.

In television we wish to pass the high frequency sidebands of the television picture transmitter and to suppress the low frequency sidebands. This type of filter is known as a *high-pass filter*, and as would be expected, the circuit elements are reversed from Figs. 57A or 57C, and shown in Figs. 57D and 57E. Obviously, the blocking condenser prevents DC from being passed by such a filter and we say that it has *infinite attenuation* at zero frequency. As the frequency is increased, the attenuation of the filter decreases until the nominal *cutoff frequency* is reached. At this frequency, the filter will pass all of the energy that is applied to it and the voltage across the output will equal the voltage across the input to the filter. At least such would be the case if an infinite number of *filter sections* had been employed. The term filter section denotes that part of the network enclosed within the dotted lines of Fig. 57E. Because of the gradual *cutoff characteristic*, this type of filter is not particularly suitable for television application, where an extremely sharp cutoff is desired; but it forms the basis of the more complicated types to be discussed presently. The filter which we have been considering is known as the *constant K* type of high-pass filter, and is distinguished by a gradual cutoff characteristic.

A sharp cutoff calls for a type of filter known as the *M-derived* type, the fundamental diagram of which is shown in Fig. 57F. The shunt coils of Fig. 57D have been replaced with series tuned circuits, which resonate at frequencies below the cutoff frequency. Neglecting losses, this causes the filter to have infinite attenuation at such frequencies, as shown by the attenuation characteristics immediately below the circuit diagrams.

The series tuned circuits effectively short-circuit the network at their resonant frequencies. One of these tuned circuits is adjusted to the sound carrier of the adjacent television band in order to completely eliminate transmission at this frequency. Other sections of the filter are adjusted to series resonate at other frequencies throughout the adjacent television band in order to produce additional *infinite attenuation points* for the unwanted sideband of the television picture transmitted. When several sections of filter are employed, and these sections are adjusted for different rejection frequencies, the network is called a *composite filter*, since it consists of several different sections.

The attenuation curves shown in Fig. 57D and Fig. 57F assume that an infinite number of filter sections are employed, and that the elements of the filter have zero loss. Obviously, neither of these conditions is a practical one. A typical filter consists of five sections, employing elements with Q's of approximately 10,000. By employing a limited number of sections, it is not possible to terminate the filter in its exact *image impedance* over the television band, since this can only be done by terminating it in an infinite chain of additional sections. Actually, the filter is terminated in as nearly a constant resistance as can be achieved by the antenna and its transmission lines. The effect of this is to round off the corners of the attenuation characteristic at the cutoff frequency.

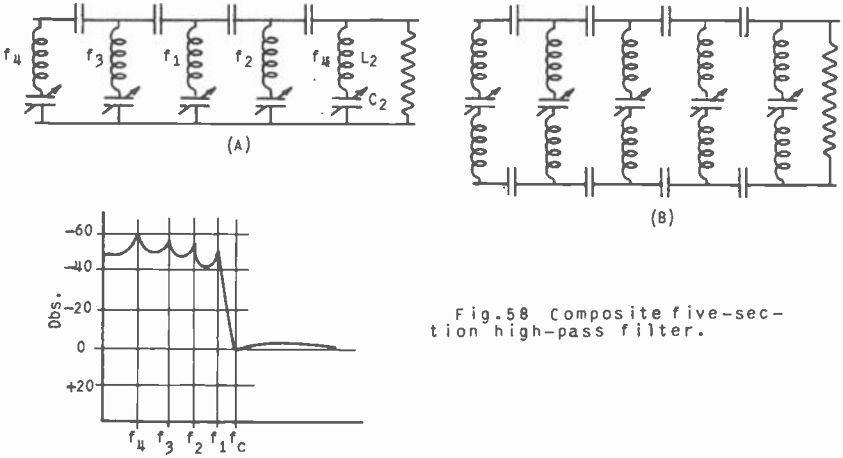


Fig.58 Composite five-section high-pass filter.

A typical five-section filter and its attenuation characteristic is shown in Fig. 58A, while the balanced form of the same filter is shown in Fig. 58B. Actually the filter is a four-section filter, since the input and output sections are each *half-sections*. That is, the reactances of L_2 and C_2 at resonance are doubled over the normal values chosen in the basic design. Terminating the filter in a half-section, makes it possible for a pure resistance to more nearly match the impedance of the filter over the television band.

In order for the impedance match to be as nearly perfect as possible over the entire pass band, it has been customary to choose the input and output sections of this type of filter so that M = 0.6. M is given by the following formula:

$$M = \sqrt{1 - \left(\frac{f\infty}{fc}\right)^2}$$

where $f\infty$ is the frequency at which the output half-section produces an infinite attenuation, and fc is the nominal cutoff frequency.

For television purposes, we are not interested in the entire pass band of the filter, but only a small band near the cutoff frequency. If the carrier is placed at 51.25 megacycles, and a cutoff frequency is chosen at 50.5 megacycles, then the pass band would extend from 50.5 to 55.5 megacycles, as explained in Section 8 of Lesson 5. Thus, we are interested in a narrow band, within approximately 10% of the cutoff frequency. Unfortunately, this is the very region which has the poorest characteristic in normal filter design. Thus, for television application, it is better to choose a filter with M equal to .4 and design it to have a characteristic impedance equal to twice the impedance of the antenna feeder. The actual formulas are omitted from this lesson, inasmuch as a considerable amount of supplementary reading and study would be necessary to apply them in a particular example.

It was mentioned that sections of transmission line were used to replace the circuit elements of ordinary filter circuits. Thus the series tuned circuits of Fig. 58B could be replaced by quarter-wave sections of balanced transmission lines open-circuited at the far ends. The characteristic impedance of these transmission lines would be chosen so that the reactance curves over the television band approximated that of the tuned circuit shown in Fig. 58B. If this gave an unreasonably high value of characteristic impedance, the situation could be improved by using half-wave sections of transmission lines short-circuited at the far ends. The series condensers could either be formed from sections of transmission line, or could be built up as air dielectric condensers using aluminum plates. The circuit would then become that of Fig. 59A, in which case the transmission lines would be bonded together to give the general appearance of Fig. 59B.

The other characteristic of filters which must be considered is the phase characteristic. The detrimental effect of phase distortion was explained in Section 8 of Lesson 5, Unit 7. Minimizing the phase distortion is an intricate problem of filter design tied up with the choice of filter sections and will not be considered here. In passing, it should be mentioned that it is not always possible to reduce the phase and amplitude distortion introduced by the single sideband filter to a sufficiently low degree. In this case, it becomes necessary to employ correction circuits at video frequencies. These circuits accentuate the video frequencies which in turn are reduced by the filters. Phase correction networks are employed, which delay all the video frequencies except the ones which are delayed by the single sideband filter, thus correcting the distortion inserted by the latter. The amount of correction that can be employed by this method is limited, particularly with regard to amplitude correction. Non-linearity in the amplifier stages may

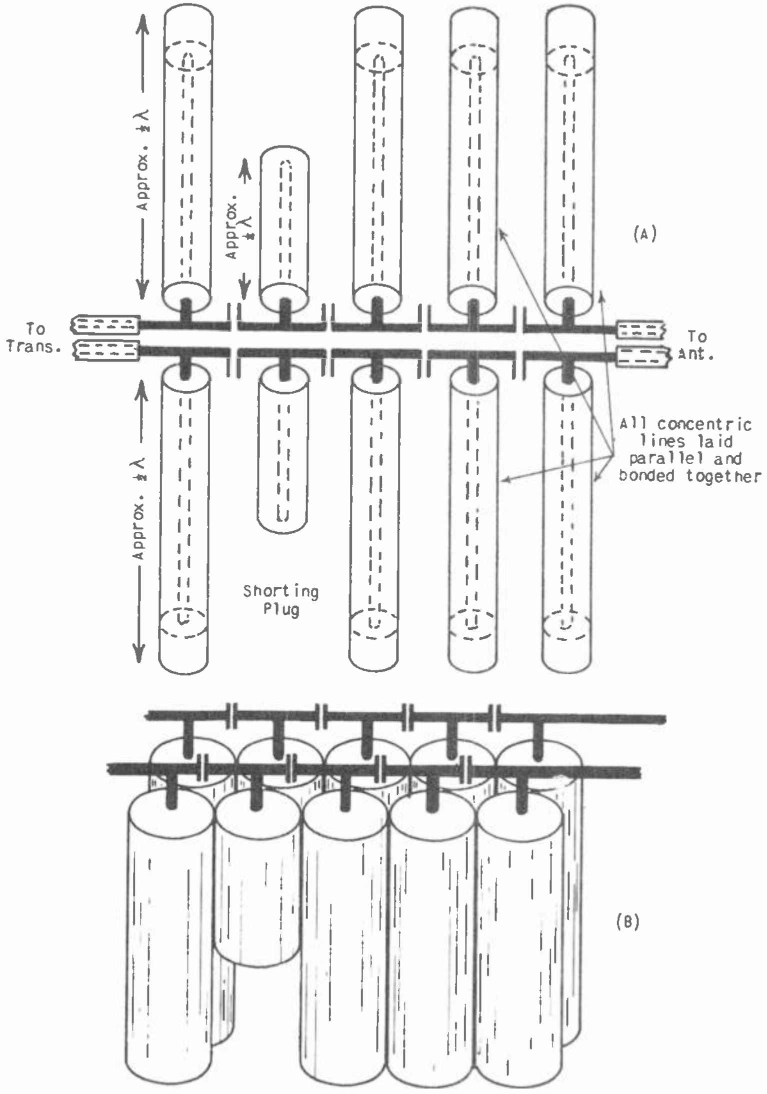


Fig.59 Five-section high-pass filter employing sections of transmission line.

result in correcting the frequency discrimination for positive modulation peaks, but not for negative ones. Furthermore, if too much video amplitude correction is involved, the transmitter will be over-modulated at certain video frequencies.

Still another problem, namely securing a constant load resistance for the transmitter to work into, is more easily solved. It is possible to combine a low-pass filter and a high-pass filter in such a way that any change of impedance in the high-pass filter is offset by an equal and opposite change in the impedance of the low-pass filter. The combination is another type of constant resistance network, one form of which has been studied earlier in the lesson. The low-pass filter, which passes the unwanted sideband, is terminated in a dummy resistor. This dissipates the undesired lower sidebands of the picture transmitter.

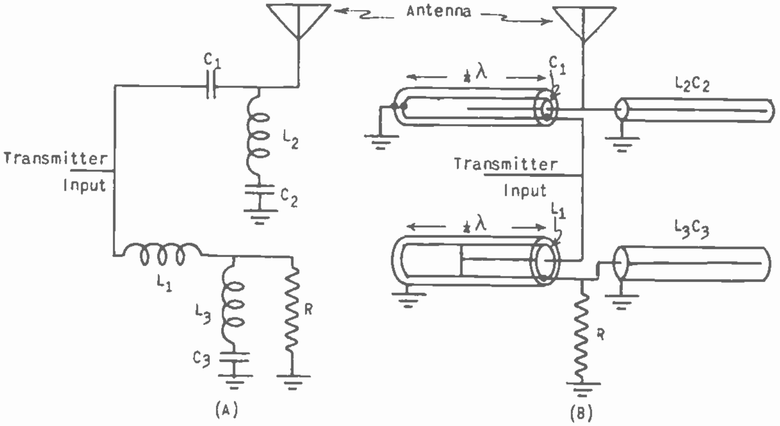


Fig.60 Single-sideband filter of the constant resistance type.

The filter shown in Fig. 60A is a single-section filter of the single-ended type, but in practice, the principle would be extended to five or six sections (either single-ended or balanced to ground) in order to obtain sufficient attenuation of all the unwanted sidebands. The application of transmission line sections to this type of filter is shown in Fig. 60B. Here it is seen that not only the tuned circuits, but also the capacitance and inductance of the *series arms* are simulated by sections of transmission lines. In order to correlate Figs. 60A and 60B, the same lettering has been applied to corresponding filter elements. The series resonant circuits are formed by quarter-wave sections of transmission lines, open-circuited at the far end. The characteristic impedances of these lines are chosen to agree with the slope of the reactance curves of the tuned circuits of Fig. 60A.

A section of transmission line shorter than a quarter-wavelength will behave as a capacity if it is open-circuited at the far end and will behave as an inductance when it is short-circuited at the far end. Therefore, sections of transmission line can be made to serve as L_1 and C_1. These transmission line sections must be so arranged that they do not produce an appreciable impedance to ground. Furthermore, they must be balanced. To obtain these characteristics, it is necessary to mount the condenser or inductor transmission line within another transmission line, as shown in

Fig. 60B, to obtain an extremely high impedance to ground. The operation of this arrangement has been described earlier in the lesson under the term of "balance converter".

If the single sideband filter is employed between the final stage of the transmitter and the antenna circuit, it is necessary to employ excellent shielding of the final stage in order to prevent the unwanted sidebands from being radiated directly from the final tank. If it is desired to maintain a stop band 40 db below the pass band in the filter circuit, then it is necessary to shield the final amplifier so that the energy radiated from it is less than 1 part in 10,000 of the energy radiated by the antenna. If a filter is employed between the modulated amplifier and the linear amplifier, it becomes necessary to secure almost perfect linearity in the linear amplifier stages; otherwise, the unwanted sidebands will be reinserted as a modulation product between the upper sidebands and the carrier. Any non-linearity in the stages themselves would cause cross-modulation to occur between these two frequencies.

The overall structure, including the antenna, the transmission line, and the single sideband filter employs a maze of copper tubing sometimes exceeding 6" to 8" in diameter in order to obtain the desired Q. The unusual "plumbing" which results, and the excellent results which can be achieved by departing from the orthodox, are a source of intrigue to the radio men and engineers who work on these problems.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. Name the important considerations in a comparison of horizontal and vertical polarization. Which type of antenna possesses superior qualities on each point?

2. Using two half-wave antennas, show how they may be connected to secure general coverage in all directions. What is such a connection called?

3. If the antenna consisted of two dipoles, how could they be arranged to secure radiation parallel to the earth's surface?

4. Outline five methods that are used to flatten the input impedance of a television antenna over a wide band of frequencies.

5. What consideration is important in the design of sound antennas, and how are the desired results secured?

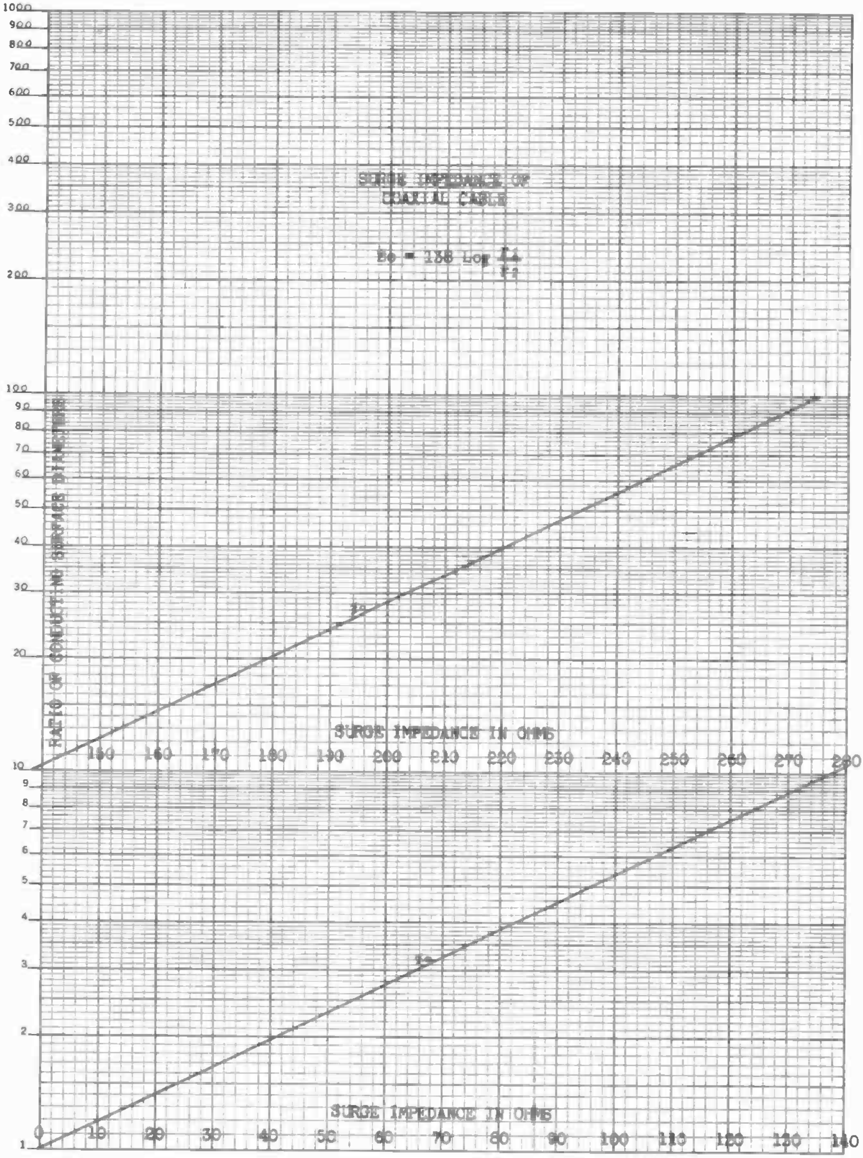
6. Why are concentric lines to be preferred to open wire lines, and what precaution must be taken in the design of concentric line?

7. Mention several mechanical considerations in the erection of television antennas.

8. What three pieces of apparatus are used in measuring the input impedance of a television antenna, and describe, in a general way, how this is done?

9. Why is it desirable to eliminate one sideband of a television transmitter, and which sideband is eliminated?

10. Suppose we wish to receive a picture from two transmitters. Transmitter "A" operates with a power of 1000 watts, an antenna height of 200 feet above average level ground, and the transmitter is six miles away from the receiver. Station "B" on an adjacent channel has an antenna height of 600 feet, a power of 4 kw., and is three miles distant. What is the relative field strength at the receiver, produced by the two transmitters?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COAXIAL CABLE
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RATIO OF CONDUCTING SURFACE DIAMETERS
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SURGE IMPEDANCE IN OHMS
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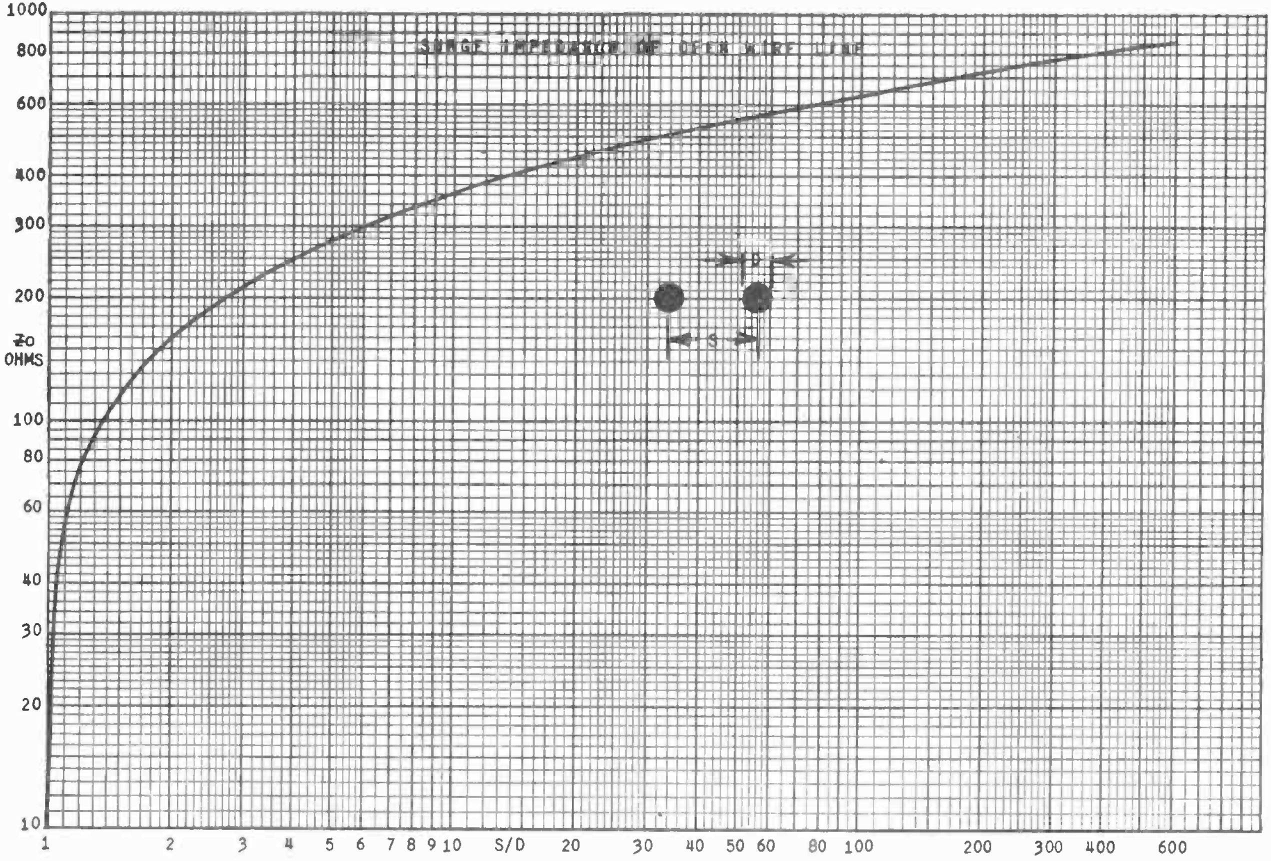

SURGE IMPEDANCE OF OPEN WIRE LINE
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D
S

Power & Light Building
Kansas City, Missouri

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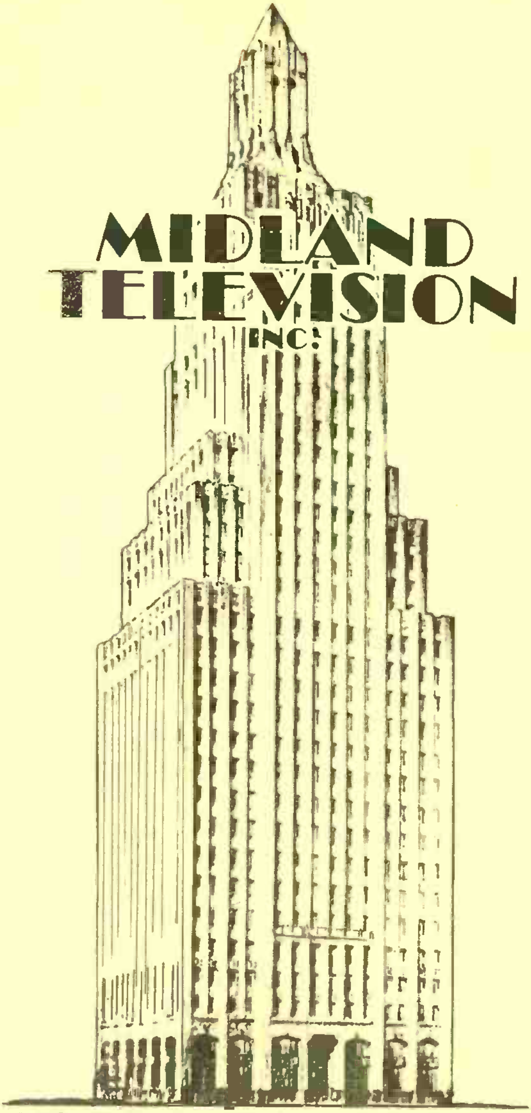

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 7

TELEVISION FILM TRANSMISSION

LESSON NO. 8

TELEVISION

film transmission.

We have written so many stories about the changes that television may bring in our lives, that this page may be, in part at least, a repetition of what you have previously read. However, the possibilities are so fantastic that they will bear repeating.

Today, moving picture films are shipped to all parts of the United States and distributed through film exchanges. Many of the original films must be made so that a considerable number of theatres may screen the play on the same day.

Let us assume that a large moving picture company should construct master television stations at strategic points in the United States. These stations, in turn, would transmit moving pictures to selected theatres within their area. Such theatres, being equipped for the reception and screening of film transmission, would present the play without film. While this may sound fantastic, it is within the realm of future possibility.

There are many other equally fantastic uses for television that may come in the future. When radio broadcasting was first inaugurated, but few people had the slightest conception of the change it would bring about in our lives. Radio was called a toy....a plaything for the man who loved a hobby. Consider what radio means to us today. Then, again, consider what television may mean to us in the future.

We, at Midland, are prepared, awaiting the day.....you, too, should remain PREPARED

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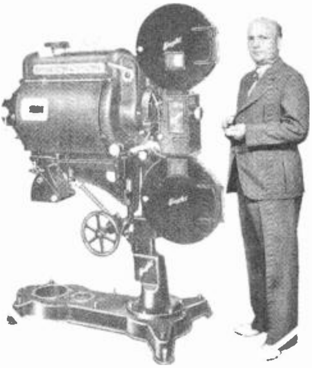
Lesson Eight

TELEVISION FILM TRANSMISSION



"One of the big problems in television is the source of program material. From the experience which has been acquired to date, it is quite evident that motion picture film will be used a great deal.

"This lesson deals with two things every television student should learn; first, the operation of a motion picture projector; and, second, how film is scanned for transmission by television."

1. USE OF MOTION PICTURE FILM FOR TELEVISION. It is generally appreciated that one of the chief features of the modern television system is the simultaneous action of the events being televised and the scenes which are witnessed on the receiver in the home. In witnessing a sports event, a large part of the interest is undeniably in the uncertainty of the outcome. The thrill at the end of a horse race would be lacking if the audience knew in advance which horse was to win. Furthermore, the public wishes to feel that the show is being put on for them purposely at that very instant. These considerations limit the extensive use of film at the present time; nevertheless, many types of shows are produced best by the use of motion picture film.

Most people object to recorded sound programs, even though they must listen carefully to distinguish between a transcribed and a live talent show in modern sound broadcasting. This attitude is probably carried over to some extent in television as a prejudice against the use of film. There is a distinct difference, however, between recorded programs for sound broadcasting and recorded programs for television. In the case of a recorded television program, it is possible to overcome certain limitations of the television studio; thereby producing shows of greater elaboration and higher entertainment value.

For the pickup of news and sports events, the motion picture camera has an advantage over direct pickup television equipment in that it is more easily transported to the scene. Until recently, mobile television equipment could not be placed in an automobile or an airplane and rushed to the scene of interest. Two special trucks were needed for the remote pickup unit. Motion picture cameras, on the other hand, are quickly and easily transported to the scene of a fire, explosion, or other unexpected event of news value. Films can be rushed to the television studio and placed on the air in instances where direct pickup with a television camera is impossible.

By the use of motion picture cameras, events may be recorded in the daytime and placed on the air at night when the largest television audience is available. Likewise, events which occur after midnight may be photographed and placed on the air the following day. A Saturday afternoon football game on the east coast can be held over by means of motion picture film and broadcast Saturday afternoon on the west coast. Even variety shows, dramas, music presentations, and other broadcasts which could easily be in the nature of a direct pickup might be recorded during the daytime and placed on the air at night in order to take full advantage of a limited number of studios and operating personnel. This reduces the cost of programs and allows an advertiser to reach a large nighttime audience without excessive production costs.

The time and space conveniences of motion pictures in television has another aspect, especially in dramatic presentations. Successive scenes may be produced which represent widely varying places and unrelated time. Flashbacks to an earlier period may be portrayed if useful to the program continuity. Changes in the locale, time, and costuming are more convenient. If the story depicts a street scene in the daytime, it need not be re-written for television simply because it is being broadcast at night, as films can be made at the most convenient time. A character in a play can be pictured as stepping on a train in Chicago, and arriving in New York, by actually filming these two locales, with much more realism than could be achieved by representing these places in the studio.

One of the best recognized advantages of motion picture film over direct pickup is the application of splicing technique. A program produced in the studio for direct pickup must be perfect in every detail. Any fault in the program which develops will be viewed by the entire audience. If the program is first filmed, defects can be thrown out of the final version. In practice, every scene is shot with two or more cameras and the scenes subsequently spliced together to register the best effect. For direct pickup, each scene would have to be tried in a variety of ways, the best method selected and then rehearsed repeatedly until it could be reproduced every time without a flaw. Elaborate rehearsals require that many studios be provided and all of them equipped with a multiplicity of cameras and television equipment necessary to audition the performance. A further advantage of film technique is that accurate timing may be achieved by cutting. Scenes may be shortened or lengthened at will and spliced together to obtain a program of split-second length. Accurate timing of a direct pickup, on the other hand, requires endless rehearsing and greater directing skill. Timing of programs in modern broadcasting is extremely important since the networks switch the programs and key in the spot announcements by the second hand on the control room clock, rather than by cues in the program itself as was done at an earlier time. Sound programs occasionally are marred by a station cutting in the middle of an announcer's speech. The difficulty of accurate timing is greatly increased in television.

When the continuous projection type of motion picture machine is used (discussed in detail later in the lesson), the film may be slightly speeded up or slowed down without any disturbance in the performance of the television circuits. This is an advantage in that the continuity can be adjusted to exactly the right length for the time allotted to that portion of the program, which is difficult to do precisely in studio technique. The change in pitch of the sound accompanying the change in film speed is usually the limiting factor, as running the film faster will cause the sound to become higher pitched, or running the film slower will cause the sound to have a lower pitch. Of course, if a change of film speed is carried to extremes, the motion of the picture also will appear unreal.

The use of motion picture film also makes it easy for the sponsor of a program to audition it, not once but many times, by simply running the film through a projector. Advertising matter may be spliced into the film at any point, and it is possible for an agency of sponsored programs to produce a program on film and then sell the program to a sponsor, whose advertising matter is inserted, after the film has been produced.

Syndicated programs will be possible as a preliminary or a supplement to chain broadcasting. If it is not economically feasible to chain a number of small stations by coaxial cable or by a R.F. relay network, a major program can be produced in Hollywood, Chicago, or New York, which may be filmed, printed, and rushed to the syndicated stations so that they can broadcast the program either simultaneously or with due regard to the different time belts in which they are located. This will require a systematic and rapid means of distributing these programs, but the distribution undoubtedly will be worked out in conjunction with commercial airline schedules.

A television program does not have to be exclusively direct pickup, or exclusively film. A dramatic production might be largely produced in a studio, but supplemented by shots taken at other times and other places and recorded on film. One scene could be a studio shot followed by a brief motion picture insert without discontinuity in the program. Special effects can be recorded on film and used as program fills. This has the advantage of reducing elaboration of the studio.

Both still slides and motion picture films are used in projecting backgrounds against which the players perform in the studio. At present, this technique is limited because of the extremely high intensity of illumination which must be projected from behind the translucent back-drop if the background is to stand out visibly in the brightly lighted studio.

The use of motion picture film in conjunction with a television pickup tube allows more light to fall on the tube than in the case of direct pickup. This makes possible several techniques for film broadcasting. Insufficient light is the chief limitation in the use of certain types of television pickup tubes. You will recall that it was this factor which prevented the use of a scanning disc and a photoelectric cell in a high fidelity mechanical television system. By projecting a large amount of light through the film, the Farnsworth image dissector tube (which you studied in an earlier lesson) will function very well and is capable of excellent resolution with entire freedom from blackspot. In

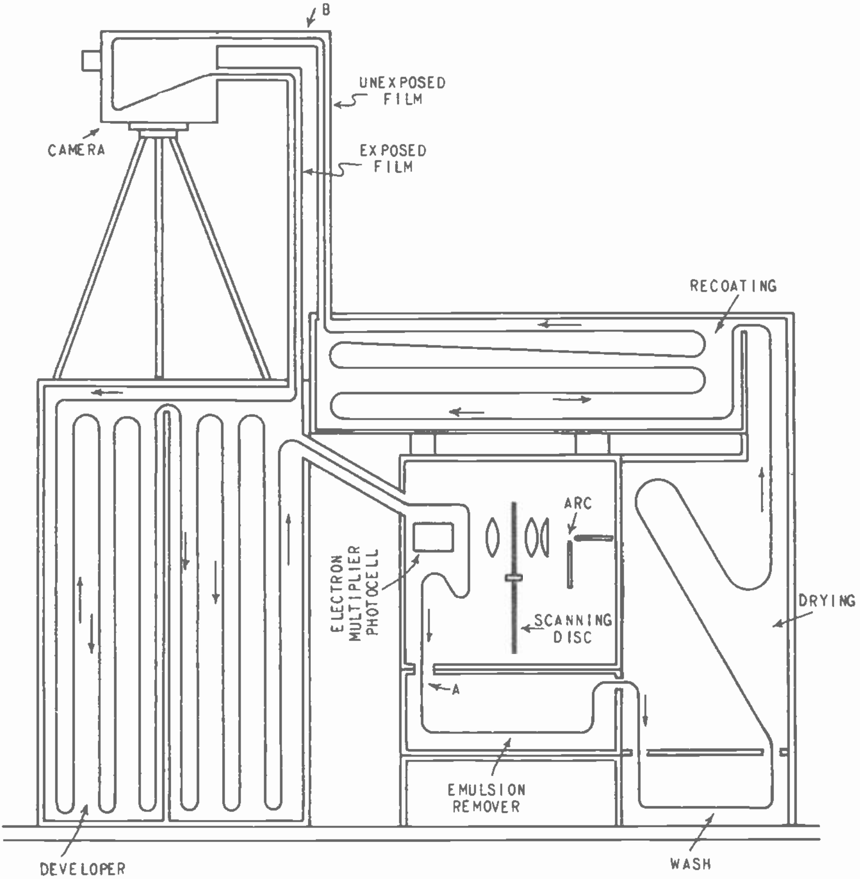


Fig.1 Intermediate film method of pickup.

systems using direct pickup, the image dissector gives insufficient signal output to override tube noise in the amplifiers unless the scanning aperture is made so large that the resolving power of the system suffers. With motion picture film, even mechanical scanning drums are possible. In Europe, some television systems have taken advantage of the excellent pictures obtainable from motion picture film and the image dissector tube, by incorporating a modified direct pickup system known as the intermediate film process.

A sketch of a continuous intermediate film scanner is shown in Fig. 1. The scene to be televised is photographed in the usual manner. The film passes through a developer; and then, while it is still wet, it is scanned with a mechanical scanning disc. The disc is fitted with lenses, and rotates at high speed in a partial vacuum, which is necessary to reduce air friction losses. Unlike the type of mechanical television studied in an earlier lesson, all of the holes in the scanning disc lie in the same radius, and the scanning disc revolves several times for every frame of the picture. The vertical motion of the picture is achieved by the continuous motion of the film. Because of the intensity of the arc illumination and the large amount of light passed by the lens disc, 400-line mechanical television is practical.

There is a delay of approximately 30 seconds from the time the picture is photographed until it is electrically reproduced in an electron multiplier photocell pickup. In the sketch shown, the film, after passing through the scanning apparatus, is salvaged and re-coated with a photographic emulsion. Thus, the band of film passing through the film scanner is continuous. However, this need not be the case. In some models, the film path between points A and B is omitted, being replaced by a takeup reel and a new spool of film at points A and B respectively.

When properly constructed, an intermediate film scanner is capable of excellent pictures, but the equipment is unwieldy, and, because of the half-minute or so delay in transmission, the sound must also be recorded and delayed an equivalent amount.

2. BASIC PRINCIPLES OF MOTION PICTURE PROJECTION. A motion picture projector flashes on the screen a series of still pictures, each of which has been photographed with a moving object in a slightly different position. These pictures occur in rapid succession and the audience has an illusion of motion. Because of the persistence of vision, successive still pictures occurring in rapid succession cannot be individually distinguished.

The motion picture industry has standardized on the number of pictures which must be shown per second in order to take full advantage of the persistence of vision in eliminating flicker. Twenty-four pictures are flashed on the screen per second, but each of these is shown twice, so that effectively 48 pictures are shown on the screen in one second. The film which is projected carries not only the picture, but also a recording of the sound which accompanies the picture. The dimensions of the film, the width of the sound track, and the dimensions of the individual

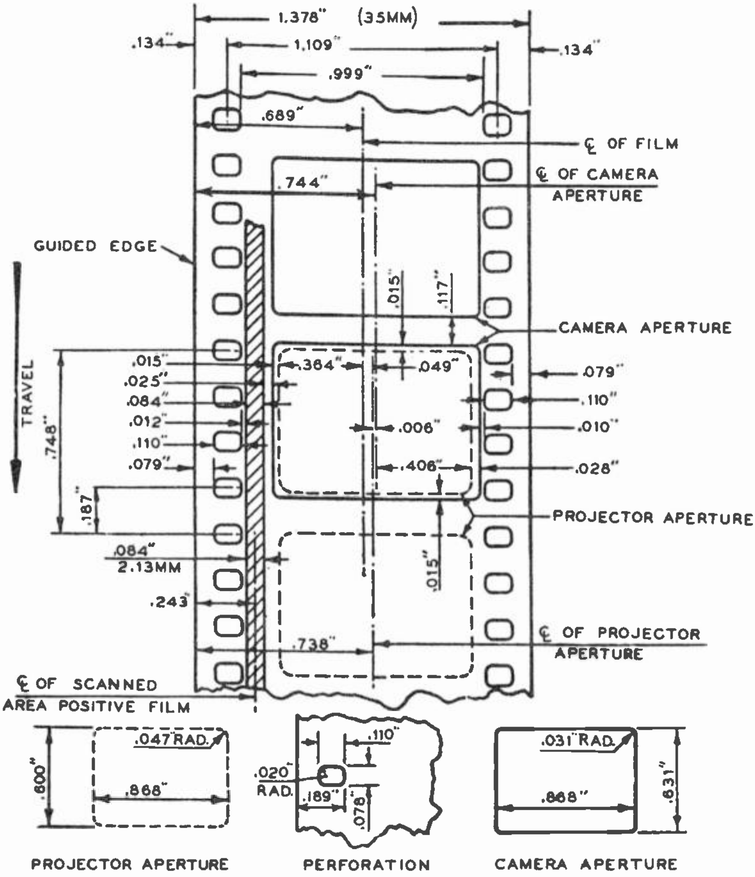


Fig.2 Standard dimensions of 35 mm. film

pictures on the film, are shown in Fig. 2. From these dimensions, the film travel can be calculated at approximately 90 feet per minute when 24 frames per second are being projected. Fig. 2 is interesting in that the small size of the projector aperture indicates the tremendous magnification necessary to project a 12 x 16 foot picture in a theatre. In order to project such a large picture without blurring the image, it is necessary that the film dimensions be held very accurately and that the projector be capable of pulling down successive frames of the picture to exactly the same position in the projector. Misalignment of successive frames by as much as .001 inch will spoil the resolution of the projected image.

A schematic diagram of a motion picture projector is shown in Fig. 3. This diagram shows typical parts in a projector, and the

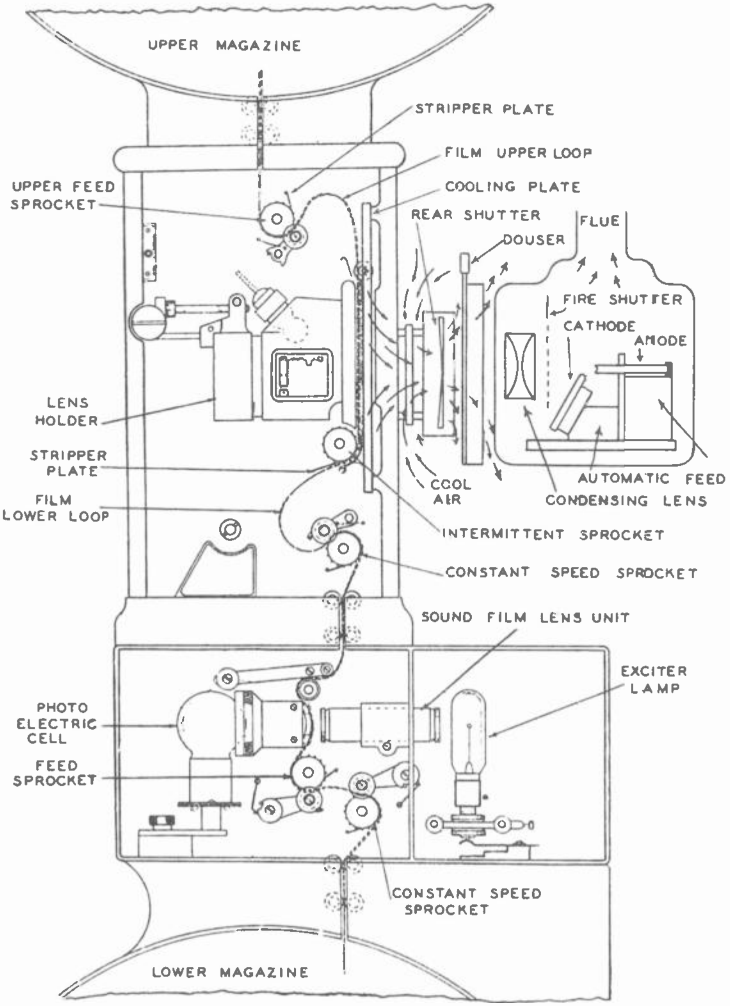


Fig.3 Component parts of a motion picture projector.

functions of these parts are more or less indicated by their names and positions. The image of the arc crater is focused to a point just beyond the film gate so that the film gate accepts the center portion of the cone of light. Some light is allowed to fall on the cooling plate which contains the film gate. The light rays cross over and enter the projection lens after passing through the film. The upper magazine contains the film which has not yet been projected. This is pulled down at a steady rate of 90 feet per minute by the upper feed sprocket. The film roller just below the magazine, and the idling roller which runs against the feed sprocket, serve to steady the film. These rollers also help to prevent the spread of fire to the upper magazine, should the film catch on fire at the gate. A loop of free film is left between the upper feed sprocket and the film gate, because the intermittent sprocket pulls the film down, a frame at a time, with a jerking motion; whereas, the upper feed sprocket rotates continuously. There is a slight drag upon the film reel in the upper magazine which prevents the film from unwinding too rapidly and becoming entangled.

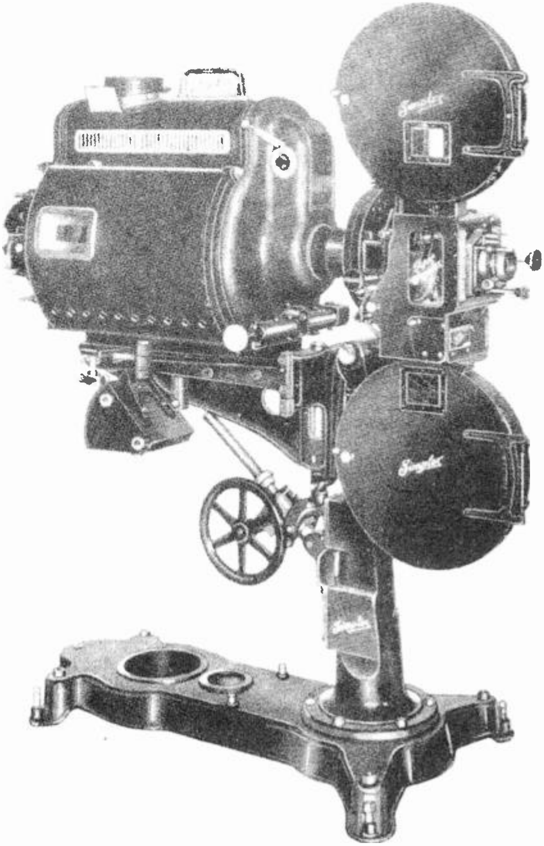


Fig.4 Simplex Super Magnarc Projector.

After the film has passed over the intermittent sprocket, a second loop is allowed to take up the intermittent film motion, since the lower sprocket and its idler once more transmit a steady motion to the film, so that it will pass the sound track at a uniform rate of speed. The uniformity of motion is further insured by an additional guide roller and a feed sprocket in the sound head. Here the sound track on the film is scanned with a thin sliver of light. The film modulates the light beam and the variations of intensity are converted to an A.F. output by a photoelectric cell. The exciter lamp is supplied with DC current.

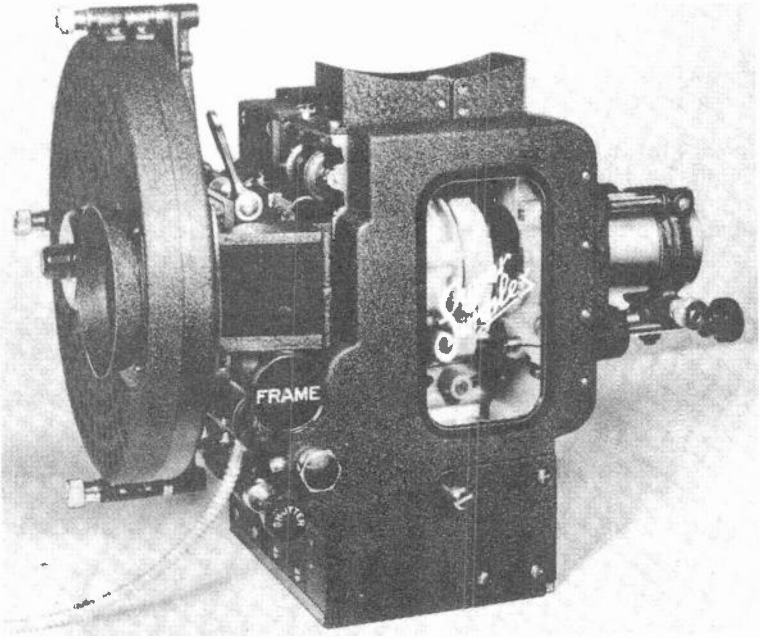


Fig.5 Mechanism of Simplex projector.

The lower magazine contains a takeup reel which is driven by a loose-fitting belt. This belt tends to rotate the takeup reel faster than it should go. Rapid rotation is prevented by the tension of the film so that the belt driving the takeup reel continually slips. The belt tension is adjusted so that the lower takeup reel will just barely turn when the reel is full of projected film.

The stripper plates indicated at the upper and lower feed sprockets prevent the film from clinging to these sprockets. They are necessary, because of the loose film loops which are provided to compensate for the intermittent film motion between the two feed sprockets.

A rotating shutter is introduced between the arc lamp and the film gate which shuts off the light once while the film is being moved to a new frame and once at the half-way interval during the projection of each frame. This serves to eliminate flicker and prevents film motion from blurring the image while the film is being transported from one frame to another. An additional purpose of the rotating shutter is to cause a draft of air past the film gate, keeping the film sufficiently cool at this point, and

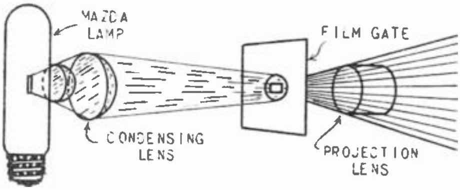
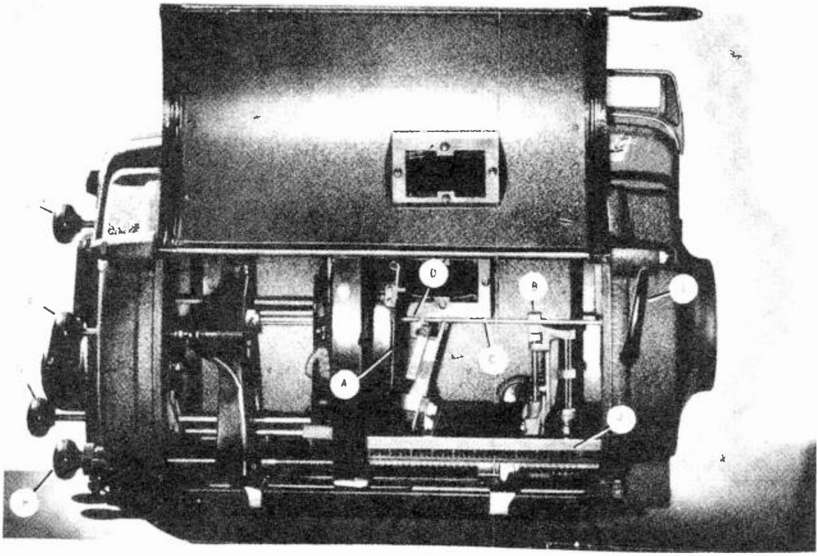


Fig.6 Formation of cone of light from a Mazda projection lamp.

also ventilating the lamp house. The douser serves to shut off the light at the beginning and end of each reel.

We shall now describe in detail the individual parts of the projector, but first look at Figs. 4 and 5, which show a Simplex



(B) Mirror Reflector. (B) Positive Carbon Jaw. (C) Positive Carbon. (D) Negative Carbon. (E) Mirror Vertical Adjustment. (F) Arc Striking Knob. (G) Mirror Horizontal Adjustment. (H) Focus Control. (I) Douser and Flame Shield Handle. (J) Positive Carbon Feed Scale.

Fig.7 Cross-section of typical lamphouse.

Super Magnarc projector and a closeup of the Super Simplex projector mechanism which is used on this projector. These illustrations are of typical theatre projection equipment, although, of course, many other models are in use. A Simplex Model E7 projector, which is very popular, is shown later in the lesson, (Fig. 25), after it has been altered for television purposes.

The light source of motion picture projectors may be either an incandescent lamp, a low or high intensity arc lamp, or an AC arc. The latter is not practical for television purposes, and will not be described. Fig. 6 indicates that a Mazda projection lamp (which burns approximately 50 to 100 hours) is used with condensing lenses and without a reflector in the lamphouse. The lamphouse for such a light source does not have to be ventilated with an external flue; this is a considerable advantage.

Fig. 7 shows the construction of one type of arc lamp. The positive carbon is held in a clamp in front of a mirror reflector. The negative carbon projects through a hole in the center of the mirror. The mirror concentrates the light from the positive carbon onto the film aperture and cooling plate. Since no condensing lenses are used in this type of lamp, a great deal of heat reaches the film, but this is not serious in a low current arc. Both the positive and negative carbons are arranged with automatic feeds so that they are advanced at the proper rate, as the carbons are burned up. Since the mirror forms an image of the crater of the positive carbon, the carbons are completely adjustable in position to take full advantage in focusing the arc through the film.

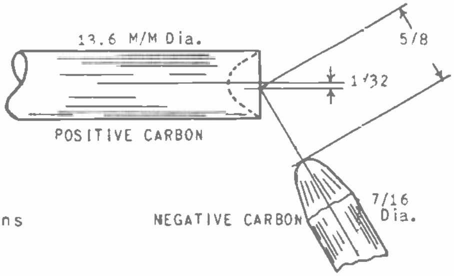


Fig.8 Carbon positions in a DC arc.

The positive carbon is formed by extruding a carbon rod with a hole in the center, and baking this rod. The hole is afterward filled with soft carbon in the form of a core. The core material burns out first and then the arc which is formed between the two carbons centers on the positive crater and prevents the arc from wandering around the face of the carbon. The negative carbon is somewhat smaller in diameter and is copper plated to increase its conductivity and to prevent it from spindling, which means burning out to a long thin point. The steadiness of the flame is enhanced in some models by the use of a magnet to direct the arc flame to the proper position. Fig. 8 shows the proper alignment of the carbons in a low intensity arc.

The arc light is produced by the carbon vapor which is formed at a temperature of about 3600 degrees Centigrade. This temperature limits the whiteness of the light which can be achieved in a low intensity arc, because increasing the arc current simply volatizes more carbon without affecting the temperature. The carbon gases must be carried off in a flue, and this is aided in most projectors by a fan incorporated with the rear shutter.

In high intensity arcs, the central core of the positive carbon is made larger and mixed with certain cerium compounds and other rare earths. These materials light up under the influence of electron bombardment in the arc. Furthermore, the deep crater formed in the positive core builds up a certain amount of gas pressure which increases the temperature of the arc. As a matter of fact, the crater of this type of arc lamp has about the highest temperature of anything known except the sun. The positive carbon is rotated as well as advanced, as it burns away, to maintain a steady flame. A picture of the flame from a high intensity arc is shown in Fig. 9. The tail flame, as it is called, serves no useful purpose, but a condensing lens system focuses the brilliant crater of the positive carbon to a point through and beyond the film aperture.

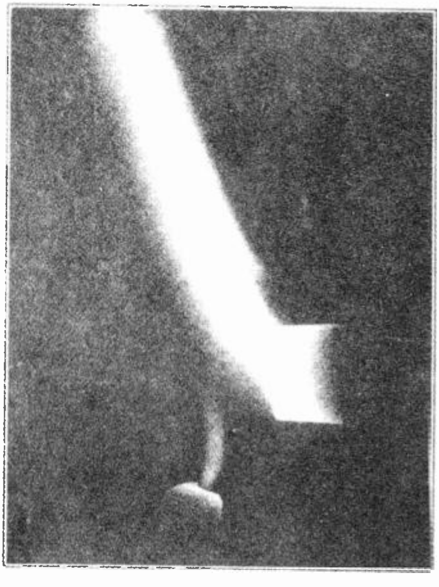


Fig.9 High-intensity arc.

In order to gather as much light as possible, the condensing lenses are placed close to the arc. However, they cannot be placed too close, since intense heat will crack them. A schematic diagram of the condensing lens system used with the high intensity arc is shown in Fig. 10. Facing the curved portions of the lenses together reduces the amount of light lost by reflection in the lenses and reduces spherical aberration.

The film path in the Model E7 Simplex projector is shown in Fig. 11A. The film comes from the upper magazine through a fire trap and around the upper drive sprocket A. A loop of film, as previously described, is allowed between this sprocket and the film gate B. The intermittent sprocket C, is driven by the intermittent movement which will be shown later in detail. Two idlers help to maintain steady motion of the film over the lower driving sprocket D. A little slack is left in the film between this sprocket and the rotary stabilizer E.

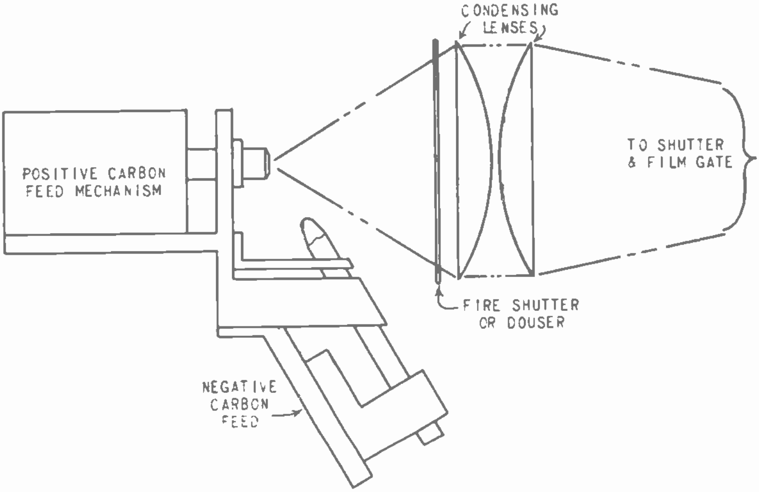


Fig.10 Condensing lens system for a carbon arc.

The rotary stabilizer is a large flywheel connected to a film sprocket by a film of oil. This flywheel is driven by, and, in turn, drives the film, insuring absolutely constant motion of the film in passing through the sound head. Any irregularity in motion at this point carried over from the intermittent motion of the film would cause the sound to flutter or introduce "wows". The path of the sound track is not clearly shown in this photograph, but will be discussed later in detail.

Fig. 11B shows a closeup view of the upper drive sprocket and the automatic fire shutter safety trip. Details of this safety trip vary with most projectors, but it is arranged to operate a free falling fire shutter to cut off the illumination if the film accidentally remains stationary in the film gate or if the film drops below a speed of 60 feet per minute. The fire shutter remains closed until the film has reached its normal speed.

Fig. 12 shows the construction of the film gate and the means for maintaining absolute accuracy in the position of the film. Fig. 12A shows the guide rollers and the accurately machined guides, (which should be replaced if slightly worn), for maintaining the correct position of the film. The film is held against the guide by long, light-pressure spring clips shown in Fig. 12B.

The guide unit (Fig. 12A), and the film gate (Fig. 12B), mount face to face in the projector.

The rear shutter shown in Fig. 13 is equipped with curved fins which cause an air stream to pass away from the film aperture and back through the lamphouse, thereby keeping the film cool.

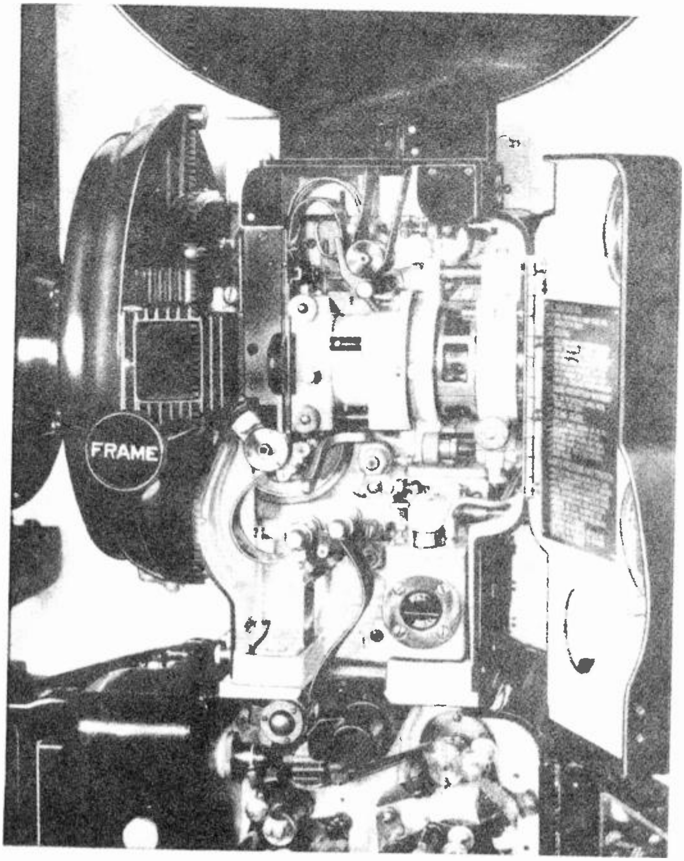


Fig.11A Film path in a Simplex E7 Projector.

Perhaps the most delicate part of the entire projector is the intermittent, which causes the film to be pulled down one frame at a time. This must move successively in steps into a new position with an accuracy of .0001 inch. This means that the individual parts of the intermittent must be machined to a closer tolerance than this. Fig. 14 is an illustration of the intermittent operation. The star, labeled A, is directly connected to the intermittent sprocket. While the cam (B) is rotating for three-quarters of one complete revolution, the star is locked rigidly in place by the curved track labeled C on the cam. In the remaining one-quarter revolution, the pin P engages the slot in the star and causes it to rotate. The star begins to rotate slowly, comes up to maximum speed, in the position shown in Fig. 14B, and the rotation is then decelerated as the pin P leaves the groove or slot in the star. The star is now locked rigidly in place while a second frame of the motion picture film is being rotated. Thus, for each complete revolution of the cam, the film is being transported for one-fourth of the time and held steadily in front of the film gate for the remaining three-quarters of the period.

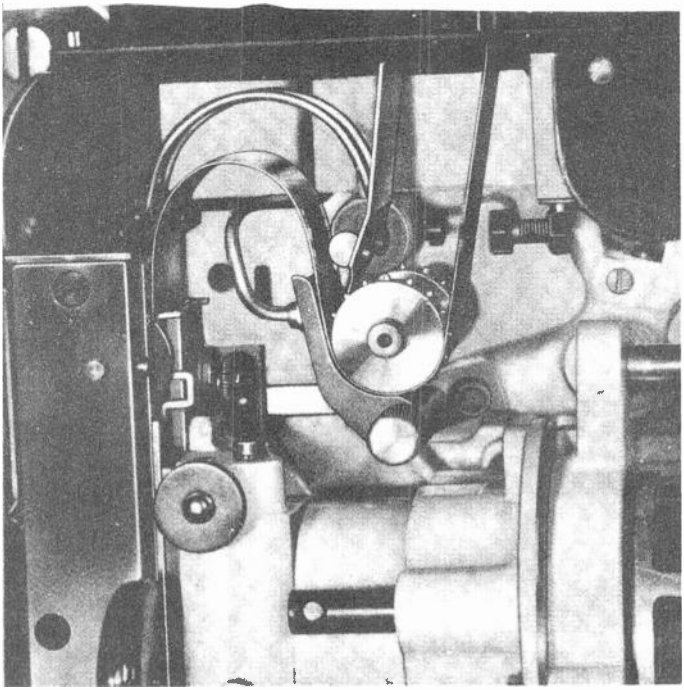


Fig.11B Automatic fire shutter safety trip and upper drive sprocket.

The smooth motion of the intermittent in starting and stopping the film reduces wear and tear on the film, and the accurate locking of the intermittent during projection increases the steadiness of the picture. Fig. 15 is a cutaway view of the intermittent mechanism during the interval of moving the film from one position to another.

A film passes through the sound head over several constant speed sprockets equipped with idlers, and steady motion of the film at the point of scanning is insured by a rotary stabilizer previously described. A generous use of flywheels connected to the constant speed sprockets is sometimes substituted for the rotary stabilizer.

Schematic diagrams of two types of scanners are shown in Fig. 16. In the first type, an image of an air slit approximately .001 inch wide is focused on the sound track of the film. Light passing through the film is picked up by a photoelectric cell. In Fig. 16B, an enlarged optical image of a portion of the sound track is focused on a shield containing an adjustable slit. This has the advantage that the *effective* width of the slit can be greatly reduced with reasonable mechanical dimensions. Fig. 17 is a pictorial representation of the operation of Fig. 16B.

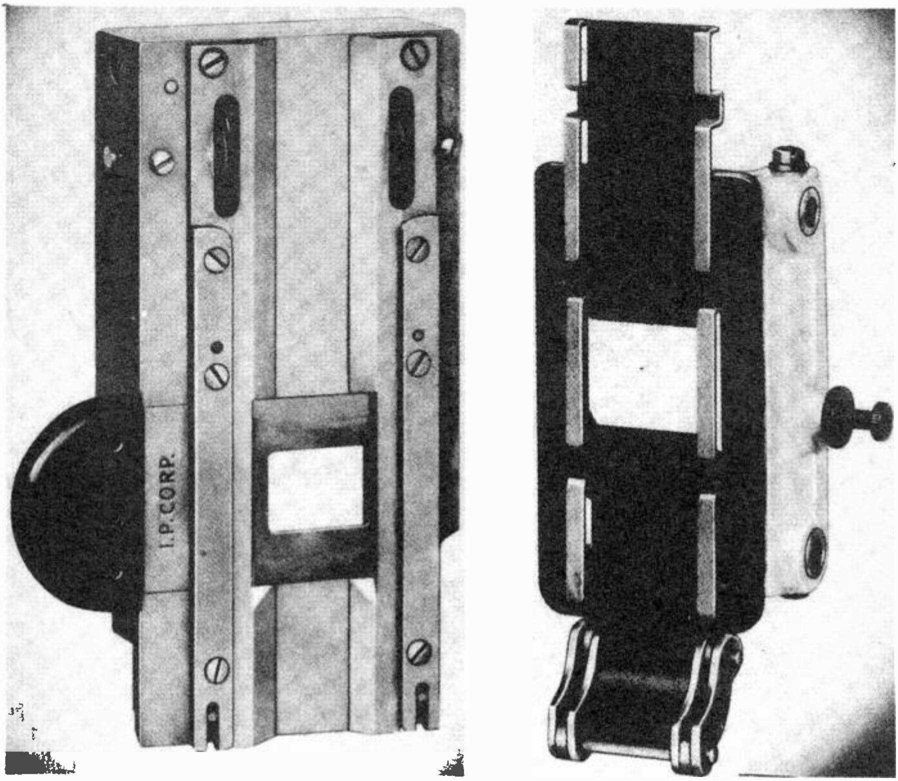


Fig.12 Film gate and film guide.

When the sound is recorded in push-pull, two film tracks are employed side by side. To reproduce sound from this type of film, a double photocell is employed and connected to a push-pull amplifier. The light beam is split and passed through the two sound tracks either by using a split mask or a system of prisms.

The sound is recorded upon the film in several ways. The density of the negative can be varied according to the audio frequency amplitude present in the recorder, or the width of the sound track can be varied. The Western Electric system of recording uses variable density, while the RCA system of recording uses variable width sound track. To reduce the noise due to dirt and scratches on the film in the variable width method, a portion of the sound track which is unused for low intensity sound is blocked out in the film. This has the effect of adding a very low frequency component on the order of 3 or 4 cycles, but it is not objectionable because the audio amplifier in the reproducing system will not pass these low frequencies.

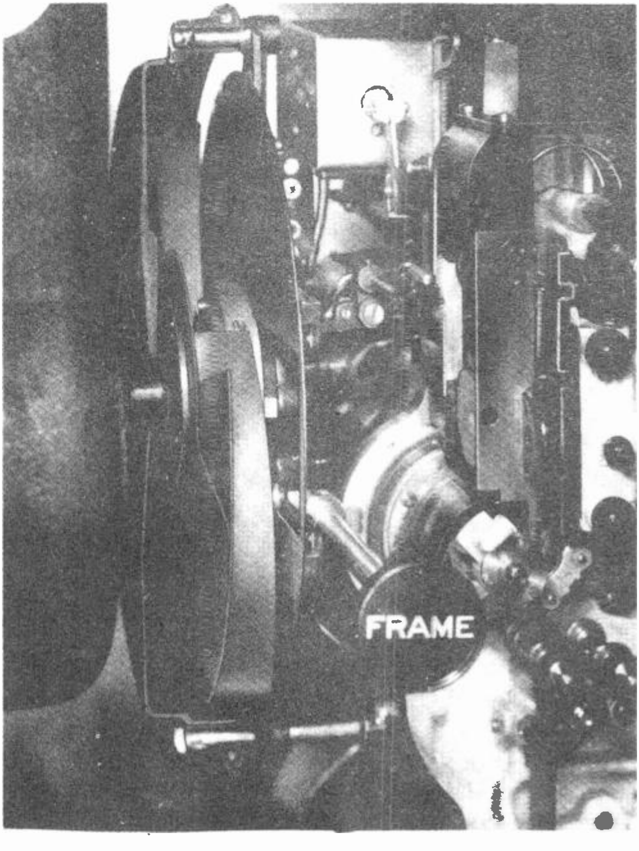


Fig.13 Rear shutter in a Simplex E7 Projector.

Either the variable density or the variable width method may be employed in push-pull; that is, the sound track is split in half and two recordings are made. In the case of variable density recording, as the width of the line is increased in one half of the sound track, it is diminished in the other. In the case of variable density recording, a dark portion of one sound track will lie opposite a corresponding light portion of the other sound track. Push-pull recording in general reduces noise due to film scratches and the like, and reduces the amplitude distortion in the photoelectric cells and audio amplifiers.

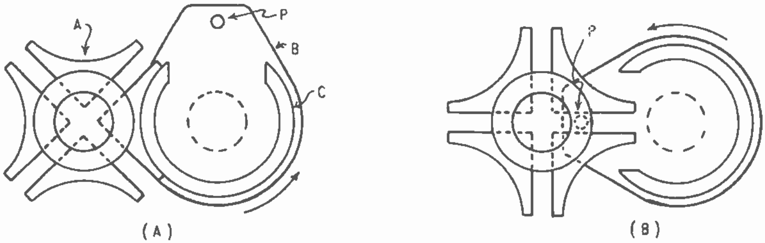


Fig.14 Intermittent star and cam.

Fig. 18 shows a photograph of a variable density recording; a variable width recording; and a variable width recording where the unused portions of the sound track are blanked out to reduce noise during the low passages.

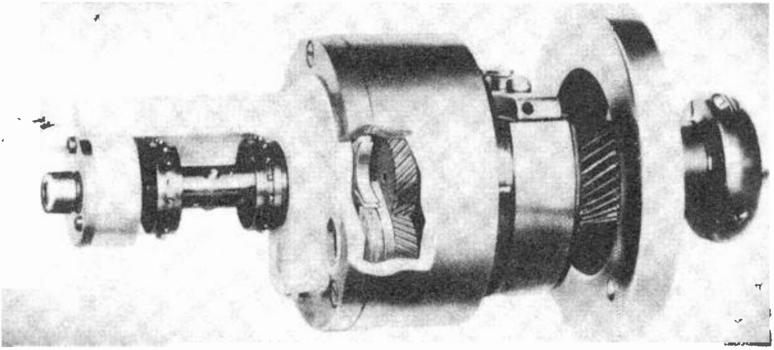


Fig.15 Cut-away view of intermittent mechanism.

3. OPERATION AND MAINTENANCE OF PROJECTION EQUIPMENT. The first step in operating the motion picture projector is to thread the machine. The reel itself is placed in the upper magazine and if it has been correctly spooled, the film, in passing through the projector, will be upside down with the emulsion side of the film toward the lamphouse. Looking from the lamphouse or toward the emulsion side of the film, the sound track will be on the right hand side of the film. In threading the machine, it is necessary to leave a loop of film between the upper drive sprocket and the aperture plate. This is necessary because the film passes through the aperture plate in jerks, one frame at a time; whereas, the upper drive sprocket advances the film continuously. The upper loop must be just large enough to take up the differential motion

of the film. The size of the loop is usually determined by the experience of the operator.

To find out just how much loop should be allowed, it is necessary to thread the machine and rotate the flywheel which causes the film to move in its normal fashion, but at greatly reduced speed. The intermittent motion of the film is absorbed by the loop, and the size of the loop is diminished until most of the slack is taken up as the frame of the picture is advanced. Some operators prefer to allow rather large loops, but they should, of course, not be made so large that they would touch the upper or lower fire shields. Between the intermittent and the lower constant speed drive sprocket, another loop of film must be allowed in order to translate the intermittent motion of the film back to a constant speed of travel. The statements applying to the minimum and maximum size of the upper film loop apply equally well to the lower film loop. Between the lowest sprocket of the projector mechanism and the constant speed sprocket in the sound head, a small amount of play is allowed in the film.

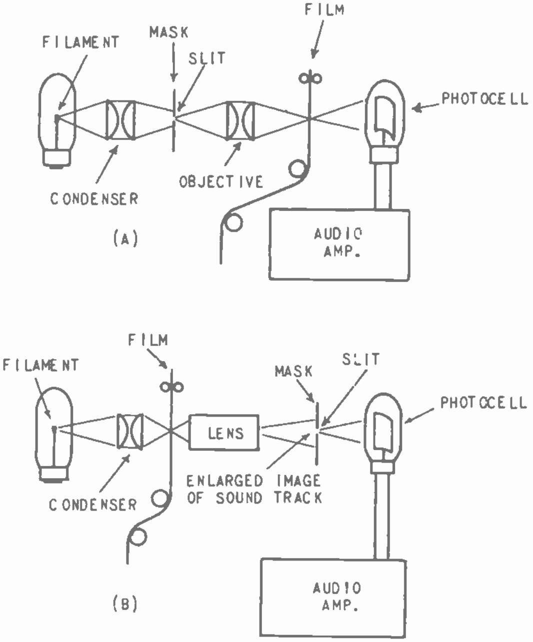


Fig.16 Alternative systems of sound track scanning.

For perfect synchronization between the picture and the accompanying sound, the frame which is threaded in the sound head must be exactly 20 frames ahead of the frame occupying the projection aperture. This location is not especially critical and one or two frames one way or the other is difficult to notice in the sound. It simply gives the illusion that the observer is sitting a little closer or a little further away from the scene than he actually is. In order to get perfect synchronization, most standard films have a diamond mark on the film leader, 20 frames ahead of the start frame. If the projector is correctly threaded, when the start frame is over the projection aperture, the diamond will lie opposite the scanning aperture in the sound head. In the case of sound on discs, if the starting frame is over the projection aperture, the needle on the sound disc should lie opposite the synchronization mark.

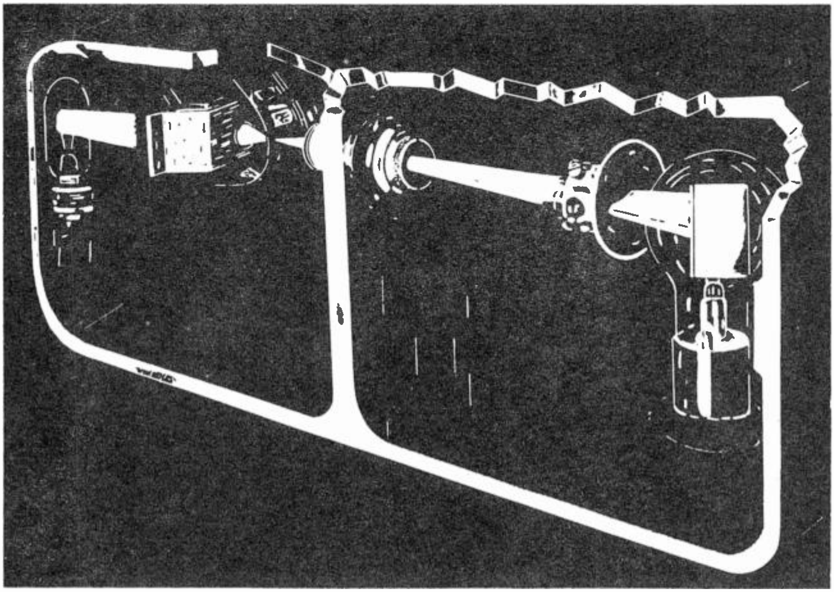


Fig.17 Pictorial sketch of sound track scanning.

After the film has been threaded, the douser should be closed while striking the arc so that particles of hot carbon will not damage the condensing lenses. The carbons are advanced manually until the arc is struck, and they are then moved back immediately to a normal operating position. A little experience will enable the operator to determine just how much spacing between the carbons should be allowed on a particular type of arc. When the arc is running, the motion picture operator starts the machine on a cue a few seconds before time for the film to flash on the screen. The fire shutter should open automatically when the projector comes up to speed, and as soon as the starting frame of the film passes the projector aperture, the douser is raised and the picture is on the screen. In case the showing is a test, the projection lens is focused and the position of the arc carbons are adjusted for maximum resolution and brilliance of the picture on the screen. These adjustments need not be made during the actual performance.

Films are received from the film exchange in 1,000 foot reels. During the inspection by the projectionist, these films are normally spliced into 2,000 foot lengths and re-wound on the 2,000 foot reels which are part of the projection equipment. If the motion picture consists of more than one reel; and any feature length picture consists of several reels; the second projector is loaded and the arc started. When the reel of the first projector is nearly exhausted, the operator must be on the alert for the motor start cue. This cue consists of a black dot in the upper right hand corner of the picture. If this portion of the picture represents a dark scene so that the spot would not be visible, or difficult to distinguish, the dot will be surrounded with a thin white line. The dot lasts for four frames and is located four feet and six frames from the end of the reel. This allows the projectionist eight seconds to start projector number two and bring it up to speed. As projector number two comes up to speed, the projectionist must now watch for the changeover cue. This is a black dot similar to the motor start cue, but it is located 22 frames from the end of the reel. This allows the operator approximately one-half second to make the changeover from one projector to the other. This is done by raising the douser on projector number 2 and dropping the douser on projector number 1. The sound

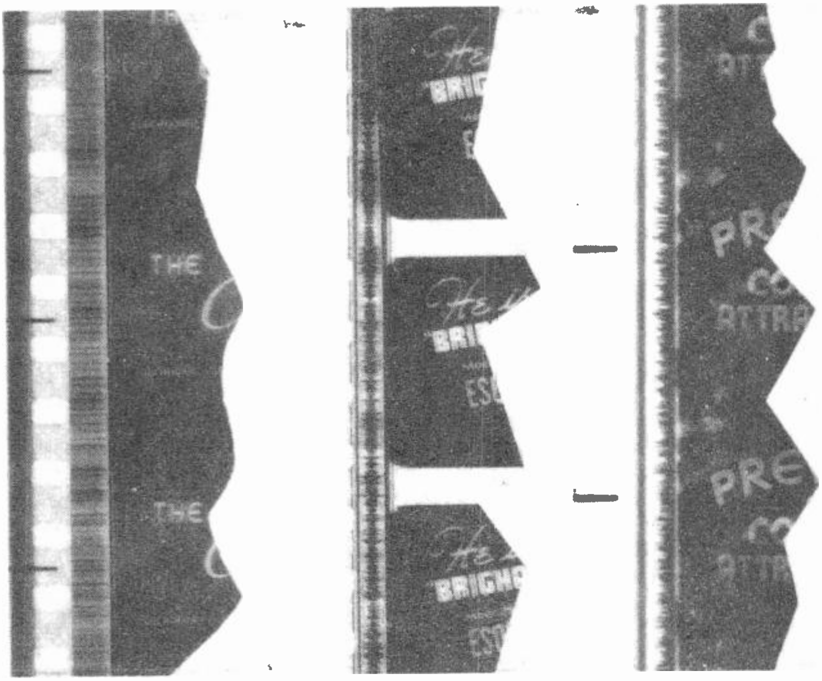


Fig.18 Sound tracks.

must be switched from projector 1 to projector 2 at the same time.

As soon as a reel has been finished and removed from the machine, it is immediately rewound before being placed in the film storage cabinet. Film should not be out of this cabinet longer than is absolutely necessary and no reels should ever be removed unless they are being inspected, projected, or rewound.

The film exchanges are responsible for supplying the projectionist with film in good condition, clean and ready for showing. However, the projectionist must inspect each reel of film before loading the machine with it. This is done by unwinding it and passing the film over a ground glass with incandescent illumination from the underside.

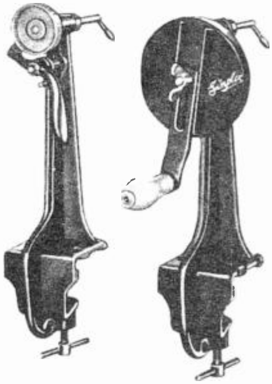


Fig.19 Hand re-winder.

Motor operated rewinds are to be preferred since they can be set up to rewind the film at a slow rate of speed. If hand rewinders are used, the general tendency is for the operator to rewind the film too rapidly, which puts unnecessary stress on the film, particularly if the rewinding spindles are not perfectly in line. A simple hand rewinder is shown in Fig. 19. The film should be rewound at the rate of 1,000 feet in six minutes. In inspecting the film on the rewinder, all splices should be inspected to see if they are tight; not too bulky to pass over the sprockets in good shape; and that the sprocket holes are correctly aligned. Portions of the film containing broken sprocket holes or tears in the film should be cut out and the remainder of the film spliced together. Oil or dirt on the film should be carefully removed, since this spoils resolution and introduces noise into the sound.

If excellent motion pictures are required, the projector must have considerable care. The tension in the upper magazine and the tension of the belt driving the takeup reel must be frequently inspected to make sure that the film is operated at just enough tension to prevent it from buckling up, but not enough to cause excessive strain on the film. The path of the film through the projector should be inspected every day to make sure there is no accumulation of emulsion from the film on various sprockets or idling rollers, and particularly in the film gate. A collection of dirt or emulsion will cause scratches on the film or may even cause the film to jump the sprocket teeth and become snarled. Should this happen, the fault not only destroys the continuity of the film, but there is considerable fire hazard.

All moving parts should be oiled frequently and should be inspected to make sure that all oil is wiped off so that no oil appears anywhere, which could get on the film or lenses. All lenses should be cleaned frequently with a mixture of half grain alcohol and water, and the lenses should be completely removed occasionally and cleaned with soap and water. Lenses should never be cleaned with a handkerchief or other type of hard cloth, since optical glass is not particularly hard and tiny scratches will destroy the quality of the picture. The condensing lens or the reflecting mirror in the lamphouse should be cleaned thoroughly in order to get the maximum amount of light on the screen with the minimum amount of heating in the projector.

All of the gears, sprockets, and other parts of the projector associated with the film motion have very close mechanical tolerances and must be frequently inspected. The sprocket teeth, particularly on the intermittent sprocket, are subject to wear in the form of under-cutting the teeth so that the film sticks to them. When such a condition occurs, the sprockets should be replaced. The tension of the idlers should be occasionally inspected. Very exact adjustment is important here and also in the guiding plates of the film gate.

Almost all standard 35 mm. films are made on nitro-cellulose stock, which is highly inflammable. Some productions are made on cellulose acetate or slow burning film. Almost all 16 mm. films are produced on cellulose acetate film. This is known as "safety film", although such film is not non-inflammable, as some people suppose. It is simply harder to catch on fire and requires a higher temperature for combustion. Film fires are dangerous, and should be avoided with every possible precaution. The automatic fire shutter in the projector should be inspected regularly to make sure that it does not stick and works properly. Film fires are almost impossible to extinguish with any normal type of fire extinguisher because no oxygen is needed to support combustion. Rollers are provided in the projector to prevent a possible fire from spreading to the film magazine, but these precautions are not always satisfactory. Film fires are not infrequent, especially in old projectors, operated by careless projectionists. Film may burn either with or without a flame, but in either case, clouds of toxic fumes are poured forth. The fumes from burning film are very dangerous and cannot be breathed for more than a few seconds without serious results. Sometimes the effects may not be noted for several hours, or even for a day or more. It is thought that part of the disastrous effect is due to formation of nitric acid in the lungs.

The only really successful method of stopping a film fire is to pour cold water on the film until the temperature of the film is reduced below the point necessary to produce combustion. Naturally, smoking is prohibited in any projection room. Film should be stored in such a way that either every reel has its separate metal container, or else a metal fire baffle should be placed between every reel of film. The film cabinet should be equipped with automatic fire plugs which release a stream of water on the film when it rises above a certain pre-set temperature. Water should also be directed on the other film containers in the storage cabinet so they will not catch fire due to the increased temperature in the cabinet.

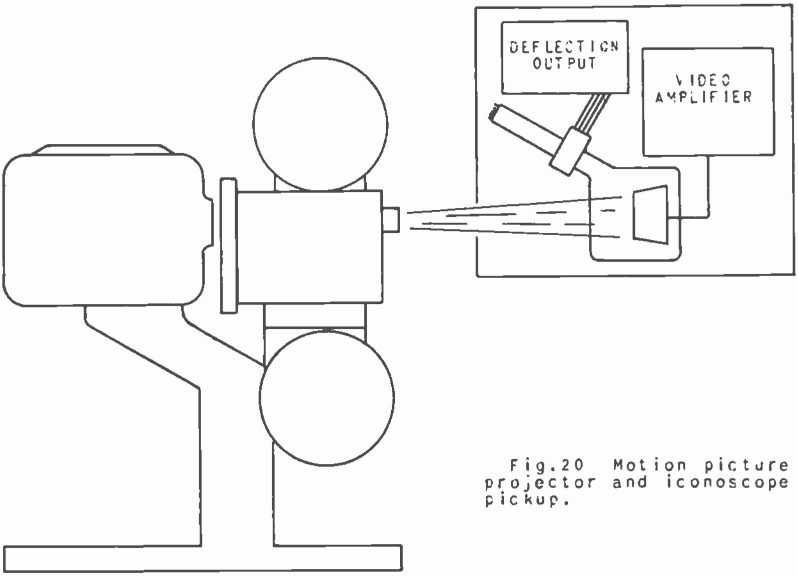


Fig.20 Motion picture projector and iconoscope pickup.

The film storage cabinet has another important function beside fire protection. All films normally contain a certain amount of moisture which is rapidly driven off when subjected to the heat of the arc lamp in the projector. Dried out film becomes very brittle and breaks easily in the machine. This can be prevented if the film is properly stored in a film cabinet which is provided with sufficient humidifiers. The humidifiers are trays of water having a large surface area and arranged so that they will be automatically filled when the water drops below a certain level, thus guarding against the forgetfulness of the projectionist. The humid atmosphere in the film storage cabinet serves to replenish the moisture in the film after it has been partially driven out of the film in the projector.

4. ADAPTATION OF STANDARD FILMS FOR TELEVISION. Standard sound movies run at the rate of 24 frames per second, whereas

television pictures are transmitted at the rate of 30 frames per second. For this reason, standard motion picture projectors can not be used for television. If a projector is set up as shown in Fig. 20 to project a motion picture film onto the plate of a standard iconoscope or other type of pickup tube, the motion picture machine and the iconoscope will not be framed at the same time. If one frame from the projector is projected on the signal plate of the pickup tube, the tube will scan the frame from top to bottom and start to scan the frame a second time before the projector changes frames.

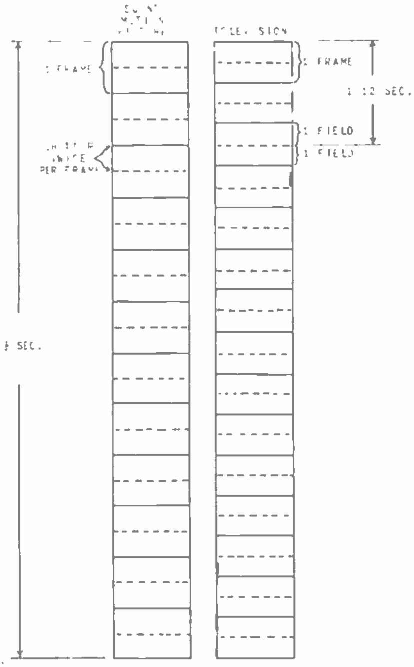


Fig.21 A comparison between the projection rate of standard motion pictures and the scanning rate of television.

Each frame of the picture projected on the signal plate of the pickup tube is interrupted once by the shutter. Consequently, 48 distinct pictures are flashed on the pickup tube per second. The pickup tube scans these at a rate of 60 fields per second. The signal is blanked out by the pickup tube once every sixtieth of a second, whereas the picture in the projector is shut off once every forth-eighth of a second. A beat will be formed between the 48 fields of the projector and the 60 fields of the television system, resulting in a 12 cycle flicker of violent magnitude.

If a motion picture projector could be speeded up so that the frames would coincide with the frames of the television picture, the above objection would be removed and a satisfactory picture would result. Thus, if a camera was used to record pictures at a rate of 30 frames per second, a projector of the standard type could be speeded up and synchronized at a rate of 30 frames per second, and with slight modification, could be used in conjunction with a television system, following RMA standards. This would necessitate that all film used for television purposes be filmed especially for television. If regular motion pictures were contemplated, it would be necessary to reprint them with a special printer which would print some of the frames of the original in duplicate on the copy, in order to change the film speed from the standard 24 frames per second to 30 frames per second.

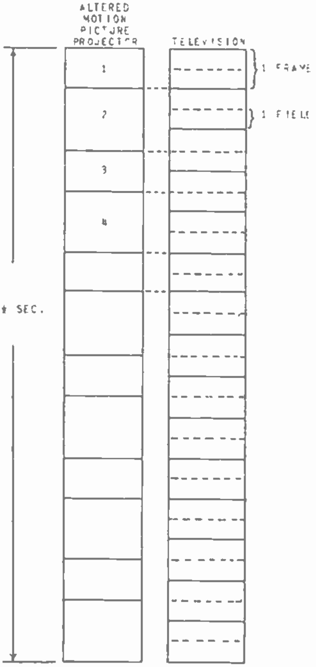


Fig.22 Required film projection rate for a television system.

In certain types of motion picture projectors where the film moves through the camera at a constant rate of speed, a system of lenses is provided which maintains a steady picture of constant illumination on the screen, and the individual frames of the picture have no significance, because one frame dissolves into another without any discontinuity. In this type of projector, which will be described in detail later in the lesson, pictures taken at the rate of 24 frames per second can be used for television at the rate of 60 fields per second, even though the film is run through the projector at a normal rate of 90 feet per minute. However, such projectors are uncommon and expensive for the reason that the intermittent type of projector has been highly developed for theater use. The large sums of money spent in development, and the semi-mass production of standard theater equipment, provide a highly developed projector at reasonable cost. For this reason, it may be preferable to use a standard type of projector with slight modifications.

To use an intermittent projector for television without resorting to special film, the projector must be modified so that every other frame is projected for a slightly longer time, and every other frame projected for a shorter duration. From Fig. 21, which compares standard sound motion picture film with the scanning rate of television, it is apparent that every five fields of the television picture will correspond to two frames (equivalent to four fields) of a standard movie machine. Thus, we have a 4:5 ratio in the field frequency of motion pictures and television.

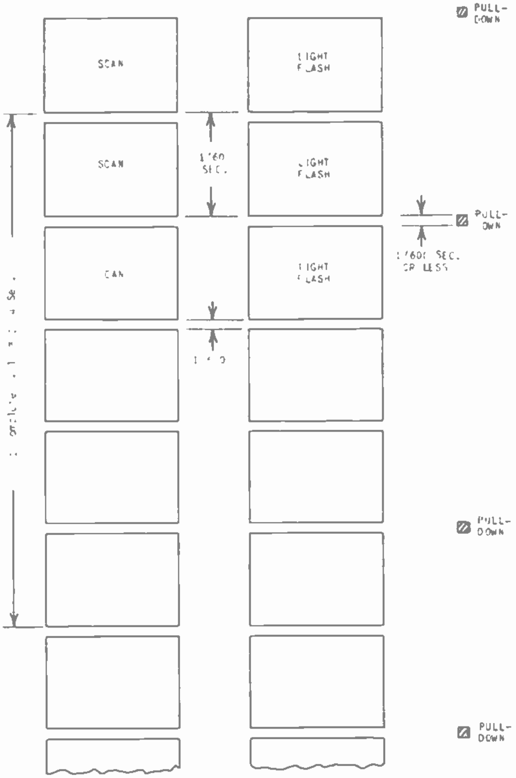


Fig.23 Operation of a theoretical projection system for television.

Fig. 22 shows the change in timing necessary to adapt standard movie films for television. Frame 1 of the picture is flashed on the pickup tube for one television frame. Frame 2 is flashed on the pickup tube for 1½ frames (1½ frames correspond to three fields); that is, the iconoscope or other pickup tube scans 262 lines from top to bottom of the picture, goes back and scans the in-between lines, and on the third trip scans the first group of lines a second time. Frame 3 in the projector is now pulled down by the intermittent and is scanned for two complete fields by the pickup tube. Therefore, frames 1 and 3 each represent a normal television frame. The only difference is that frame 1 first scans the even numbered lines in the picture and then fills in the odd lines; whereas, frame 3 begins by scanning first the odd lines and then filling in the even numbered lines.

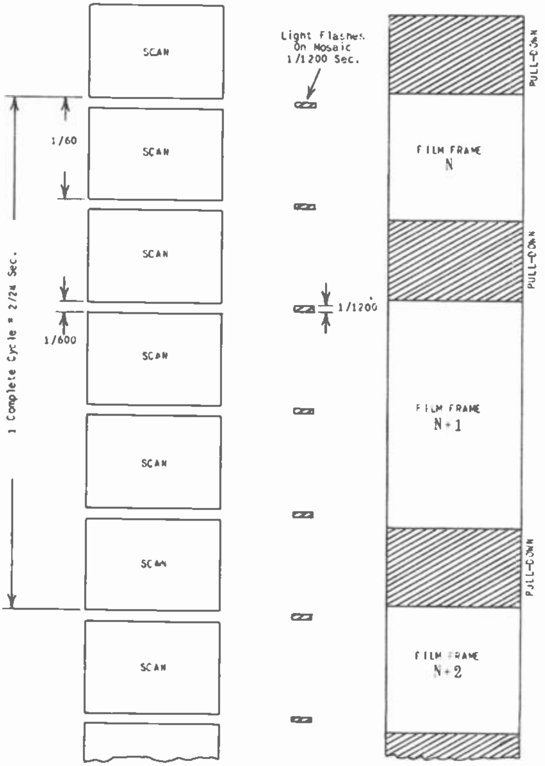


Fig.24 Practical operation of a television projector.

In order for the pickup tube to be active during the entire period of scanning, the intermittent would have to transport the film the complete distance of one frame during the blanking interval of the pickup tube. This is illustrated in Fig. 23, which indicates that the film is pulled down one frame, scanned twice, pulled down one more frame and scanned three times, according to the principles which we have already discussed. If this system could be followed, an intermittent type projector could be used for either the iconoscope (to operate on the storage principle) or with an image dissector (which must be active during the scanning interval). However, the short time, (1/600 second), allowed for the pulldown mechanism to operate, is impractical, and no present-day intermittent is capable of pulling down standard film at this rate without excessive wear on the sprocket holes of the film.

The method shown in Fig. 24 allows a much longer interval for transporting the film. In this case, the film is pulled down and a light is then flashed on the pickup tube for less than 1/600 second (1/1200 second in the example shown). The light flash occurs during the *blanking interval*, and the electrical surge generated by flashing the light on the photosensitive pickup device does not cause flicker in the picture. The pickup tube is in darkness during the scanning interval, and this method is consequently applicable only to pickup tubes utilizing the storage principle, such as the iconoscope. The flash of light during the blanking interval creates a latent electrical image on the face of the iconoscope. The iconoscope must be of sufficiently low leakage that this electrical image will be maintained for at least 1/60 second, and in this case, uniform intensity of the picture will be preserved from top to bottom of the picture, even though it is scanned in darkness.

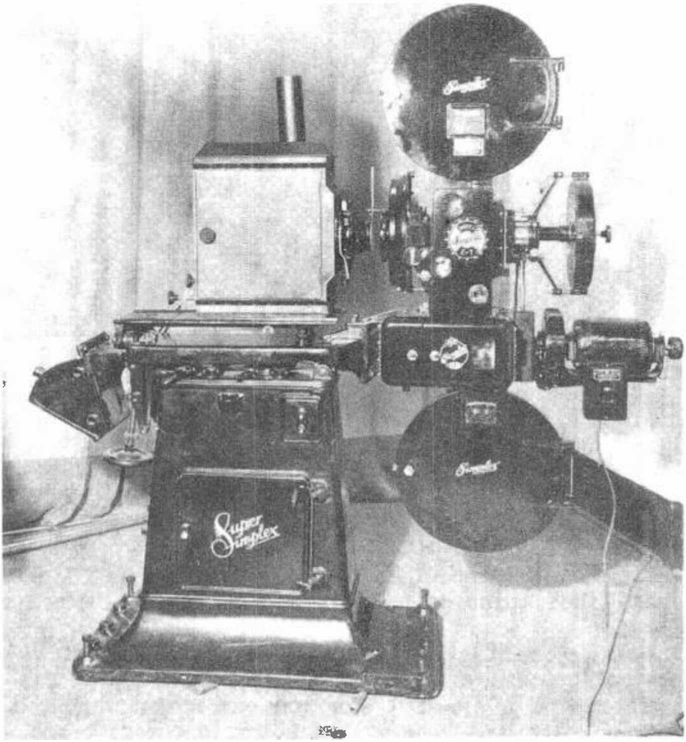


Fig.25A A Simplex projector modified by G.E. for television.

5. INTERMITTENT PROJECTORS AS MODIFIED FOR TELEVISION.

Standard Simplex projectors made by the International Projector Company are modified to meet RCA or General Electric specifications, following the principles outlined above and are sold by

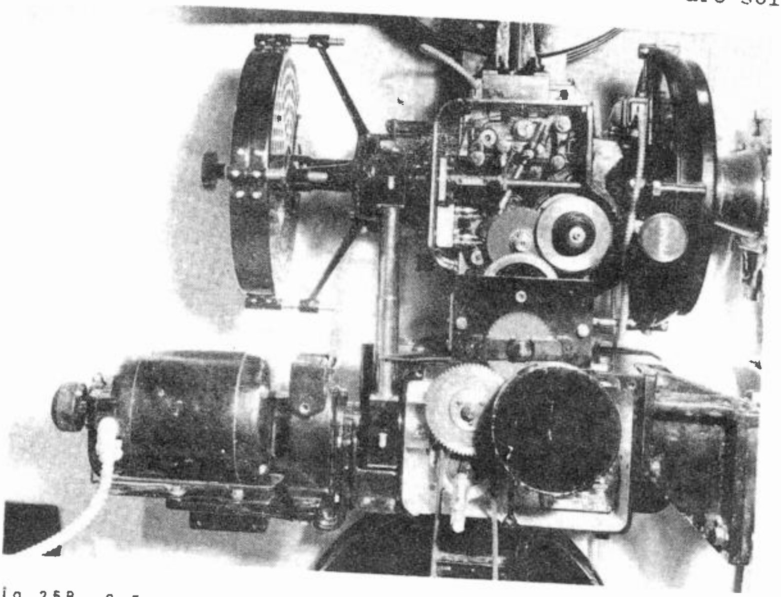


Fig.25B G.E. projector showing gears and internal mechanism.

these two companies. Fig. 25 shows the appearance of the General Electric projector, while Fig. 26 pictures the RCA projector. The modifications of the standard projector are:

1. Addition of a photoelectric cell to interpret the average illumination or background intensity of the picture.
2. Altering the intermittent to obtain a 2:3 ratio of projection intervals.
3. Addition of a special synchronous motor which will lock on only one phase of the 60 cycle power source.
4. Change in the shutter to obtain projection during a brief fixed interval substantially less than the blanking time of the television system.

A schematic sketch of such a modified projector is shown in Fig. 27. The film passes through a light beam which interprets the average light level of the picture within a few frames of the picture being scanned. A voltage is thus obtained which is proportional to the background of the picture. The film is pulled down according to the chart of Fig. 24, by an intermittent cam and spider follower. The shutter allows the picture to be projected on the iconoscope only during the brief blanking interval. The irregular motion of the film is then removed after passing through the lens system, by the film loop and sprocket K which rotates at a uniform rate of speed, being stabilized by a flywheel. The motion picture film then passes through the sound head at a uniform rate of speed and the sound track is scanned.

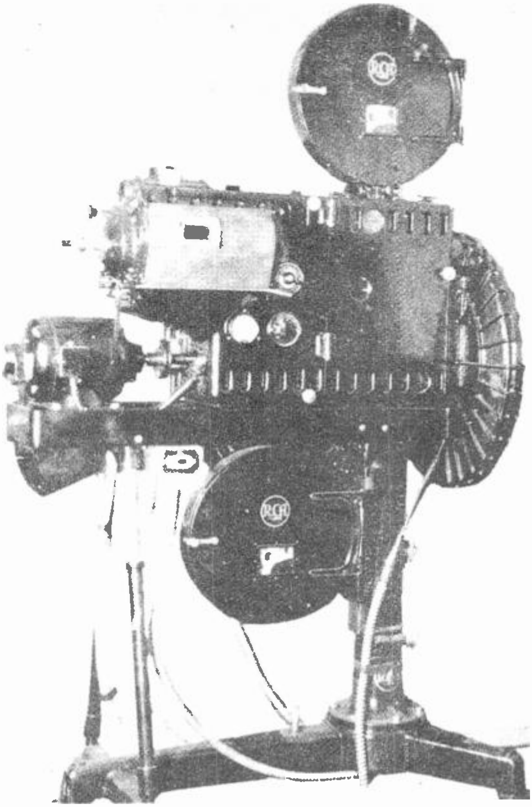


Fig.26 RCA television film projector.

Since the video amplifiers from the camera are AC coupled, only the contrast of the picture is carried as information. The background of the picture is inserted as a change of bias in the output stages of the video amplifier. In studio pictures, this background insertion is done manually, since the average illumination is approximately the same at all times and in any case would be anticipated by the operator because of previous rehearsals and cues. In the case of motion picture film, the background varies widely from scene to scene, and if manual control were used, the operator would take a fraction of a second to readjust the background level as the scenes change, resulting in a surge, or bounce in the picture background. The output of a background control photocell, shown in Fig. 27 is amplified by a DC amplifier and the output applied as changes in bias to the background control circuit which you have studied in a previous lesson.

A normal intermittent rotates at a rate of 24 rps. A stud on the cam engages the spider follower once each revolution and rotates it by an amount sufficient to transfer the film a distance of one frame. For television, the intermittent cam is redesigned and equipped with two studs which throw the spider follower twice each revolution instead of once. A further alteration which is necessary is a change in the gearing mechanism so that the intermittent cam rotates at 12 rps instead of 24. Since it rotates at half speed, but has twice as many throws per revolution, the film is transported at the same average rate as previously; that is, 24 frames per second. The throws are arranged 144° apart and 216° apart. This corresponds to 2/5 and 3/5 of a revolution, thereby maintaining the 2:3 ratio of scanning which we have shown to be necessary.

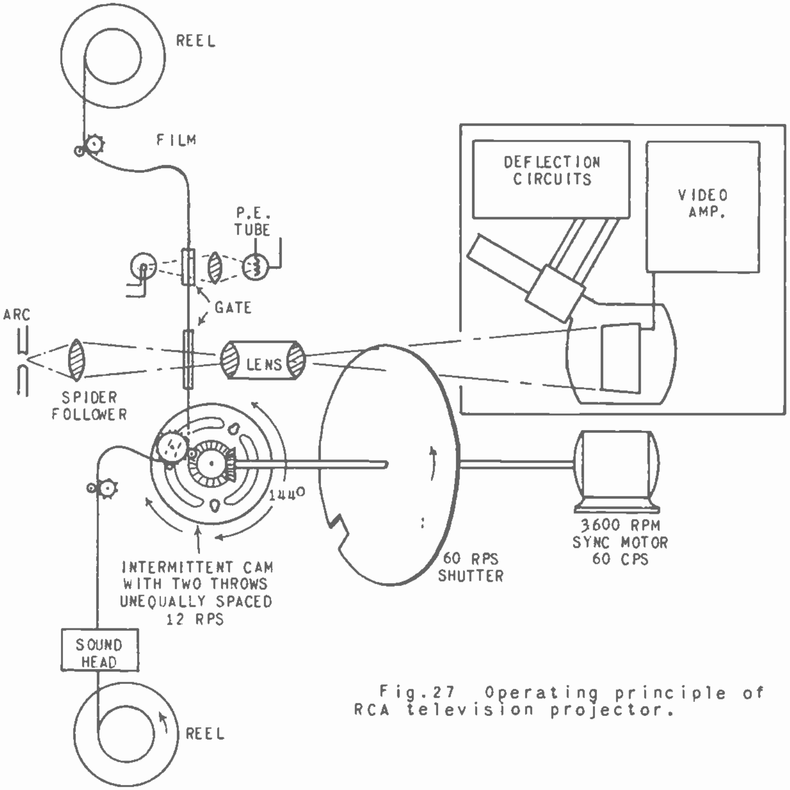


Fig.27 Operating principle of RCA television projector.

The intermittent cam and the shutter are driven by a 3600 rpm synchronous motor which may be either a two phase or a three phase machine. This motor is operated from the same supply to which the synchronizing generator of the television circuit is connected. Since the synchronizing generator is tied in to the 60 cycle supply, the blanking pulses will occur at exactly the same rate as the shutter operation and pull-down mechanism. However, in order for the film projection to occur during the blanking interval, the motor drive of the projector and the blanking generated by the synchronizing generator, must be in phase. Adjustment of this phase relation may be made either by changing the phase of the synchronizing generator with respect to its supply line (by one of the means given in a previous lesson), or the frame of the synchronous motor may be rotated in its mount on the projector chassis. In either case, some means must be provided for insuring that the synchronous motors will always start on the same phase. This automatic synchronization is supplied in either the General Electric or RCA machine. In the case of the RCA projector, a special synchronous motor is used, which includes a DC winding for fixing the polarity of the poles.

The shutter may be equipped with one slot and rotated at 3600 rpm or the shutter may contain two slots and be rotated at 1800 rpm. In either case, the picture will be flashed on the pickup tube 60 times per second. The width of the slot in the shutter must be at least as wide as the aperture of the lens in order not to reduce the light efficiency. Consequently, the diameter of the shutter must be quite large; approximately two feet. However, the increased momentum of such a large shutter increases the stability of projection. The rotating shutter is normally supplied with a slot of fixed width, but a variable width slot can be obtained by using two shutter discs clamped together but arranged to rotate slightly with respect to each other by means of set screws. By rotating the two shutter discs with respect to each other so that the slots do not coincide, the effective opening is reduced.

In the General Electric projector, two shutter discs are mounted on a single shaft and rotate one before and one after the projection lens. Since the picture is inverted in passing through the lens system, the effect is as if two shutters were used on opposite sides of the camera and rotated in opposite directions; that is, if a single shutter is off to one side and rotating in such a way that the bottom of the picture becomes exposed before the top of the picture, the effect would be to place an additional shutter on the opposite side of the camera and rotating in such a way as to expose first the top and then the bottom of the picture. Any distortion which may be introduced by the trapezoidal shape of the shutter opening is thereby reduced because of the balancing effect of the additional shutter. Likewise, any uneven change which might be produced by exposing the picture from top to bottom is somewhat offset by the simultaneous exposure from bottom to top.

The projector shown in Fig. 25 is powered with a 1/3 hp., 220 volt, three phase, 60 cycle, self starting motor with automatic synchronization. The heavy cast base of the machine prevents vibration, and a vertical picture stability of 1/8 of 1% is claimed. This means that successive frames will register one on top of the other to within 1/8 of 1% of the height of the frame. In a motion picture projector in a movie theater, vertical instability would show up as vibration or flicker in the picture. When an iconoscope is used, a charge is stored on the signal plate of the pickup tube. If any vertical movement of the picture is involved, the energy stored over various parts of the mosaic will be modified an instant later by the change in picture position. This will have the effect of destroying the vertical resolution of the picture.

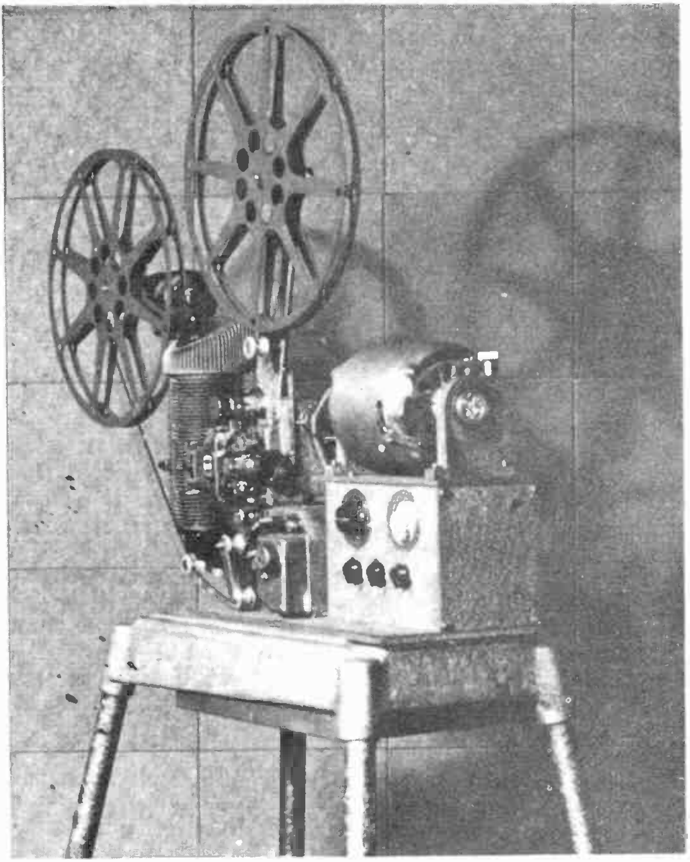


Fig.28 Bell & Howell 16 mm. projector for television.

Sixteen millimeter movies, as well as standard 35 mm. movies, are used in television. A complete discussion of the relative merits of the two film widths will be taken up in detail later in the lesson.

One standard 16 mm. projector which has been modified for television is shown in Fig. 28. This one is manufactured by the Bell and Howell Co. In this projector, pull-down time is so short that it is possible to stagger the pull-down interval with the projection interval in such a way that a standard intermittent can be used; that is, the pull-down occurs once every 1/24 second, even though one frame is scanned twice while the following frame is scanned three times. The timing of the pull-down which makes this arrangement possible is shown in Fig. 29.

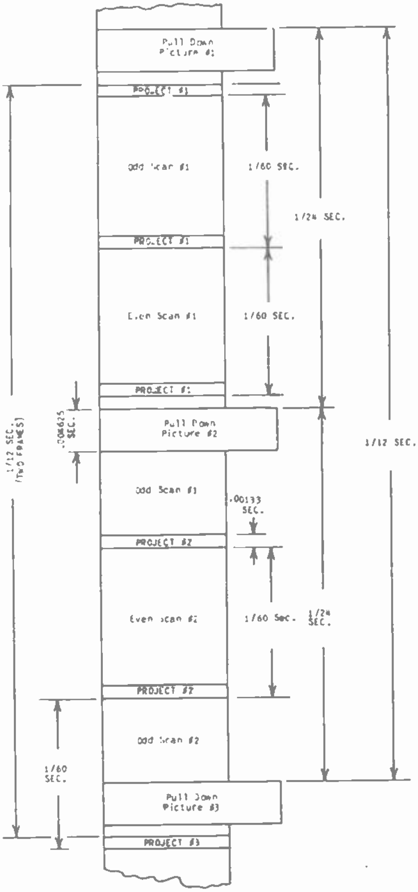


Fig.29 Timing of operations in Bell & Howell television model projector.

A rotating shutter between the projection lamp and the film gate is incorporated to allow one projection every 1/60 second. The duration of a projection is 8% of the time for one field; that is, 8% of 1/60 second or .0013 second. The projector is powered with a 3600 rpm synchronous motor which can be rotated about its axis during operation for accurate adjustment of the protection interval with respect to the phase of the 60 cycle supply. Incandescent illumination in conjunction with the 8% shutter opening and F:2 lens supplies an illumination of 140 foot-condles in the highlights.

The standard 35 mm. projector can be equipped with either an incandescent lamp or a standard motion picture arc. The projector shown in Fig. 25 is equipped with incandescent illumination, while the projector shown in Fig. 26 is equipped for arc light.

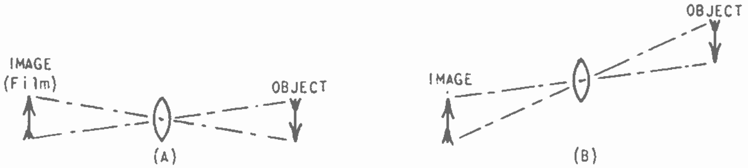


Fig.30 Principle of "rising front" camera.

6. CONTINUOUS PROJECTOR. If the student happens to be an amateur photographer, he will be acquainted with the term "rising front", which means a camera whose lens may be moved up and down with respect to the negative. By means of the rising front on a camera, it is possible to take a picture of an object which is elevated above the level of the camera without tilting the camera up at the object in the normal way. This is shown in Fig. 30. Part A shows the object in the normal position, and the inverted image on the film. Part B indicates that by raising the lens, the position of the object may be raised, and still focused on the same portion of film.

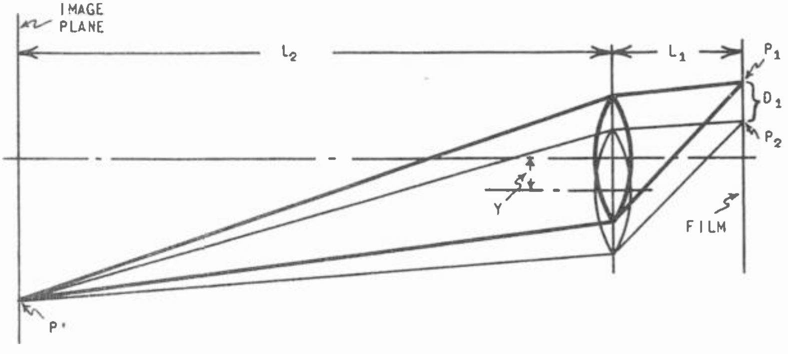


Fig.31 Principle of optical rectification as applied in a continuous motion projector.

The above illustrates the action of a rising front camera, but the situation in a projector is somewhat different. The film occupies the position of the object of Fig. 30, whereas, the image is flashed upon a pickup tube. It is apparent that if the film is moved to occupy the position of the object, it is possible to move both the lens and the film at the same time, while the image remains stationary on the screen of the pickup tube. Fig. 31 will portray this idea more clearly. The motion of the film and the motion of the lens may be equated as follows:

$$D_1 = -y\ (1/M - 1)$$

Where: D_1 = film displacement,
y = lens displacement,
M = magnification due to lens.

Since M is constant, we can say that the motion of the lens is equal to the motion of the film divided by a constant. In other words, if the film is traveling at a uniform rate of speed, and a system of lenses is also made to travel at a uniform, but slower rate of speed, the image of the film will remain stationary upon the screen. Several systems are possible for transporting the film and a system of lenses at the same time. One method for doing this has been developed by the Farnsworth Company and is discussed in detail later in the lesson.

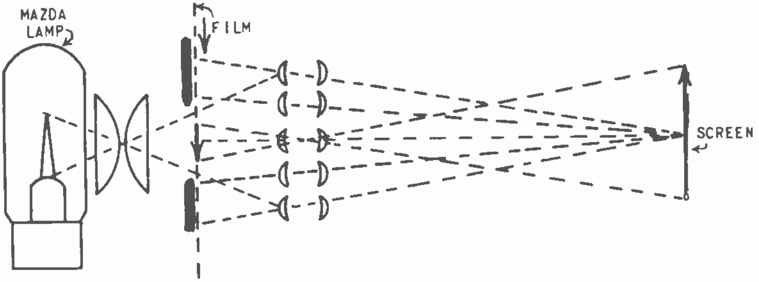


Fig.32 An optical rectifier employing a system of moving lenses.

A system of moving lenses or any other device in a non-intermittent projector which compensates for the motion of the film is known as an "optical rectifier". An optical rectifier of the type described in the last few paragraphs is diagrammed in Fig. 32. The lenses may be transported in a variety of ways. They can be mounted in holders which slide down past the film gate in close contact with each other and are then transported by means of a chain or endless belt to a position above the film gate. They may be operated with a system of hinged levers and cams, or they may be placed on the circumference of a wheel having a radius sufficiently great so that the vertical motion of the lenses will be essentially a linear function against time. All of these methods require extreme precision in the operating mechanism in order that the images from each of the lenses will coincide on the screen. Failure of complete register of the images will result in loss of resolution.

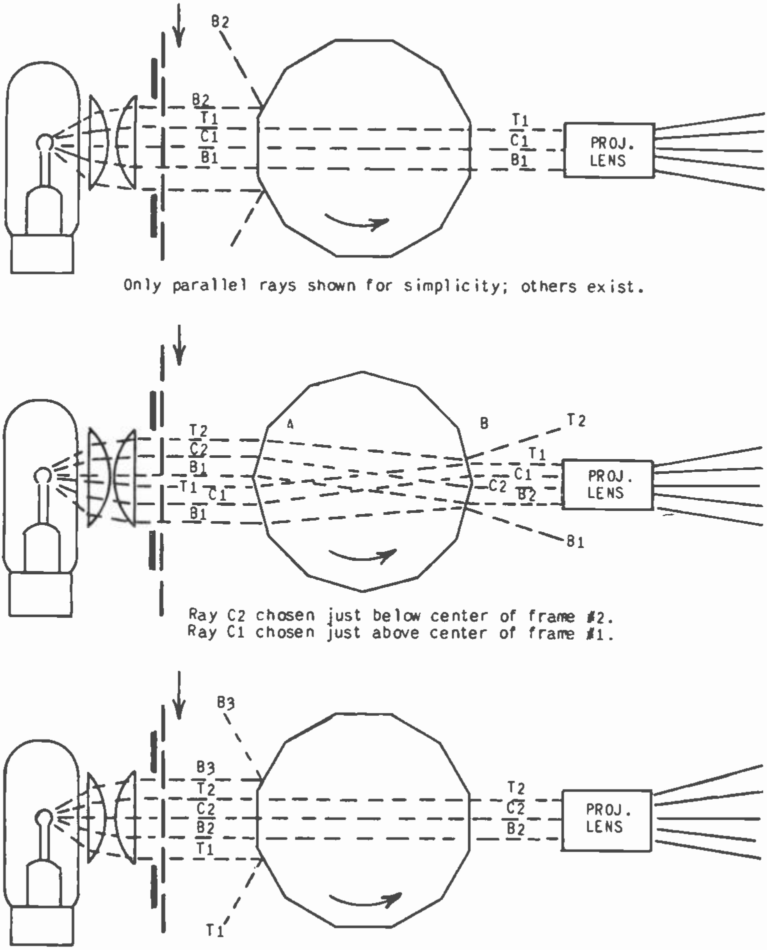
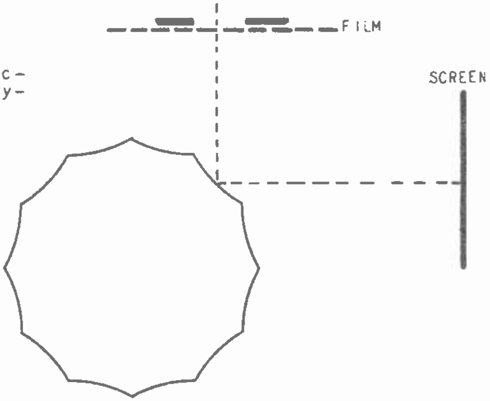


Fig.33 An optical rectifier employing a polygon glass prism.

One simple type of optical rectifier is shown in Fig. 33. In this system, a block of glass in the shape of a polygon is rotated in front of a moving film. This causes a movement of the object in the opposite direction to that produced by the film's motion. The principle of rectification is the successive bending at points AB due to refraction by the glass. Vertical motion of the image follows a sine function instead of the desired linear motion. This error of rectification can be diminished, but not entirely corrected, by using a large polygon, having many sides. Furthermore, the glass prism introduces some chromatic aberration which cannot be completely ocrrected by supplementary lenses. A slight modification of this method uses a polygon mirror instead of a polygon prism as shown in Fig. 34. The sine wave error of rectification can be reduced by properly shaping the faces of the mirror.

Fig.34 An optical rectifier employing a polygon mirror.



If a continuous motion projector is to be used with an iconoscope pickup tube, it is essential that the projector have constant light output. This is not easy to do and many continuous motion projectors suffer from variation in light output at the frame frequency, which cause surges in the video amplifiers. Certain rules must be followed to maintain constant light output. In the system employing a multiplicity of lenses, the image of the light source, in the plane of the lenses, must cover three or four lenses and the gate must be an integral number of frames in height; usually one or two. If a rotating prism or polygon mirror is used, the gate must expose two frames of the picture and the image of the light source must fill one face of the prism with uniform light. Furthermore, the projection lens system must cover three faces of the prism. By observing such refinements, practically constant light output can be achieved.

The continuous motion type of projector is difficult to construct because of the precision which must be maintained to prevent motion in the picture, but the method has the advantage of flexibility. By using continuous motion, constant illumination projection, any film may be used regardless of the rate at which it was recorded. For instance, 24 frame pictures recorded for standard motion picture work would be interchangeable with 30 frame pictures produced specifically for television. Furthermore, a picture can be slowed down or speeded up slightly in order to adjust its timing to fit the program schedule.

Continuous motion type projectors are not used in standard motion picture work because of their greater complexity and lower light output; however, the latter drawback is not objectionable for television. Ordinary theaters require about 500 times the light output necessary for television. With a continuous motion projector, either an iconoscope or image dissector tube may be used.

Since the motion of the optical rectifier must bear a definite relation to the speed of film travel, some means must be provided for accommodating film shrinkage. Depending upon the manner in which the film was processed, the length of film for any given number of frames will vary slightly, as will the frame-to-frame spacing. Some shrinkage can be compensated by the use of supplementary lenses to modify the magnification of the projector lens, according to the following relations.

$$\frac{\text{Speed of lens travel}}{\text{Speed of film travel}} = \frac{M}{M + 1}$$

Where: M is the magnification of lens, and is variable by the use of supplementary lenses. The speed of lens travel is fixed, but speed of film travel depends upon film shrinkage.

Another method of compensating for film shrinkage will be discussed under the CBS projector.

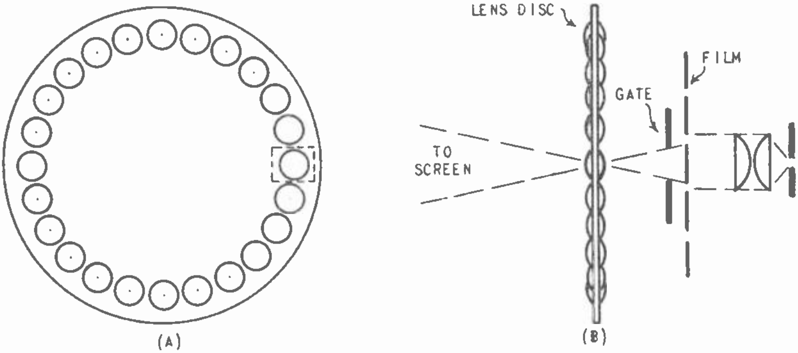


Fig.35 An optical rectifier employing a system of moving lenses in a disc scanner.

7. FARNSWORTH PROJECTOR. In the Farnsworth projector, the optical compensator which causes the projected image to remain stationary while the film is in motion, consists of two sets of lenses mounted in rotating discs. If only one lens disc were used to rotate in front of the picture gate as shown in Fig. 35, the lenses in moving on the circumference of a circle would have a component of horizontal as well as vertical displacement. Optical rectification in the vertical direction would be achieved, but the picture would also be moved sidewise on the screen, resulting in a loss of resolution. In practice, two discs are used as close together as possible, and rotating in opposite directions as shown in Fig. 36. Any horizontal motion contributed by one disc is offset by an equal amount of opposite horizontal motion contributed by the other disc.

The use of twin discs has a further advantage in that the two element lenses formed by the two discs have adjustable focal lengths and are set by changing the spacing between the individual elements. Thus, every pair of lenses may be adjusted to have the same focal length as every other pair. This condition is necessary to prevent the picture from going in and out of focus as various pairs of lenses are used.

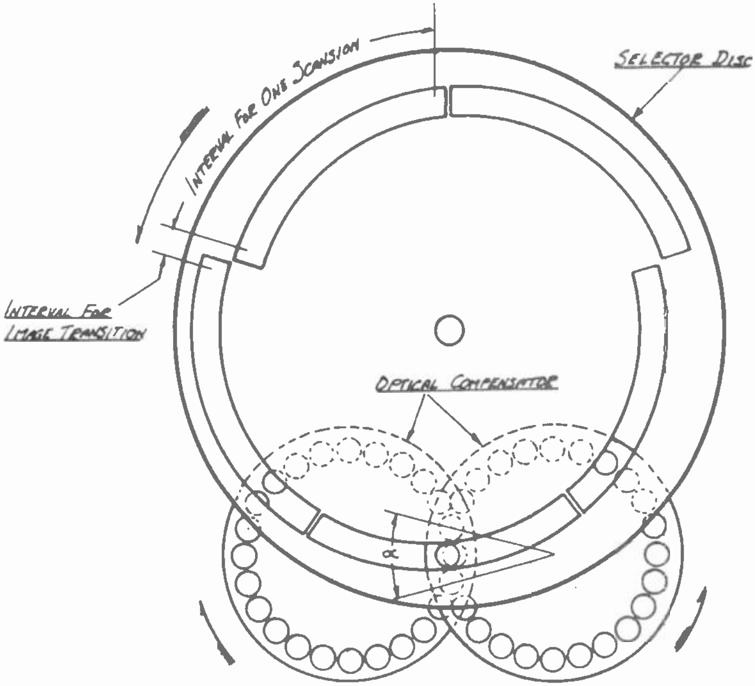


Fig.36 An optical rectifier employed in the Farnsworth continuous motion projector.

The position of the lenses is adjustable on the supporting disc and in this way, very accurate registration of the images from all of the lens pairs can be had. Fig. 37 clearly shows the mounting of the individual lenses and the adjusting screws.

The selector disc in Fig. 36 and 37 is arranged with one group of three slots on a spiral and a second group of two slots on a similar spiral. These slots are arranged to select only one lens pair at a time so that the picture is projected through only one pair of lenses at any one instant. This is necessary because the lens elements do not travel uniformly in a vertical direction, but follow a sine law since they are laid out on the circumference of a circle. The distortion produced is not prohibitive, provided multiple images are not supplanted, one on the other by being projected through several pairs of lenses, occupying different positions in the vertical plane, and therefore, traveling at different rates of speed.

The slots in the selector disc move downward at the same rate as the individual lenses in the optical compensator. One frame is followed down for a time equivalent to two scanning intervals of the pickup tube; in other words, for one complete frame. The next motion picture frame is followed downward for a time equivalent to three scanning intervals in the pickup tube, corresponding to $1\frac{1}{2}$ television frames. Thus, the 2:3 ratio is preserved as required in the use of 24 frame motion picture film, for 30 frame television.

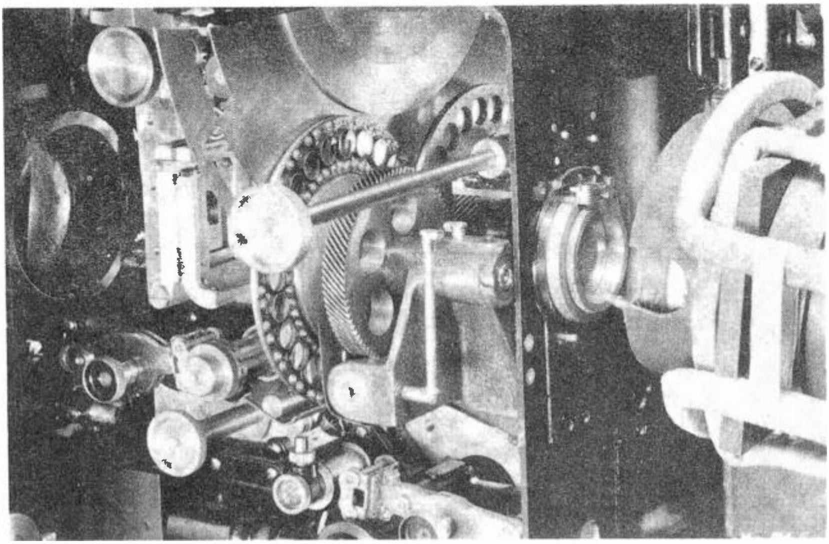


Fig.37 Interior view of Farnsworth projector.

A picture of the complete projector is shown in Fig. 38, and it is seen to be no more bulky than the ordinary intermittent type of projector, in spite of the multiplicity of lenses. Because the picture is projected during the entire scanning time, the illumination intensity required is very much smaller than in the case of an intermittent projector where the picture must be flashed at brief intervals of less than 10% of the total time. For this reason, the lenses in the optical compensator do not have to have a particularly wide aperture. The choice of the optimum *f* number and focal length is fairly complex and will not be discussed in this lesson. The lens disc is geared directly to the sprocket for transporting the film so that the downward motion of these two factors will be in exact synchronism. The gearing mechanism performing this operation is shown in Fig. 39. The lens selector disc is clearly shown in Fig. 40. An ordinary projection Mazda incandescent lamp will project a total light in excess of 40 lumens. An auxiliary lens is incorporated for modifying the focal length of the lens disc; thereby changing the image size.

In the intermittent type projector, precise vertical positioning of successive frames is necessary in order that vertical resolution of the picture be maintained. This necessity is partly a consequence of the use of interlacing. Accurate horizontal positioning is achieved by passing the film through a close-fitting gate, to prevent horizontal weave of the film. In the case of continuous motion projectors, it is essential that the film be run at an absolutely constant rate of speed; otherwise, vertical vibration of the film will result in loss of vertical resolution. This means that the driving sprocket must be accurately machined and designed so that changes in film shrinkage can be accommodated. Furthermore, registration between pictures projected by successive lenses in the lens disc must be perfect. In the Farnsworth projector, the combined error, due to gears, lens settings, etc., is so small that image unsteadiness does not exceed 1/8 of 1%. This is equal to the steadiness obtained from a correctly designed intermittent type projector.

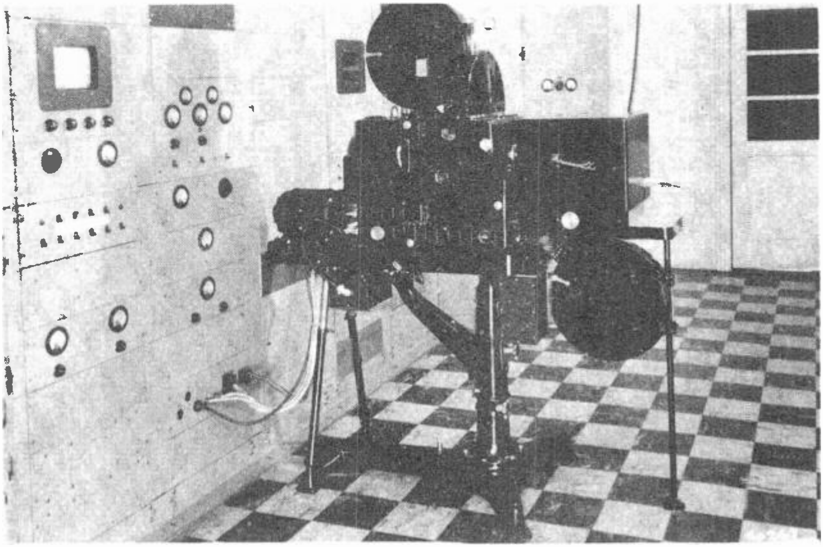


Fig.38 Farnsworth projector.

8. CBS PROJECTOR. In the intermittent type projector, the film, the lens system, and the image; all three remain stationary during the projection interval. In the continuous motion type of projector, the film and the optical rectifier were each in motion, while the image remained still. In the television film scanner developed by P.C. Goldmark for the Columbia Broadcasting System, the lenses remain stationary, while the film is in continuous motion. This causes the image to move on the pickup tube in a vertical direction. This image is then scanned in a direction

opposite to the motion of the image. In other words, the vertical component of scanning consists of a combination of picture motion and scanning motion. The scanning motion is 1½ times as fast as the picture motion.

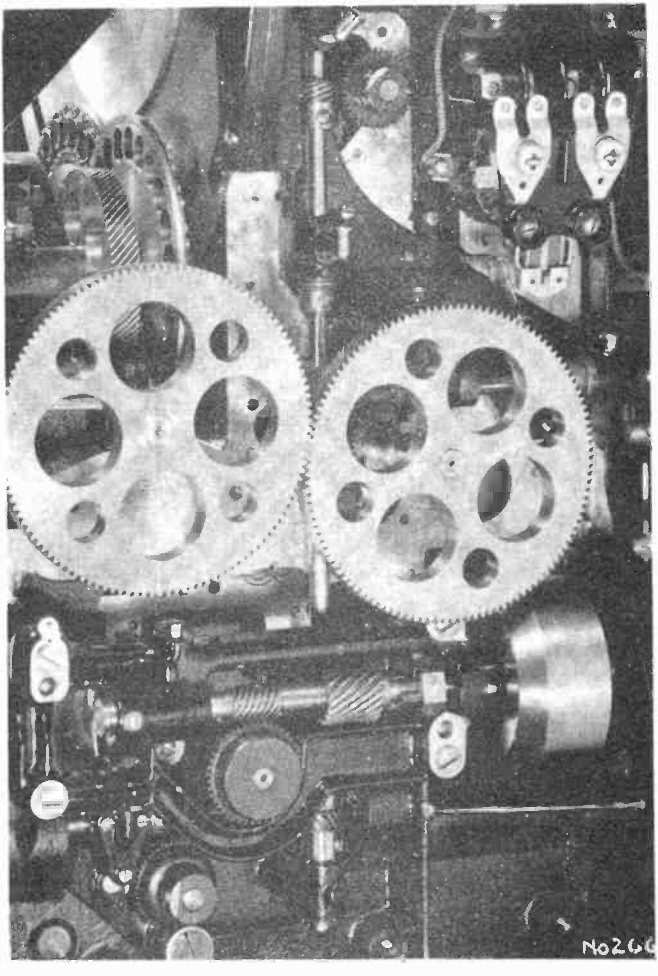


Fig.39 Gear train for driving the lens disc.

Suppose the film is traveling downward through the gate as illustrated in Fig. 41. The gate is 1½ times the width of one frame and at the first instant shown, the picture frame to be scanned has entered 3/5 of the way within the gate. A scanning spot starts from the bottom of this frame at point 1 and scans upward. By the time the spot reaches the upper limit of the gate, the film will have moved downward so that the entire frame is within the opening of the gate. Thus, during the interval between columns one and two of Fig. 41, the spot has scanned one complete frame. Now, if it were possible to cause the scanning

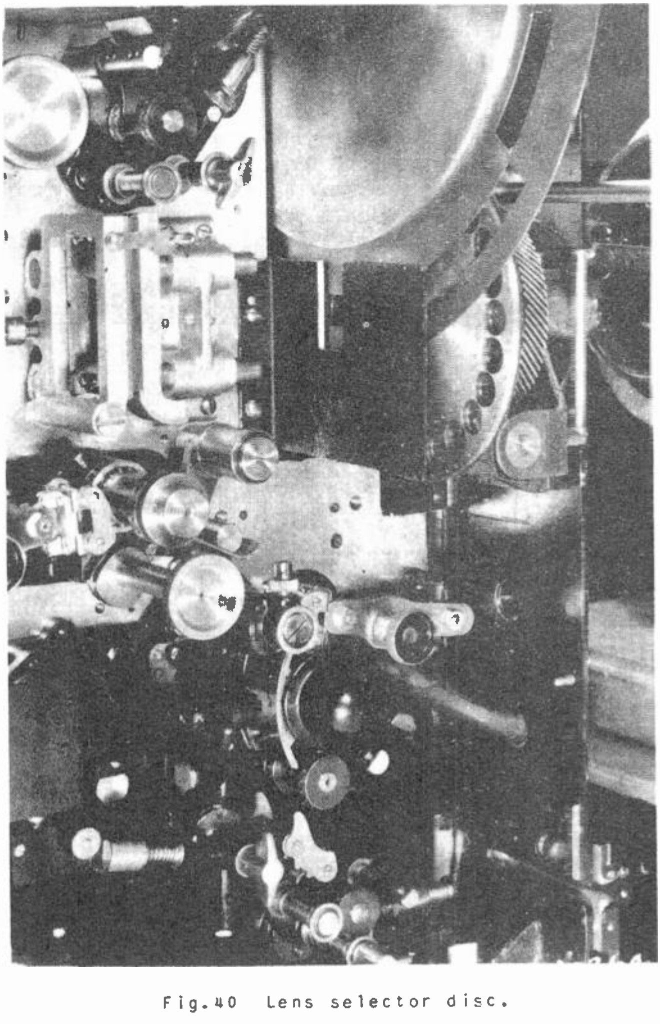


Fig. 40 Lens selector disc.

spot to return to the base of the picture, (which is not in the same position, but has been moved downward by two-fifths of a frame from its original position), the scanning of the same frame a second time would be illustrated by the second column of Fig. 41. Once more, an upward motion of 3/5 of a frame plus a downward motion of 2/5 of a frame results in a complete scansion.

Suppose that 1/60 second later the scanning spot could once more be transported to the bottom of the picture. This would now coincide with the bottom of the gate, and the scanning spot would scan frame 1 a third time. It could not scan it a fourth time because the frame would now have passed beyond the bottom of the gate.

If the scanning spot jumps up one frame and begins to scan the following frame, its position would then be indicated by column 4 of Fig. 41. This frame could be scanned a second time as indicated by column 5. It could not be scanned a third time because frame 2 would now pass beyond the bottom of the gate. If the spot jumps up to begin scanning the bottom of the third frame as indicated in column 6, everything is now exactly as it was in column 1 except that two frames have passed through the gate. This completes a cycle of two frames; frame 1 was scanned three times, and frame 2 was scanned twice.

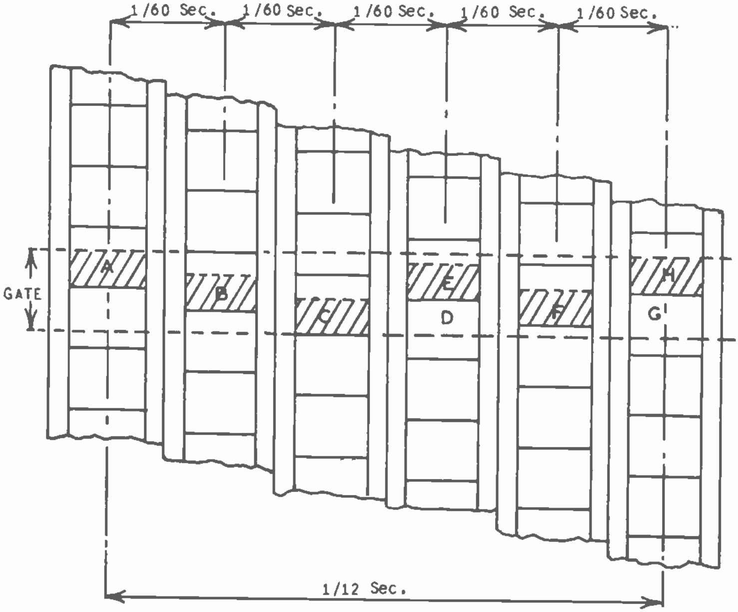


Fig.41 Film travel in CBS projector.

If the film is traveling at the rate of 24 frames per second, a television picture would be scanned at a rate of 60 fields per second by the above process. However, it would be impractical to cause the scanning spot to jump around on the signal plate of a pickup tube as we have suggested in the preceding discussion. A picture would result, by jumping the vertical position of the beam around, but it would not be practical to secure accurate positioning of the scanning lines, which is required for high definition television. A loss of vertical resolution would result.

The scanner developed for the Columbia Broadcasting System which operates on the above principle utilizes a group of five similar lenses which are displaced so that a single image is projected through the five lenses from five similar objects, having the positions shown in Fig. 42. Since the film is moving, these displaced lenses, if exposed at the correct time, will cause

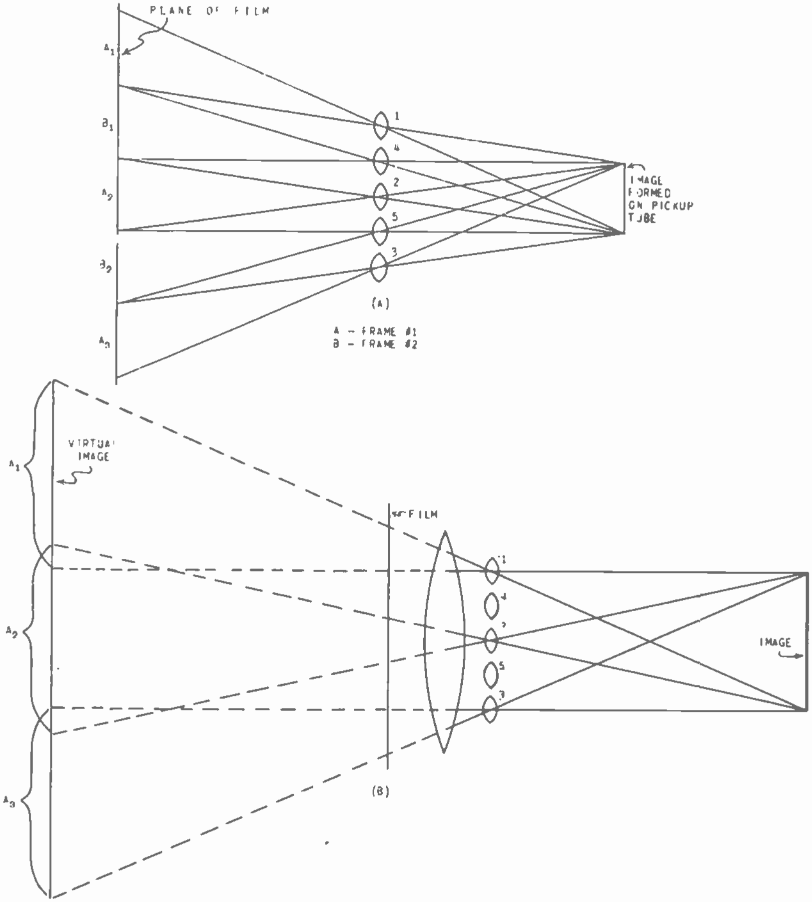


Fig.42 Lens system in CBS projector.

the moving images of the film to be returned to the same point on the photosensitive cathode of the pickup tube. In Fig. 42A, the three positions of frame A are shown separated by a distance sufficient to insert frame B. Actually, the three positions of frame A are not separated, but overlap. This is necessary because the frame is scanned in 1/60 second, which is insufficient time for the film to travel a distance of one frame. To achieve overlapping of the objects, an additional lens is inserted between the film and the co-planer lenses (numbered 1, 2, 3, 4, 5). The film lies inside the focal length of the added lens so that an enlarged virtual image of the film is formed back of the film gate. This is shown in Fig. 42B. It can be seen that the virtual images of the film overlap at successive intervals. Frame B was omitted to avoid complicating the diagram; construction lines for the virtual images were omitted for the same reason. The student should construct these virtual images as a review of a previous lesson.

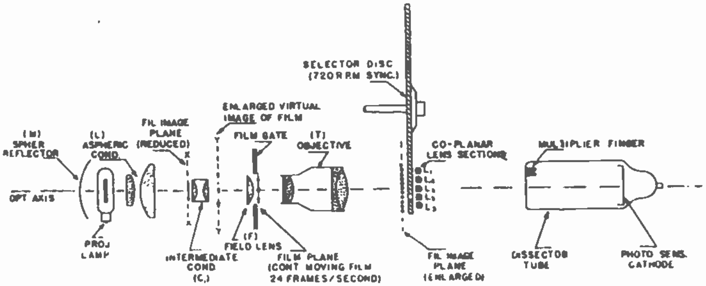


Fig.43 Schematic diagram of CBS projector.

Fig. 43 shows the practical schematic of the CBS projector. The lenses are numbered in the order in which they are exposed through slots in the selector disc. By splitting the picture into five images, the total light delivered to the multiplier dissector tube is only one-fifth that obtainable from other types of continuous motion projectors. However, since the light is incident during the entire scanning time, adequate illumination is available from the 1500 watt projection lamp which is incorporated in the practical projector shown in Fig. 44. The five co-planer lenses of the projector are rigidly mounted on a heavy metal plate by means of adjustable brackets. Since the image is alternately projected through each of the five lenses, it is essential that the images coincide exactly if good vertical resolution is to be preserved. The method of aligning the lenses is as follows. A reel of motion picture film carrying a resolution chart is projected, while three of the lenses are covered up. The images from the other two lenses are brought into coincidence by adjusting one lens mount and observing the monitored image on the kinescope. Adequate resolution is obtained. The other three lenses are uncovered, one at a time, and the image from each brought into coincidence with the first pair before another is uncovered. The images can be aligned to within a fraction of a scanning line.

As in other types of continuous motion projectors, the adjustment of the lenses depends upon the frame-to-frame spacing of the film being used. This changes from film to film because of variations in shrinkage and some method must be provided to compensate for changing film lengths. In this projector, the correction is accomplished by slightly changing the distance between the projector and the dissector pickup tube, and then refocusing the projection lens to secure a sharp image. In practice, the shrinkage of the film is first determined by measuring a length of film against a metal standard. The distance between the dissector pickup tube and the projector is predetermined for various values of shrinkage, and the control used to set this distance is calibrated directly in film shrinkage. When the projector is set up for one particular value of film shrinkage, all that remains when the projector is started is to focus the single projection lens.

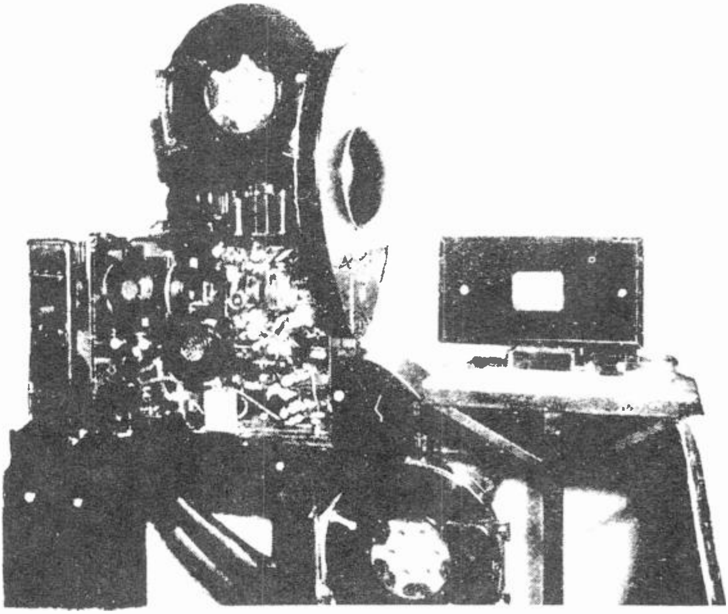


Fig.44 The CBS projector developed by Dr. Goldmark.

Since the image is in motion while it is being scanned, pickup tubes which operate on the storage principle, such as the iconoscope or orthicron, cannot be used with this projector. The projector was built primarily to be used with a Farnsworth image dissector type of pickup tube, which has a distinct advantage over the iconoscope in freedom from shading difficulties and in its ability to produce pictures of greater contrast. When sufficient light is available, as in the projector just described, the image dissector is capable of producing pictures having excellent resolution and range in tone.

9. COMPARISON BETWEEN 35 MM. AND 16 MM. FILM. At one time, 35 mm. film was used almost exclusively for television, inasmuch as it was readily obtainable and standard projectors with the

alterations described above could be used. At present, both 16 and 35 mm. films are being used. Small gauge film has several advantages, the chief of which is its vastly lower cost. It occupies less space and it is lighter, which makes it easier to transport and to store. Because of the low cost of 16 mm. film, safety film is used almost exclusively, which together with the reduced size, results in lesser fire hazard. Another advantage of 16 mm. film is that portable camera equipment is available at reasonable prices, making it possible for a television station to record its own film either for entertainment purposes, or to record sports and news events when it is impractical or undesirable to make a direct pickup.

In selecting the correct width of film for television, a size must be chosen which will satisfy the maximum utilization of the television standards. Quality of motion picture film below the maximum possible quality of television cannot be tolerated. Thirty-five millimeter film, even when it is carelessly handled, will meet the requirements of the best 525-line television. The grain is sufficiently fine and the film speed sufficiently great that sound can be recorded out to 15 kc.

Sixteen millimeter film can be made to have sufficient resolution to satisfy the best television pictures *if* considerable refinement is included in the projector. Whether or not 16 mm. film is capable of high fidelity sound recording is somewhat problematical, and depends, for one thing, on the maximum fidelity of sound which will eventually be carried on a television chain.

Film resolution is normally stated in lines per millimeter which can be resolved. The standard is somewhat different than in television. In motion picture work, if we have 10 black lines occupying a space of 1 millimeter, we say that we are resolving 10 lines per millimeter. Actually, 20 spaces have been resolved, because each of the black lines was separated by a white one. In television, if we have a 500-line picture, this means the total number of black *and* white lines which can be placed in the height of the picture. For a picture 1 millimeter high, 10-line motion picture resolution would be equivalent to 20-line television picture resolution.

Various types of film exhibit various grain sizes and this is tied up to a certain extent with the speed of the film. An extremely fast film will have, in general, a somewhat larger grain size. The standard types of 16 mm. film used today can resolve between 40 and 60 lines per millimeter, which is equivalent to a 576 and an 865-line television picture respectively. Therefore, the film itself is not a limiting factor as any of the standard black and white films will have sufficient resolution for television purposes.

The resolution of the picture is also determined by the circle of confusion of the lens in the camera and in the projector. If we wish to resolve 40 lines per millimeter on black and white film, the circle of confusion of the lens must be at least as small as the width of one line, or in other words, it must have a diameter of 1/80 of a millimeter. This is equivalent to a circle of confusion of .0005 inch, which requires a very high quality lens. Since the circle of confusion of both the taking and the projection lenses play a part in the final resolution, the lenses should have a resolving power somewhat better than .0005 inch. Fortunately, modern, high quality lenses are capable of this resolution, even at wide apertures.

Another factor which plays an important part in determining the resolution of the final picture, is the precise position of successive frames of the picture. If the picture moves ever so slightly, the image movement is multiplied by the magnification of the lens system in the projector. For instance, if the image moves as much as .001 inch, blurring of the picture will result equivalent to reducing the resolution of the television system to 284 lines. This is not good enough for television. The motion of the film causing this loss of resolution may be due to a number of things. The most obvious cause is inaccuracy in the intermittent of the projector. Unless the bearings in the intermittent are entirely free from vibration or play of any kind, and unless the sprocket which pulls down the film is very carefully machined, it will not be possible to pull down one frame of the film to within .001 inch or better of its correct position. A guide must be provided to maintain the horizontal position of the film in the gate to within a fraction of .001 inch, particularly in using the storage type of pickup, such as an iconoscope. In this case, horizontal motion would result not only in unsteadiness of the picture, but also in loss of resolution.

Accurate positioning of the film is necessary not only in the projector, but also in the motion picture camera. Any errors in these two mechanisms would add up to destroy the final picture quality. Extremely accurate registration is required in printing copies of the original film, as error in the location of the picture with respect to the sprocket holes at this point would result in inaccurate positioning in the projector. Accurate location in the camera and the projector is achieved by the use of registration pins which form a snug fit in the sprocket holes of the film. These registration pins are reciprocating and enter one of the socket holes during the time of exposure or projection.

In ordinary 16 mm. movies and projectors used by amateurs and in certain types of commercial work, the sum of all of the above errors averages .001 inch in the placing of the film in the projector. As mentioned, this is equivalent to a television picture of only 284 lines resolution, and is, therefore, not satisfactory. By the use of special cameras and special projectors using registration pins and excellent lenses; also through the use of a special accurate printer and fine grain film, a net resolution may be obtained which is equivalent to a 575-line television picture. The refinements necessary for this resolution are mechanical in nature, and, once set up, do not add to the operating difficulty.

For standard 16 mm. films which run at the rate of 36 feet per minute, flat frequency response to 6,000 cycles is obtainable without getting into excessive noise due to film grain, thermal agitation, and shot effect in the photoelectric pickup and amplifiers. Film which is recorded with a flat frequency response to 6,000 cycles, will have useful response as high as 8,000 cycles. Push-pull recording is used to reduce noise generated by scratches and dirt on the film, and to reduce amplitude distortion.

If 16 mm. motion picture films are recorded at the rate of 30 frames per second for television purposes instead of the standard 24 frames per second, the increase in film speed of from 36 to 45 feet per minute will allow higher fidelity sound recording. For instance, if 6,000 cycles can be recorded successfully on standard 16 mm. film, 7,500 cycles could be recorded if the film were speeded up to 45 feet per minute.

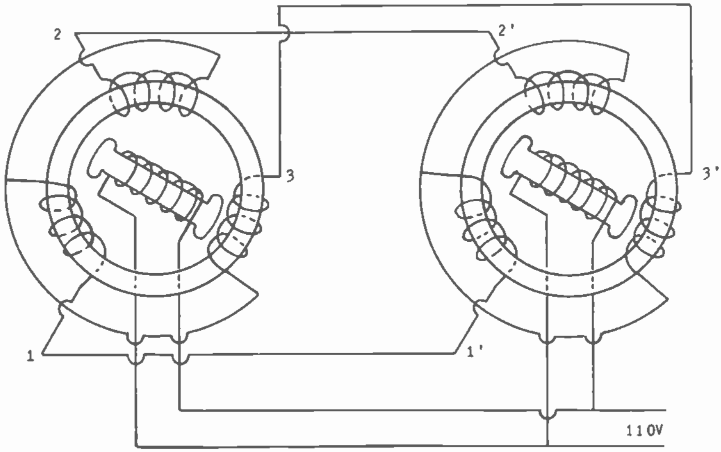


Fig.45 Connection diagram of a pair of Selsyn motors.

The figures quoted are for the best types of film stock and recording equipment. In some comparisons made between 16 mm. motion picture film and the highest quality vertical cut recordings, it was found that the high quality recordings had slightly lower noise and somewhat better frequency response than the 16 mm. film recording used in this particular test. High quality vertical cut records are normally equalized to 6,500 cycles and the response from there on is purposely reduced in the pickup equipment to eliminate needle scratch which predominates above this frequency.

For sound motion picture work, there is, of course, the added difficulty of synchronizing the recording with the motion picture being projected. In one type of projector, this is taken care of by driving the record with a Selsyn motor, which is powered from a similar unit on the motion picture projector. If the record is started in synchronism with the projector, it cannot fall out because of the interlocking properties of the Selsyn motor.

A connection diagram of the Selsyn motors is shown in Fig. 45. One of these motors is connected to the driveshaft of the projector. The other motor is fastened to the driveshaft of the sound turntable. The armatures are powered from a common AC line and the stators are interconnected. The fields set up by the armatures induce voltages in the stator windings of both machines. The voltages induced in one set of windings will buck those induced in the corresponding windings of the other machine. If the voltages generated are equal and opposing, no current will flow between the stator windings of the two machines, and no torque will be generated. However, if one armature is misplaced slightly, as it will be when the projector begins to function, the voltages induced in the stator of the rotating machine will not be equal and opposite to those induced in the stator of the stationary machine connected to the turntable. Current will flow between the armatures, and the turntable motor will take up a new position so that the voltages generated in its stator will buck the voltages generated in the stator of the other machine. Consequently, as long as one armature is rotating, the second armature will also rotate, to keep up with the first. The faster the first armature goes, the faster the second armature will go and

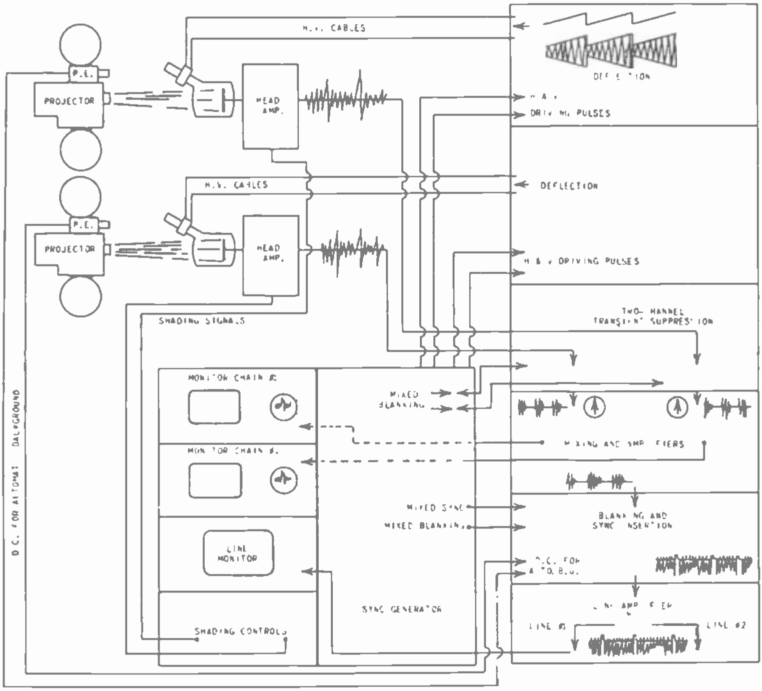


Fig.46 Block diagram of motion picture equipment for television.

they will always stay within a fraction of a turn of the same position. If the two armatures become displaced by as much as 20°, a considerable torque is generated, which prevents them from falling further out of step.

10. ELECTRICAL CIRCUITS NECESSARY FOR THE APPLICATION OF MOTION PICTURE FILM IN TELEVISION. Fig. 46 shows a block diagram of the equipment in a typical television studio with the exception of the signaling controls and the communication system between the motion picture operator and the studio control man. The equipment is practically identical to that required for direct pickup, the only difference being the increased necessity for the transient suppression amplifier and the automatic background control. Two camera chains are shown, although more than two may be incorporated. At least two are necessary for program continuity of films in excess of 2000 feet.

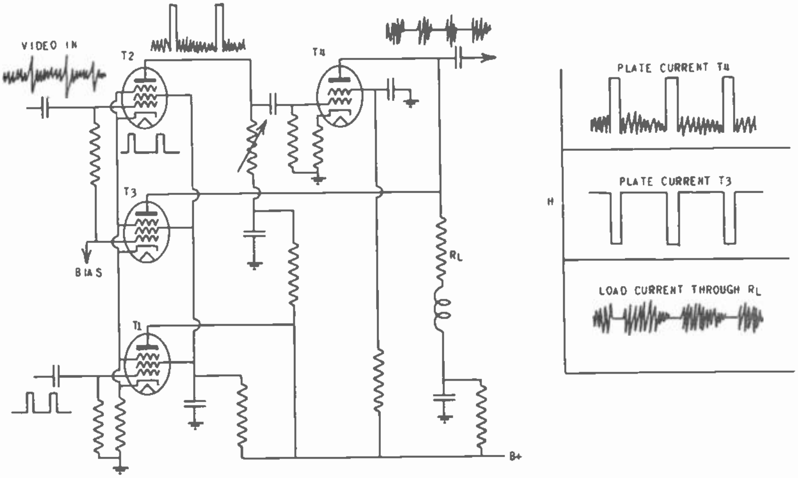


Fig.47 Transient suppression amplifier.

The iconoscopes are mounted upside down in the system shown. This takes care of the fact that the picture is projected right side up, whereas the picture is inverted on the mosaic in direct studio pickup. If the iconoscopes are mounted right side up, the connections to the horizontal and vertical deflection coils must be reversed and the polarity of the keystoning must be inverted. This may not be convenient, especially if identical deflection circuits are desired for the studio and the motion picture cameras.

The circuit for suppressing the transient which is generated during the vertical and horizontal return times was shown in a previous lesson, but it will be repeated briefly here. The need for this circuit is greatest in the case of the storage type of pickup tube. When an iconoscope is used, the picture is flashed on the screen during a blanking interval. The sudden flash of high intensity illumination causes an exodus of photoelectrons from the sensitized plate. These photoelectrons which are emitted from the signal plate result in a violent surge being produced in the video amplifier during the blanking time, notwithstanding the fact that the beam is cut off during this interval by the application of negative blanking pulses to the control grid of the iconoscope. In the absence of a transient suppression circuit, the surge generated would drive some of the amplifiers into grid current and generate a sawtooth shading voltage in the picture signal. Furthermore, any variation in total light from frame to frame causes the surges to be of unequal magnitudes and a bad flicker would result.

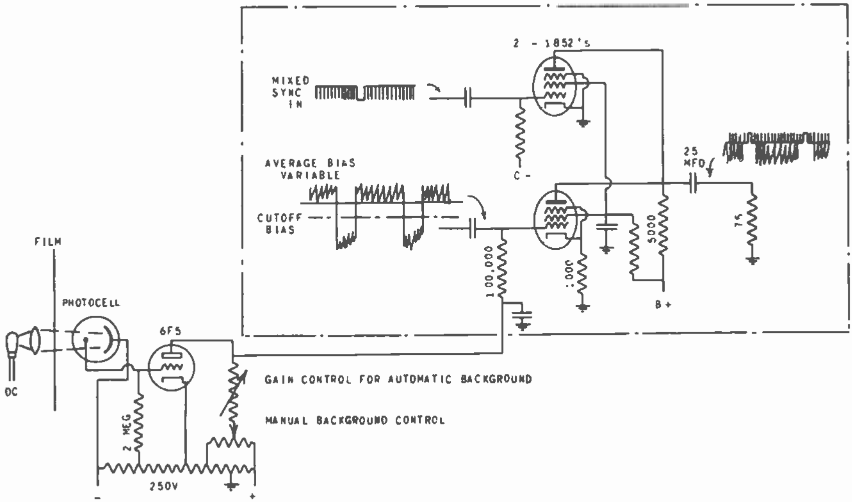


Fig.48 Automatic background insertion circuit.

Fig. 47 shows a transient suppression amplifier in simplified form. When the cathode of tube T_1 is driven positive by a positive blanking voltage on its grid, tubes T_2 and T_3 are cut off, because their cathode is made positive with respect to their grid potential. One of these tubes, T_2, has a normal video signal applied to it. The other tube, T_3, has its grid returned to ground as far as video signals are concerned. Thus, the normal video signal is applied to only one of these two tubes, whereas, the blanking signal is applied to both of them. Tube T_4 acts as a phase inverter to invert the signal from one of the video amplifiers. In the plate circuit of T_4, the inverted signal and the normal signal are combined and cancel out. Therefore, the blanking pulse which drives both of the video tubes to cutoff does not add into the video signal because it is balanced out in the plate circuit of the phase inverter. The waveforms of the picture signal shown, make the operation clear.

The block diagram of Fig. 46 indicated an automatic background control circuit. The details of this circuit are shown in Fig. 48. A photocell is used to generate an output voltage which is proportional to the total amount of light passing through the film, or in other words, proportional to the background of the picture. This output voltage is applied through a one-stage amplifier as a change in bias to a clipper tube in the video chain. The blanking pedestal height is a measure of the background of the picture, and it is set by the bias voltage on the video tube, because the pedestal is generated by driving this tube to cutoff.

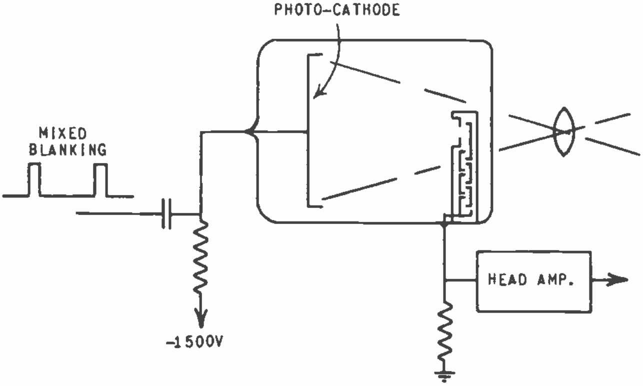


Fig.49 Pedestal Insertion in a Farnsworth image dissector.

In the CBS projector which was pictured in Fig. 44, an automatic background control circuit was not needed. Here it is possible to set up the pedestal height directly in the image dissector tube and thus the background information in the video signal is carried through the entire video chain. This is possible because a simple DC restoration or leveling circuit may be applied at any point in the chain to bring the top of the blanking pedestals to the same level. A circuit for inserting the pedestals is shown in Fig. 49. Positive blanking pulses are injected in the cathode of the image dissector tube. When a cathode is driven positive, the photoemission is suppressed so that during the interval of blanking, no photoelectrons will find their way between the cathode and the first stage of the electron multiplier. This corresponds to the conditions for black, because without illumination on the cathode, there would likewise be no photoemission.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. What is the purpose of the upper and lower film loops in a projector?

2. What alterations must be made in a standard projector to adapt it for television?

3. What is meant by the term "optical rectifier"?

4. What is the purpose of the selector disc in the CBS projector?

5. Sketch the connection diagram of a Selsyn motor.

6. How does the automatic background circuit operate?

7. Would you use an iconoscope or image dissector for the following projectors? Why?
(A) G.E. Projector; (B) Farnsworth Projector; (C) CBS Projector.

8. Why is an intermittent used in a normal motion picture projector?

9. Sketch the selector disc and lens discs used in the Farnsworth projectors and indicate the directions of rotation by arrows.

10. What are some of the comparative advantages and disadvantages of intermittent and continuous motion projectors?

Notes

(These extra pages are provided for your use in taking special notes)

Notes

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Notes

(These extra pages are provided for your use in taking special notes)

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UNIT
NO.
7

PROJECTED TELEVISION

LESSON
NO.
9

PROJECTED TELEVISION

.....group entertainment and education.

Television, comparable in quality and proportions to full-size movies, presents a field for group education and entertainment possessing exceptional possibilities.

Before considering the opportunities in this field of television, it is important that you bear this thought in mind continually. The extent of your success will depend upon your willingness to WORK, your AMBITION, your KNOWLEDGE, and the FUTURE EXPANSION of the industry to which you devote your energy. No matter how hard you work, or how much you know, if the industry is limited, your success will be equally limited. Therefore, it is vitally important that you look into, and consider, the future.

In the years to come, projected television may enable movie theaters in many cities to feature important football games and the world series. Thousands of people, unable to secure tickets to the games, will flock to the theaters, while throngs in distant cities will do likewise. When the colorful crowds in stadiums madly cheer a brilliant play, other crowds, witnessing projected television reproductions of that same play, will become equally excited. And the sponsor of such broadcasts will be able to place a verbal and visual story of his product before them all.

Students gathered in school auditoriums may be able to witness life in the depths of the ocean. Television-equipped diving bells may televise the mysteries of the deep, so that they may be reproduced on screens equal in size to those used in movie theaters today.

We have merely touched upon the future possibilities of projected television. If you are inclined to believe that our predictions are exaggerated, imagine the reactions of the Pilgrims if they were to come to life and visit our world today. The future of the Radio-Television industry is almost limitless.

Lesson Nine

PROJECTED TELEVISION

"While this type of television reception has not become very popular in the United States, still it is finding rather wide use in some foreign nations. Because of this, I felt that our students should have a good working knowledge of how large television pictures are produced.

"This lesson is devoted to a discussion of the problems of obtaining large television pictures. Several of the present day systems for projecting large scale pictures are described".

1. INTRODUCTION. The maximum size picture produced by modern television receivers is not adequate for audiences of over six people. For a 441-line picture, the best viewing distance is between four and eight times the height of the picture. For a 12 inch tube, this means that the best viewing distance is from 30 to 60 inches, as the picture is approximately 7 inches tall. The most convenient viewing distance in the average home is from 4 to 6 feet. This means that the picture should be around 6 to 9 inches tall. In order to have adequate entertainment value for a large audience, a television reproducer must project a picture comparable in size to the average motion picture.

2. THE PROBLEMS INVOLVED. One method of obtaining larger pictures for home receivers is through the use of larger cathode ray tubes. It is impractical both economically and physically to build cathode ray tubes with screens larger than the 12 inch tubes that are now available in this country. However, television cathode ray tubes with screen diameters of 15 inches are made in some foreign countries. These tubes have very thick walls as the external crushing force due to the air pressure is more than five tons. The force on the face of the tube alone is approximately one and one-quarter tons. In order to make the face of the tube reasonably flat, the overall length of the tube, even when mounted vertically, requires a rather large cabinet. Also, the tubes must be made long in order to obtain adequate deflection of the electron beam with the output power tubes available. The manufacturing cost and possibility of breakage increase with the size of the cathode ray tube.

Another solution that immediately suggests itself is to use a small cathode ray tube and project the picture on a screen with a lens. This does not work out very successfully with ordinary cathode ray tubes, as the illumination level of the projected picture is too low for viewing. The low light level causes eyestrain. Also, the room must be completely darkened in order to see the picture at all.

The intensity of the fluorescence of a cathode ray tube screen can be increased by increasing the accelerating voltage and the beam current. Cathode ray tubes suitable for projection purposes will be described a little later in the lesson. Even with these specially designed projection type cathode ray tubes, the intensity of the picture is lower than is desirable.

Perhaps the most satisfactory solution to the problem is to use a high intensity light source, such as a carbon arc, and modulate the light beam with a light valve controlled by the picture signal. The modulated light is then projected on a screen. The modulated light beam must be deflected both vertically and horizontally to re-assemble the picture elements to form a replica of the transmitted picture.

Light valves are devices used to control the intensity of a light beam by means of an applied picture signal. The light beam from the source is directed through the light valve by means of a lens system. A light valve suitable for television picture reproduction must be able to follow faithfully the variations in the picture signal, that is, it must have a flat frequency response from zero to the maximum frequencies present in the picture signal. Also, the power required to operate the light valve should not exceed the power output of conventional output tubes when working into the low values of load resistance required for the wide frequency range of the television signal. The light transmission efficiency of the light valve must be high, that is, the light flux transmitted by the light valve for white in the picture signal should be very nearly equal to the light flux incident on the light valve.

Many types of light valves have been developed. Mechanical light valves are used in facsimile (described in the next lesson) and in recording sound on film. In these valves a light metallic ribbon or reed is moved more or less across an aperture by a magnetic field. The exciting voltage for the field is supplied by the picture signal. Mechanical light valves are very satisfactory where the frequency range of the applied signal covers only audio frequencies, but they cannot be used in television.

Light valves that have been found suitable for television are based on the change in the light transmission characteristics of certain mediums by the presence of electric and magnetic fields and by the passage of supersonic waves through the medium. (Supersonic waves exist in a substance when the particles of the substance vibrate mechanically with frequencies in the radio spectrum). A more complete description of these light valves will be given later in the lesson.

The only methods that have been found practical in deflecting the modulated light beam are mechanical in operation. The first scanning methods used in television were mechanical. Early television transmitters and receivers used the Nipkow scanning disc as described in a previous lesson. The Nipkow disc cannot be used for modern high-definition television because of the extremely low light transmission efficiency for discs with diameters that can be used conveniently. Also, the cost of building a high-definition Nipkow disc is prohibitive. In overcoming the defects of the Nipkow disc, several mechanical scanning systems have been developed which are quite satisfactory for modern high-definition television picture reproduction in conjunction with a modern light valve. One of these will be described a little later in this lesson.

The synchronizing generators used with modern electronic television systems are controlled by the power line frequency. The control frequency for the horizontal synchronizing and blanking impulses, and the equalizing impulses, is obtained either through multiplication of the power line frequency or by automatic frequency control of a master oscillator by the power line frequency. There is a variable phase shift of the horizontal frequency with respect to the control frequency in both these systems. This phase shift does not cause trouble when inertialess electronic deflecting circuits are used. The deflecting circuits at the receivers can follow faithfully the slight variations in timing of the synchronizing impulses, and there is no distortion in the reproduced picture. However, mechanical scanning systems have inertia and cannot follow the small rapid variations in timing of the horizontal synchronizing impulses. In those countries where mechanical scanners are used for modern high-definition television picture reproduction, the synchronizing control frequency is generated by mechanically driven generators. Such a generator will be described later in this lesson.

One reason for the low light intensity in television pictures is due to the fact that only one picture element is reproduced at a time, and the retentivity of the eye integrates all the separate picture elements into the complete picture. With cathode ray tube receivers, this condition is partially overcome by the light radiated through phosphorescence after the scanning beam has left a picture element. One of the modern light valves transmits several picture elements simultaneously and the picture brilliance is increased many times.

In the remaining part of this lesson are described the principal methods that have proved successful for the production of large size television pictures. The only one used in this country at present projects a magnified image of the picture produced on the screen of a specially designed cathode ray tube. This method will be described first.

3. **PROJECTION CATHODE RAY TUBE.** There are two ways in which the brilliance of the fluorescence can be increased in a cathode ray tube. These are: (1) increasing the beam current, and (2)

increasing the second anode voltage. In the ordinary electron gun, most of the electrons emitted from the cathode are collected by the first anode. Thus, if more of the cathode emission can be forced into the electron beam to the screen, the brilliance of the picture will be increased many times. Increasing the second anode voltage increases the velocity of the electrons in the beam, and therefore, their kinetic energy. The higher speed electrons will increase the light emission of the fluorescent screen.

The operation of the ordinary type of electron gun was described in Lesson 6 of Unit 5. The essential parts of a simple electron gun are the cathode, control grid, the first anode or focusing electrode, and the second anode or accelerating electrode. The purpose of the control grid is to vary the number of electrons in the beam. The combination of the cathode, control grid, and first anode, form a crossover or concentrate the emission from the cathode into a small spot in front of the aperture to the first anode. The combination of the first and second anodes produces an image of the crossover on the fluorescent screen of the tube. The first anode limits the divergence of the beam from the crossover so that the electrons pass through the center of the electrostatic lens formed by the first and second anodes. The divergence of the electron beam is limited by the collection of all the electrons outside of the desired limits by the first anode. Most of the electrons coming from the cathode are thus removed from the electron beam.

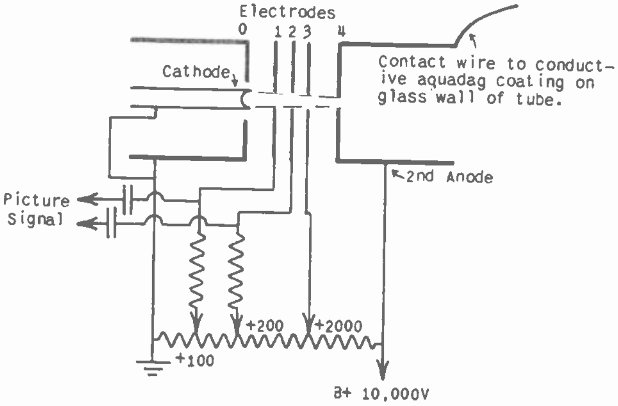


Fig.1 Construction of electron gun in projection type cathode ray tube.

Then, in order to increase the density of the electron beam reaching the screen, the design of the cathode, control grid, and first anode structures require modification. Fig. 1 shows the construction of an electron gun that forces more of the electrons into the beam. All of the sections are operated at a positive potential with respect to the cathode. This structure produces a very small crossover and confines the limits of the divergence of the electrons by forcing them to pass through a smaller crossover. The actual size of the crossover is determined by the aperture in the final or second anode of the gun structure. In this gun, the total accelerating voltage is utilized in forming the crossover, and the velocity of the electron through the crossover is so high that their tendency to diverge beyond the crossover is reduced.

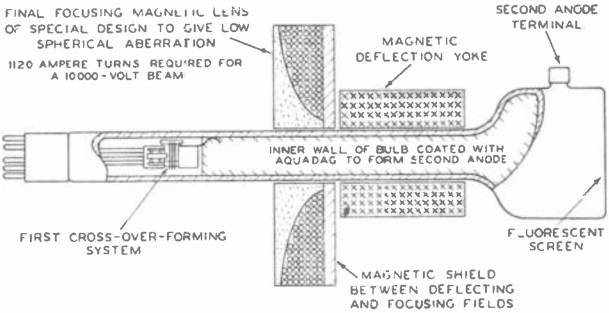


Fig.2 Projection cathode ray tube.

Since the total accelerating voltage has been used to force the electrons through a small crossover, there is no potential available to form an electron lens to form an image of the crossover on the screen. Then another type of lens system must be used to form an image of the crossover on the screen. Fig. 2 shows a complete schematic of a projection cathode ray tube. The inner coating forms part of the second anode. The second anode voltage is 10,000 volts. The beam is modulated by applying the picture signal to the electrodes 1 and 2 in Fig. 1. The image of the crossover is formed by the magnetic focusing coil around the outside of the tube. The aperture of a magnetic lens produced by an external focusing coil is much greater than the aperture of an internal electrostatic lens. The aperture of the internal lens is limited by the diameter of the neck of the tube, which in turn is limited by the available power of the magnetic deflecting circuits. The advantage of a large aperture lens is the reduction in spherical aberration as the electron beam passes through a smaller section of the lens aperture.

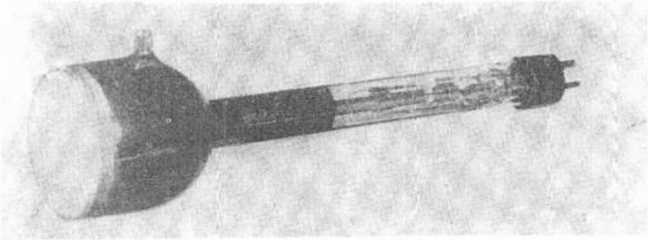


Fig.3 Projection cathode ray tube.

Fig. 3 is a picture of an RCA projection type cathode ray tube. Fig. 4 shows the tube mounted in an experimental projection unit.

The picture shows very clearly the magnetic focusing coil, the magnetic deflecting yoke, and the lens system for projecting the picture on a screen. Fig. 5 is a modern cathode ray projection unit. The power supplies for the tube, the deflecting circuits, and the synchronizing separator circuits are located in the cabinet. The unit will project a satisfactory picture on a 3 × 4 foot screen.

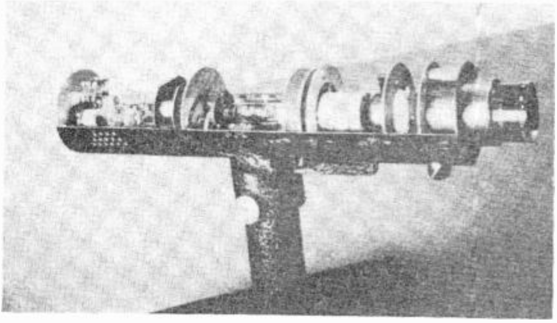


Fig.4 Projection tube mounted in housing.

In order to obtain the most brilliant picture possible, the picture is projected on a directional screen. A directional screen is one that reflects most of the light back toward the projector. An observer located at the side of the screen will not be able to see the picture very well, while an observer in front of the screen will see a brilliant, sharply defined picture. Directional screens have a surface similar to tracing cloth.

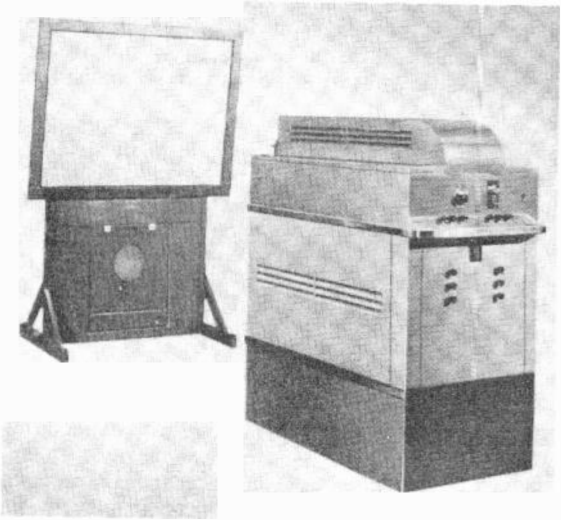


Fig.5 Projection cathode ray tube receiver.

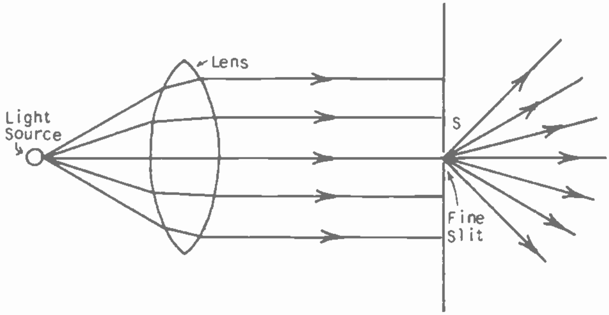


Fig.6 Light rays diverging from slit.

Projection cathode ray tubes have been developed using second anode voltages as high as 30,000 volts. Baird in England has built a cathode ray projection unit that will produce a 6 × 8 foot picture of sufficient brilliance to prevent eyestrain.

The next few sections of the lesson will be devoted to a discussion of the Scophony system for producing large scale television pictures. At present, Scophony receivers are available only

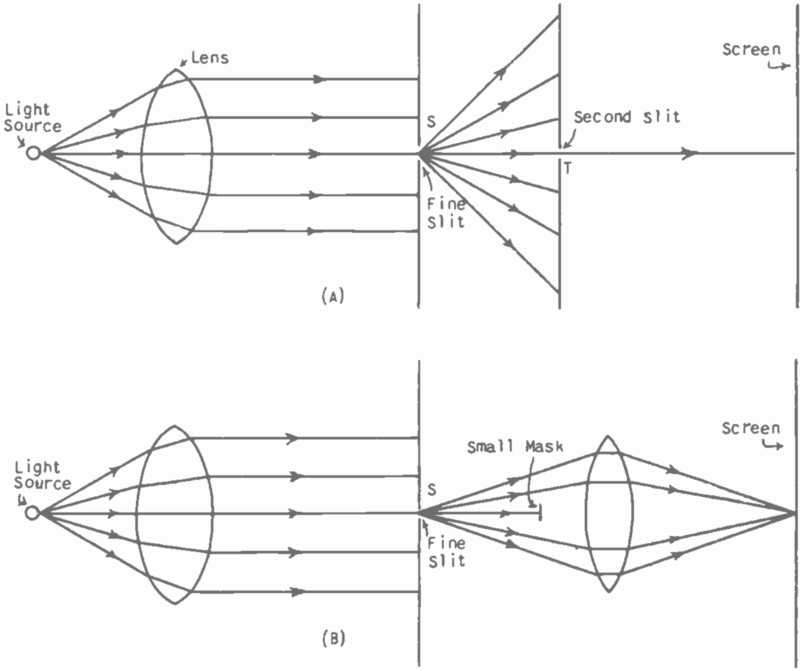


Fig.7 Simple light valves based on variable divergence of light.

in England, but it is probable that they will be available in this country shortly. This system uses a light valve in conjunction with a mechanical scanning system.

4. SUPERSONIC LIGHT VALVE. The heart of the Scophony system is the supersonic light valve. Before going into a description of the supersonic light valve, it will be necessary to describe another property of light. This is called "diffraction". Diffraction is the bending of light when it passes by an edge or corner. Whenever light passes through a very fine slit, the slit acts as a source of light, and the light leaving the slit is diverging. This effect is illustrated in Fig. 6.

If the diffraction of the light passing through the slit can be controlled, it provides a method of operation for a light valve. A simple light valve based on this principle is shown in two forms in Fig. 7. If a second slit is placed in line with the first slit (Fig. 7A), the light diverging from the first slit is cut off and only the light in the center of the beam coming from the first slit passes through the second and reaches the screen. By varying the divergence of the light passing through the first slit from zero to maximum, the light passing through the second slit will be maximum for zero divergence and minimum for maximum divergence. If the light in the central beam is cut off by a small rod with the same dimensions as the second slit (Fig. 7B), the light reaching the screen will be completely cut off when the divergence is zero, and will be maximum when the divergence is maximum.

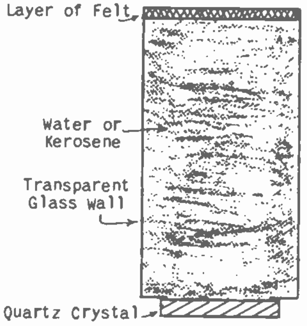


Fig.8 Construction of supersonic light valve.

The operation of the supersonic light valve is identical in operation to the simple light valve described in the preceding paragraph. The extent of the divergence of the light is controlled by the picture signal.

Fig. 8 is a schematic of a supersonic light valve. It consists of a transparent glass container filled with a liquid such as water or kerosene. At one end of the container in contact with the liquid is a quartz crystal. At the other end, opposite the crystal, is a small layer of felt. The crystal is excited into oscillation by a conventional crystal oscillator circuit. The vibrating crystal is in contact with the liquid and radiates high-frequency waves into the liquid. The waves travel to the other end of the cell where they are damped out by the layer of felt. These waves are known as supersonic waves. They have the same frequency as the crystal.

Fig. 9 shows the effect on the liquid when the supersonic waves are traveling through it. The particles of the liquid vibrate in the same direction that the wave motion is moving. When the crystal expands, it compresses the liquid next to it. This area of compression or condensation is transferred to the adjacent particles in the liquid and is transmitted through the length of the liquid. When the crystal contracts, the liquid next to it expands, and this expansion or rarefaction follows the condensation through the liquid. Thus, the supersonic waves consist of alternate condensations and rarefactions traveling through the liquid. The density of the liquid is greater than normal in a condensation, and less than normal in a rarefaction. The speed of the light passing through the condensations will be much less than the speed of the light passing through the rarefactions.

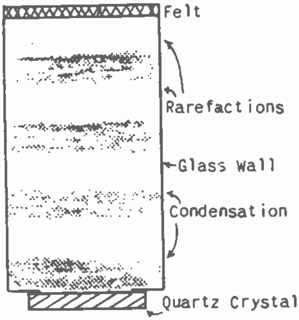


Fig.9 Condensations and rarefactions produced by vibrating crystal.

When there are supersonic waves traveling through the liquid and a beam of light is passed through the liquid perpendicularly to the direction of the movement of the waves, the rarefactions of the supersonic waves act as fine slits. The rays of light passing through the rarefactions or slits, diverge. Thus, the supersonic cell consists of a large number of fine parallel slits. The magnitude of the divergence depends on the ratio of the densities of the condensations and rarefactions. This ratio varies directly with the amplitude of the supersonic waves. If the supersonic waves are amplitude modulated with the picture signal, the divergence of the light will be proportional to the amplitude of the picture signal. Since the supersonic waves are generated by the vibrating crystal, the divergence of the light can be controlled by amplitude modulating the output of the crystal.

For a modern high-definition picture, the crystal frequency must be at least 20 megacycles in order to handle the picture sidebands without attenuation. When a crystal oscillates in the air, it has a very high Q, as the damping of the air is negligible. However, when the crystal oscillates in contact with the liquid, the damping is sufficient so that the resultant Q is low enough that the picture sidebands are passed without excessive attenuation for a carrier frequency of 20 megacycles.

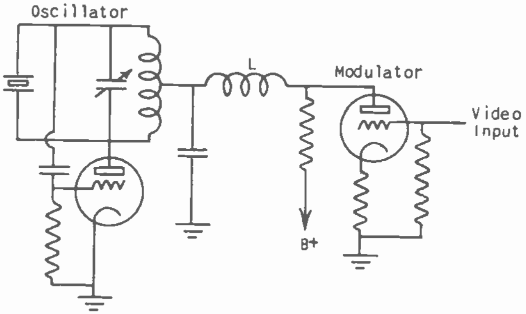


Fig.10 Circuit of oscillator and modulator suitable to drive the crystal in a supersonic light valve.

Fig. 10 shows a schematic of an oscillator and modulator that can be used to drive the crystal in a supersonic light valve. In this circuit a Hartley oscillator is used. The voltage developed across the oscillator tank circuit is applied across the crystal. The crystal is mounted so that its vibration will occur along the smallest dimension of the crystal. The oscillator is plate modulated by the picture signal. Plate modulation can be used, as the power requirements are small. The constants of the oscillator are selected so that the capacity loading on the modulator will be a minimum. The choke L, keeps the oscillator frequency from the modulator and corrects the high-frequency response of the modulator (series peaking).

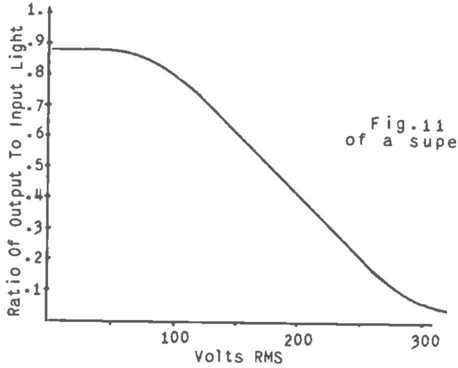


Fig.11 Characteristic curve of a supersonic light valve.

Fig. 11 shows the relation between the applied crystal voltage and the light output of a supersonic light valve. The data was plotted for a valve using the principle shown in Fig. 7A. The ratio of the output to the input light flux is plotted along the vertical axis of the graph. The maximum output of the cell is limited by the light transmission efficiency of the liquid in the supersonic light valve. From the graph, it is evident that the light beam cannot be completely extinguished to obtain black. However, if the method illustrated in Fig. 7B is used, the output will be zero when the voltage applied to the crystal is zero.

For kerosene, at room temperatures, the 20-megacycle supersonic waves have a length of approximately .065 millimeter. The velocity of the wave train is about 1300 meters per second. Since the number of picture elements transmitted in one second by a modern television system is approximately 8 million, the amplitude change of the supersonic waves for a single picture element will last over about two and one-half cycles. If the distance in the supersonic light valve between the crystal and the damping layer of felt is 10 millimeters long, the information contained in approximately 62 picture elements will occur simultaneously in the light valve (10÷2.5×.065 = 62). Thus, if the optical and scanning systems are properly designed, several picture elements can be projected simultaneously on the screen. This will result in a marked increase in the brightness of the picture.

5. MODERN MECHANICAL SCANNERS. In order to reproduce a picture, the output light beam of the light valve must be deflected both vertically and horizontally to reassemble the picture elements into a replica of the original. The Nipkow scanning disc cannot be used because of its low optical efficiency. Each hole in the disc would pass only a very small fraction of the available output of the light valve. However, the beam can be deflected with mirrors, and a high optical efficiency is obtained.

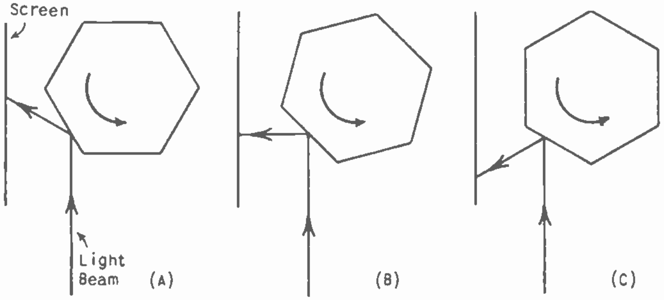


Fig.12 Deflection of a light beam by a rotating polygonal drum.

The light beam is deflected by the method illustrated in Fig. 12. The light beam is directed on a rotating polygonal drum. The surfaces of the drum are plane mirrors. When the drum rotates in a counter-clockwise direction, the light beam is first deflected upward as a new surface of the drum comes before the light beam. As the drum continues to rotate, the deflection of the light beam changes as shown in the figure. When the drum surface is in the position shown in Fig. 12C, the light beam is deflected downward. As the drum continues to rotate, the light beam goes off the edge of one surface and onto the next adjacent surface. When the light beam changes surfaces, it immediately jumps up to the starting point as shown in Fig. 12A. Thus, the light beam is deflected slowly downward and rapidly upward as the beam in a cathode ray tube. However, with a rotating drum, the return time is practically instantaneous.

Horizontal deflection of the light beam can be obtained by directing the light beam on another rotating polygonal drum whose axis is perpendicular to the axis of the vertical drum. Usually, the beam is deflected horizontally first, and the beam reflected from the horizontal drum is then deflected vertically by the second drum. Since the horizontal deflection will be along the axis of the vertical drum, the vertical drum has to be longer than the horizontal deflection at that point. A complete schematic of a mechanical deflecting system will be shown a little later in the lesson.

The vertical drum must deflect the light beam at the field frequency rate, and the horizontal drum must deflect it at the line frequency rate. If the vertical drum has six surfaces, the beam will be deflected vertically six times for one rotation of the drum. Then the drum speed will have to be 600 r.p.m. in order to deflect the beam 60 times per second. For a drum with twelve surfaces, the speed will have to be 300 r.p.m., etc. Since the horizontal deflecting frequency is 13,230 cycles, both the number of surfaces and the speed must be increased. For a drum with 30 surfaces, the speed will have to be 26,460 r.p.m.

Since the horizontal drum has a very high speed, it must be made as small as possible in order to reduce the driving power. The driving power increases very rapidly as the speed of rotation is increased. The drum speed could be reduced by increasing the number of reflecting surfaces, but for a given size drum, this results in lowered optical efficiency, as the reflecting surfaces are smaller. Also the adjustment would be more critical with the smaller surfaces.

Each drum is driven by a separate synchronous motor. The motors are controlled by the synchronizing pulses transmitted with the picture signal.

6. CYLINDRICAL LENSES. The operation of the Scophony system is dependent on the use of cylindrical lenses. In Lesson 3 of Unit 5, ordinary lenses were discussed. An ordinary lens is a spherical lens. The principal focus of a spherical converging lens was defined as the point through which light rays from a distant source pass after refraction by the lens. The principal focus of a cylindrical lens is defined as the "line" through which parallel rays from a distant source pass after refraction by the lens. This line is parallel to the axis of the cylinder.

Fig. 13 shows the formation of the image for a cylindrical lens. The object is the figure ABCD, and the image is the rectangle

A'B'C'D'. A cylindrical lens refracts or bends only the light rays that are in planes perpendicular to the axis of the cylinder. The lines D'A' and C'B' are the images of the lines AD and BC respectively, and D'C' and A'B' are the images of the lines DC and AB respectively. The images of AD and BC are inverted and smaller than their respective objects, while D'C' and A'B' are identical in size and position as their respective objects. Thus it is evident that a cylindrical lens focuses light rays in one plane only.

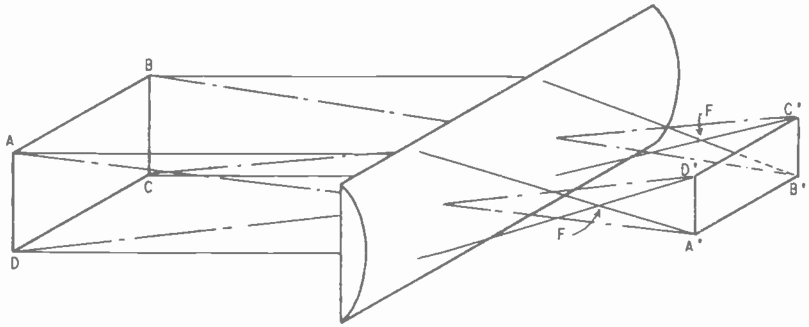


Fig.13 Formation of an image by a cylindrical lens.

7. THE SCOPHONY SYSTEM. Fig. 14 shows schematically the operation of the Scophony system for producing large scale television pictures. The lines representing the paths of the light rays through the optical system consist of two types, solid and broken lines. The solid lines represent the paths of the light rays that are bent in one direction by the cylindrical lenses L_1, L_5, and L_6. The broken lines represent the rays that are bent in the perpendicular direction by the cylindrical lenses L_2 and L_3. The two crossed arrows, labeled plane 1 and plane 2, at the bottom of Fig. 14 show the directions of these two planes of focusing. The lens L_4 is a combination of two cylindrical lenses of different focal lengths. The focusing planes of the two lenses are perpendicular.

The light from the source leaves the slit S in a diverging beam with a rectangular cross section. The long dimension of the rectangle is horizontal. The lens L_1 focuses the rays diverging horizontally so that they cross in the light valve. The rays diverging vertically, pass through L_1 unchanged. The light beam, as it enters L_2, has a rectangular cross section, but the long dimension of the rectangle is perpendicular to the long dimension of the slit, S. The lens L_2 converts the rays diverging vertically into a parallel beam, but it has no effect on the light rays converging horizontally. The cross section of the light beam in the center of the light valve will be a short vertical line.

The light valve is mounted so that the supersonic light waves travel upward. Therefore, the slits formed by the rarefactions are horizontal. As stated before, for a distance of 10 millimeters between the crystal and the damping felt layer, the information contained in approximately 62 picture elements of a scanning line, was present simultaneously in the valve. Therefore, the light beam leaving the light valve will be a replica of a section of a line of the original picture 62 elements long. The number 62 was selected arbitrarily in order to illustrate that several picture elements can be projected simultaneously by the supersonic light valve. Theoretically, a much greater number could be projected simultaneously, but the actual number is limited by conditions that will be explained later.

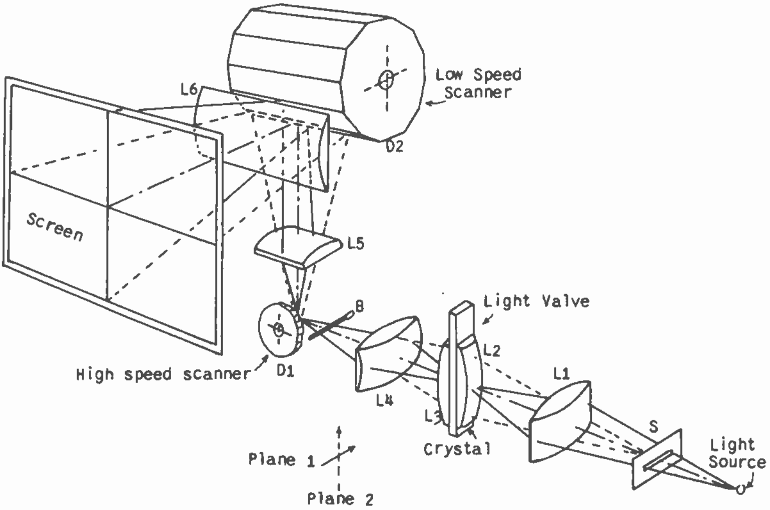


Fig.14 Optical arrangement of Scophony system.

The next problem is to separate the diverging and undiverging light from each slit. As stated before, it is better to block out the central beam from the slit in order to get a good black. The light that reaches the screen is the light that is diffracted or diverged and passes by the mask that blocks out the central beam (See Fig. 7B.).

The rays of light leaving the light valve are parallel in the vertical plane. The cylindrical lens L3 causes these parallel rays to converge and pass through a line at B. Thus, the central beams from all the slits pass through the line at B. If a small mask is placed at B with the same area and shape as the cross section of the central beams, all of them will be blocked out. Therefore, only the light diverging from the central beams from each slit will continue on to the screen. The diverging light rays from the light valve in the horizontal plane are focused on the horizontal deflecting drum by the lens L4. Lens L4 also converts the light rays in the vertical plane slightly. The light rays in the vertical plane are diverging when they reach the horizontal deflecting drum.

The image projected on the horizontal deflecting drum consists of a section of a line of the picture several picture elements long. The line lies in the plane of the drum. The rays are reflected upward by the mirror surfaces on the horizontal drum. The rays are represented by the broken lines which lie in the plane of a scanning line, and are reflected parallel to the plane of the horizontal drum. The rays represented by the solid lines and which determine the width of the scanning line, are reflected parallel to the axis of the horizontal drum. The light rays in both planes are diverging when they leave the horizontal drum.

The light rays from the horizontal drum are reflected to the screen by the vertical deflecting drum. The rays which are represented by the broken lines, and which lie in the plane of the scanning lines, reach the screen without change by the cylindrical lenses L_5 and L_6. Therefore, the lens L_4 focuses the rays in the plane of a line on the screen. The lens L_5 converges the rays which represent the width of a line in order to keep the size of the reflecting surfaces on the vertical drum within reasonable limits. Lens L_6 focuses these rays representing the line width on the screen.

As stated before, the information corresponding to a section of a line of the picture exists simultaneously in the supersonic light valve. At the beginning of a line the wave train leaving the crystal will have a modulation percentage corresponding to the first element in the scanned line at the transmitter. As the wave train progresses through the cell, it is followed by wave groups with modulation percentages corresponding to the succeeding elements in the line. Each wave group travels through the valve to the layer of felt at one end of the valve and is damped out.

The question may arise in the reader's mind why the movement of a wave group for a picture element through the light valve does not cause a corresponding movement of the picture element on the screen. Let us consider the line of light that moves across the screen as made up of a series of light beams, and that each has a cross sectional area equal to a picture element. Also, let us consider that the light valve has a series of apertures arranged in a line along the length of the valve and each has an area equivalent to a picture element. Each of the light beams that reach the screen pass through one of the apertures in the light valve. Each of the light beams is deflected across the screen at the line frequency rate.

When the wave group corresponding to the first element in a line is radiated into the liquid of the light valve, it will be before the first aperture. The light beam passing through that aperture will be modified according to the shade of gray represented by the first picture element. At this instant, the light beam passing through the first aperture will fall on the screen at the begining of a line in the reproduced picture. During the interval that this first light beam is deflected across the screen, a distance equal to a picture element, the wave group corresponding to the second element in the line will have been radiated into the liquid and will be before the first aperture in the light valve. Thus, when the light beam has moved into the position of the second picture element in the line on the screen, the wave group corresponding to the second picture element is before the aperture of the light beam in the light valve.

Since the light beams follow each other in succession across the screen, the light beam passing through the second aperture in the light valve comes on the screen in the position of the first picture element of the line at the instant that the wave group for the first picture element passes in front of the second aperture in the light valve. Thus, as each wave group progresses through the light valve before the series of apertures, its corresponding shade of gray is projected on the same spot on the screen by the progressively moving light beams.

The total number of picture elements that can be projected simultaneously by means of the supersonic light valve, ranges from 150 to 200. There are two principal factors which limit the number of picture elements. One is the variation in the velocity of the supersonic waves with temperature. The other is the attenuation in the amplitude of the supersonic waves as they progress through the cell.

Variations in the velocity of the supersonic waves change the complete number of waves or slits that can occur simultaneously in the cell. The wavelength is equal to the ratio of the velocity to the frequency. Therefore, an increase in velocity is accompanied by an increase in wavelength. Then, in order to maintain the same number of waves or slits in the supersonic light valve for normal temperature changes, there is a definite limit to the maximum length of the light valve that it is practical to use.

The attenuation of a wave or group of waves as they travel through the light valve will have the same effect as a reduction in the percentage of modulation as far as the projected picture element is concerned. If a wave train starts from the crystal with the percentage of modulation corresponding to a white picture element, the attenuation will cause the projected picture element to become darker as the wave train travels through the light valve.

8. SYNCHRONIZATION. The deflecting drums in mechanical scanning systems are usually maintained in synchronism with the transmitter by the use of synchronous driving motors. These motors are operated on the amplified synchronizing impulses that are transmitted along with the picture signal. In the case of the slow speed vertical drum it is possible to develop sufficient power from the received synchronizing impulses to start the driving motor and bring it up to speed. However, in the case of the high speed horizontal drum, it is necessary to bring the drum up to speed by a conventional induction motor and then maintain synchronism by a synchronous motor operating on the amplified horizontal synchronizing impulses removed from the picture signal.

It was mentioned earlier in the lesson that it is impractical to try to synchronize a mechanical scanning system by means of the impulses generated by the ordinary electronic impulse generator.

The frequency of the horizontal synchronizing and blanking impulses is often controlled by comparing the generated frequency to the power line frequency by methods similar to those used to control the oscillator frequency in superheterodyne receivers having automatic frequency control. With this system of control, the generated horizontal frequency tends to oscillate slightly about a mean value. This does not cause any trouble with inertialess electronic deflecting circuits used with Iconoscope and cathode ray tubes, as the sawtooth oscillators will be under perfect control of the synchronizing impulses. However, synchronous motors and phonic wheels will not be able to follow the small variations in the timing of the synchronizing impulses, because the inertia of the rotating drums and motor armature. Therefore, there is a loss in detail of the picture reproduced by mechanical methods when synchronized by impulses whose frequency is controlled electronically.

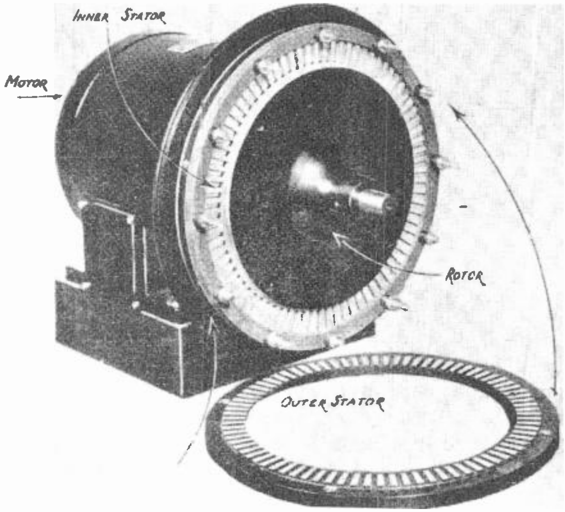


Fig.15 Scophony horizontal frequency generator.

Fig. 15 is a picture of a mechanically driven electrostatic generator of high stability that has been developed by Scophony for generating the horizontal control frequency. A disc, with teeth around the circumference, is rotated between two rings containing teeth on the inside. The disc is driven by a synchronous motor operating on the power line frequency. As the disc rotates, the capacity between the rotor and stator rings is maximum when the teeth on the disc are in line with the teeth on the rings, and is minimum when the teeth on the disc are in line with the spaces between the teeth on the rings. Fig. 16 shows the circuit of the generator. The rotor is connected to ground. The two stators are connected to high voltage DC sources of opposite polarity through the two inductances. The two inductances and the variable condenser form a parallel tuned circuit across the stators of the generator.

There will be a current flow in and out of the condensers, formed by the stators and the rotor, as the capacity is varied through the rotation of the rotor. The currents through the two inductances will be in phase because the two stators have DC potentials of opposite polarity. If the parallel tuned circuit is resonant to the frequency of the current changes, the output of the generator will be maximum.

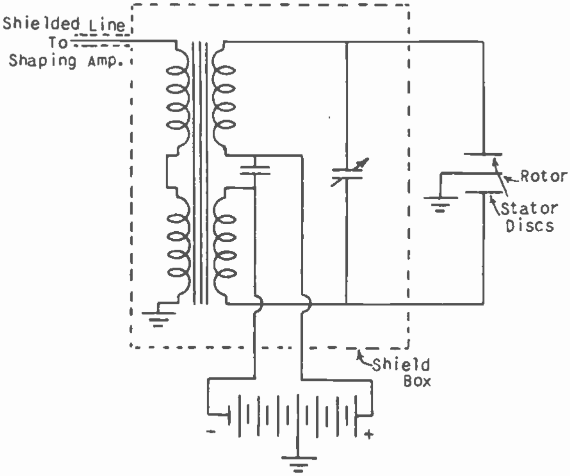


Fig.16 Circuit of Scophony generator.

The output is in the form of a sine wave. The frequency stability is very high, because all of the teeth on both rotor and stators contribute to the generation of each cycle. In this way, mechanical imperfections cancel out. The output is converted into the synchronizing and blanking impulses by conventional shaping circuits.

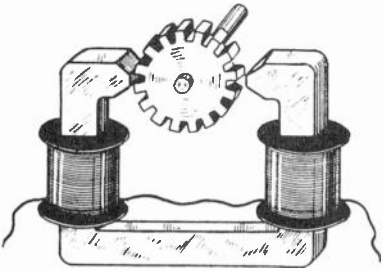


Fig.17 Phonic wheel.

9. PHONIC WHEELS. Fig. 17 shows schematically the construction of a synchronous motor or phonic wheel, suitable for driving the deflecting drums of a mechanical scanning system. The motor consists of a slotted disc rotating between the poles of a magnetic

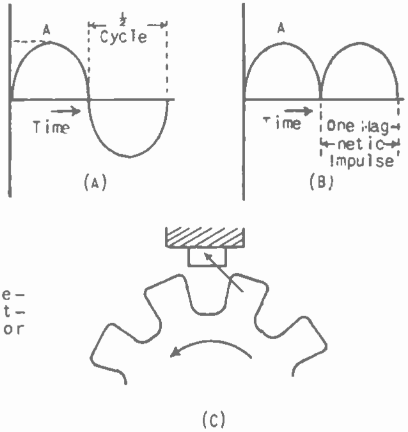


Fig. 18 Relations existing between field current, magnetic attraction, and position of rotor for a phonic wheel.

field. The rotor is made up of soft iron laminations. Rotation is produced by the interaction of the field of the electromagnet, and the field induced in the rotor. The simple phonic wheel, as illustrated in Fig. 17, is not self-starting. Also, it can be made to rotate in either direction. The motor in an electric clock is an example of the simple type phonic wheel.

The operation of the phonic wheel is best explained in terms of Fig. 18. The speed of rotation is such that two teeth of the rotor pass by each pole of the field coil for one cycle of the AC current through the field. Fig. 18A shows one cycle of the AC current through the field. Fig. 18B shows the corresponding variations in magnetic attraction between the field of the electromagnet and the induced field in the rotor. The force between the two fields is always attractive, as the induced field through the rotor is a continuation of the field between the poles of the electromagnet. Assume that the speed of the rotor has been brought up to the syn-

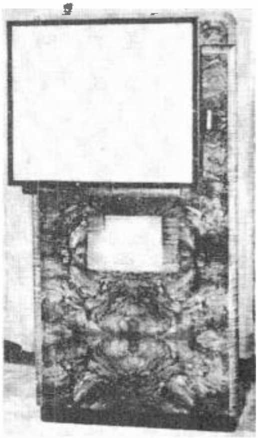


Fig. 19 Scophony home television receiver.

chronous speed by an external agency and the rotor is in the position shown in Fig. 18C. At that instant, the magnetic attraction is maximum, and in the direction indicated by the arrow between the pole and the approaching tooth. As the tooth approaches the pole, the current through the field, and the magnetic attraction, become less. As the tooth passes under the pole, the current passes through zero and reverses. At the same instant, the magnetic attraction is at its minimum. The inertia of the rotor carries the tooth beyond the pole. The residual magnetism left in the tooth, and the magnetic field produced by the reversed current, repel until the residual magnetism has been reduced to zero by the opposite induced field. However, there is an attractive force between the pole and the induced field in the next approaching tooth, and this force is maximum when the pole is midway between the two teeth. The magnetic attraction reaches a minimum, and the current again reverses as the second tooth passes under the pole.

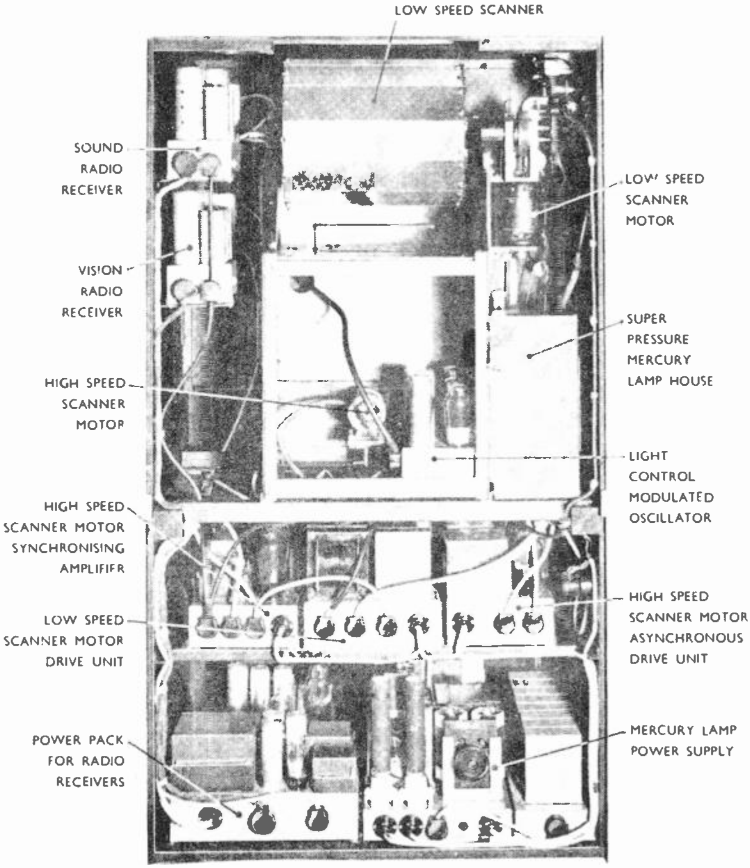


Fig.20 Rear view of Scophony receiver.

Two teeth pass by the field poles for each cycle of the applied AC voltage. Therefore, the speed of the phonic wheel in revolutions per second is equal to the ratio of the applied frequency, to half the number of teeth on the rotor.

Phonic wheels can be designed so that they are self-starting. Since an explanation of the method used to make them self-starting is not necessary for an understanding of projection television it has been omitted.

10. THE SCOPHONY TELEVISION RECEIVERS. Fig. 19 is a picture of a recent model Scophony home receiver that uses the Scophony system for producing a large picture. The picture is projected on the back of the ground glass screen. The picture size is 24 by 20 inches. Fig. 20 shows a rear view of the receiver. Fig. 21 shows a picture of the scanning and light modulation unit. Fig. 22 shows the essential parts of the combination induction motor and phonic wheel used to drive the high-speed deflecting drum in the receiver.

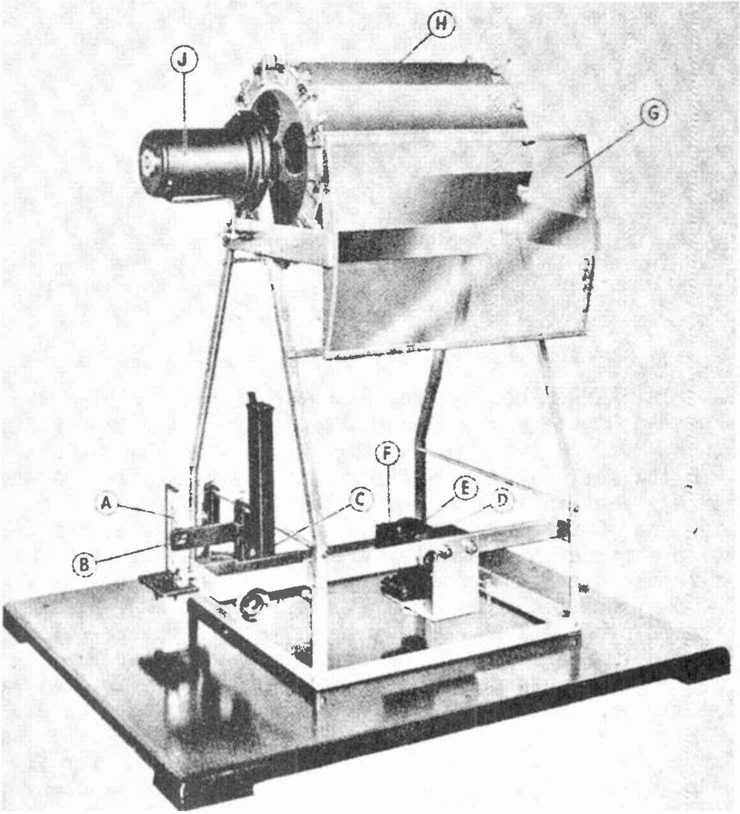


Fig. 21 Scanning and light modulation chassis.

Fig. 23 is a picture of a Scophony television projector suitable for small theaters. The picture is projected on the rear of a ground glass screen. The picture size is 6 by 5 feet.

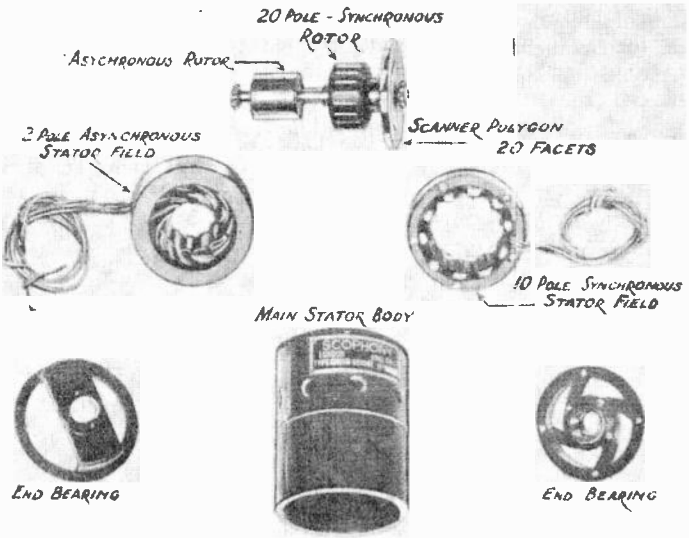


Fig.22 Parts of driving motor for high speed drum.

11. THE KERR CELL. There is another type of light valve that has been used for the projection of large scale television pictures. This is the Kerr cell. Before going into a discussion of the operation of the Kerr cell, it will be necessary to discuss another property of light that is called "polarization".

Light is a transverse wave motion; that is, the particles of the medium through which the light wave is passing, vibrate perpendicular to the direction of the propagation of the light wave. Normally, the vibration occurs in all the directions radial to the direction of propagation. When a light ray is "plane-polarized", the vibration is only in one direction perpendicular to the line of propagation. These two conditions are shown in Fig. 24A and B. The direction of propagation is perpendicular to the surface of the paper. Fig. 24A shows the directions of vibration for a normal light ray, and Fig. 24B shows the single direction of vibration for a plane-polarized light ray. The plane of polarization may be in any direction. A ray of light may also be elliptically polarized as shown in Fig. 24C.

Certain substances have the property that allows only light vibrations in one particular sense to be transmitted through them. Therefore, they have the property of plane-polarizing a light ray. A tourmaline crystal will polarize light when the incident ray bears a certain relation to the optical axis of the crystal. Polaroid is a commercially prepared substance that has this property of polarizing light.

The transmission of light in these substances is not limited to only the light that is vibrating in the sense of the polarization of the emitted ray, but also includes the components of the light rays vibrating in other senses that are parallel to the direction of polarization of the emitted ray. Let the light ray shown in Fig. 24A be incident on a piece of polaroid, and assume that the transmitted ray is polarized in the vertical direction as shown in Fig. 24B. Light vibrating in a direction other than vertical can be resolved into two components; one parallel to the direction of polarization, and the other perpendicular to the direction of polarization. This is illustrated in Fig. 25. The vertical component of the ray shown in Fig. 25 will be transmitted by the polaroid, but the horizontal component will not be transmitted. Thus, components of all other rays except the one perpendicular to the direction of the polarization will be transmitted. The magnitude of the transmitted components for each direction of vibration increases as the angle between the direction of vibration and the direction of the polarization decreases.

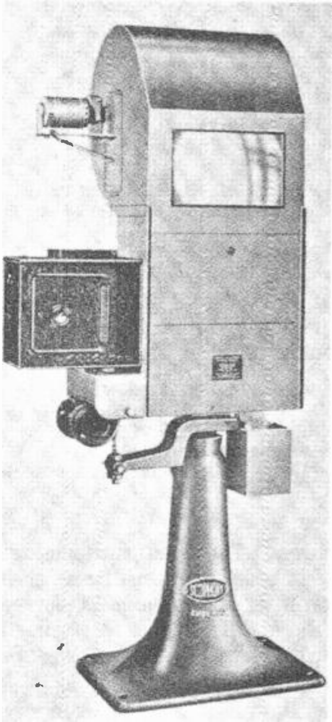


Fig.23 Scophony projector for theaters.

In some substances like quartz and calcite, the transmitted ray is separated into two rays vibrating in mutually perpendicular planes. This happens because the velocity of transmission is different for the mutually perpendicular vibration planes. By eliminating one of the rays, the output ray will be plane-polarized. This can be done in the case of calcite by splitting the crystal into two sections and then cementing them together again with Canadian balsam. The ray vibrating in one direction is totally reflected from the layer of cement to one side of the crystal, and the ray vibrating in the perpendicular plane is transmitted through the layer of cement without change in path. The crystal splitting must be properly orientated to the optic axis of the crystal. This type of polarizer is known as a Nicol prism after the man who originated the idea. The construction of a Nicol prism is illustrated in Fig. 26.

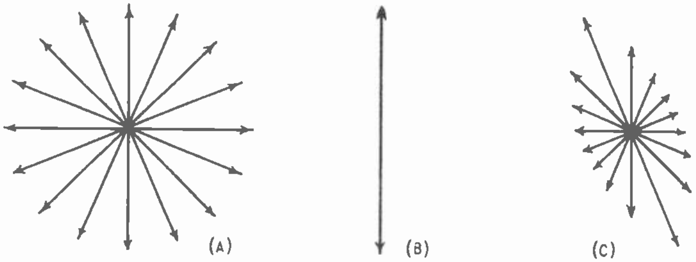


Fig.24 Vibration of light in: (A) normal ray, (B) plane-polarized ray, and (C) eliptically-polarized ray.

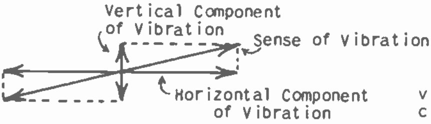


Fig.25 Resolution of a single vibration into two perpendicular components.

If two Nicol prisms are arranged so that the planes of polarization of the transmitted ray are perpendicular, no light will be transmitted by the combination. If the two are arranged so that the planes of polarization are parallel, the plane polarized ray from the first will pass through the second. If the angle between the planes of polarization is between 0 and 90 degrees, part of the light transmitted by the first will be transmitted by the second. The plane-polarized ray from the first prism can be resolved into

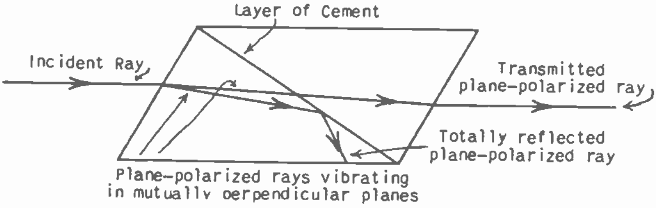
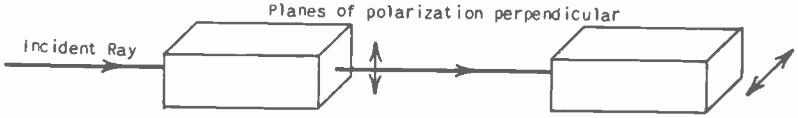
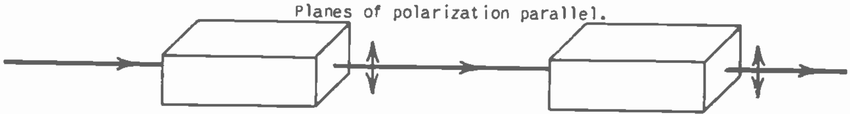


Fig.26 Construction of Nicol prism.

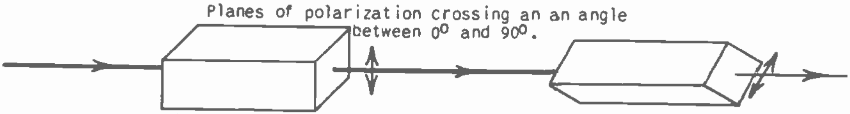
two components vibrating in perpendicular planes; one parallel to the plane of polarization of the second prism, and the other perpendicular to the plane of polarization of the second prism (See Fig. 25). The parallel component will be transmitted by the second prism. The amount of light getting through the second prism will depend on the orientation of the planes of polarization of the two prisms. This effect is shown in Fig. 27. The first prism is called the polarizer, and the second is called the analyzer.



(A) No light transmitted by the second prism.



(B) All the light transmitted by second prism.



(C) Part of the light transmitted by the second prism.

Fig.27 Transmission of light through two Nicol prisms.

Certain liquids, when under the influence of an electrostatic field, have the ability to polarize light which is passing through the liquid perpendicular to the field. This effect was first noted by Kerr. The ability of the liquid to polarize light increases as the intensity of the applied electric field is increased. Therefore, this effect can be utilized in the operation of a light valve. Such a light valve is known as a "Kerr cell". The simple Kerr cell has two metallic parallel plates at the top and bottom of a container which holds a liquid such as nitrobenzene. The walls of the container are transparent. The electrostatic field is produced by applying a DC potential between the two plates.

The operation of a light valve using a Kerr cell is best explained in terms of Fig. 28. The light valve consists of a Kerr cell placed between two Nicol prisms. The planes of polarization of the two prisms are perpendicular, and are inclined at a 45 degree angle with the direction of the electrostatic field between the plates in the Kerr cell. When there is no potential applied across the plates of the Kerr cell, it has no effect on the polarization of the light passing through it. Since the polarization planes of the two Nicol prisms are perpendicular, no light will pass through the second prism. When a potential is applied across the plates of the Kerr cell with the polarity shown in Fig. 28, light is transmitted through the combination of Nicol prisms and Kerr cell. The intensity of the transmitted light depends on the magnitude of the applied potential. In other words, the Kerr cell seems to rotate the plane of polarization of the light passing through it. A complete explanation of the action of the electric field on the plane-polarized light ray passing through the liquid, is beyond the scope of this lesson. However, as far as its application to television is concerned, we shall assume that the Kerr cell does rotate the plane of polarization of the light ray passing through it. Full modulation of the emerging light beam can be obtained by varying the rotation of the plane of polarization from 0 to 90 degrees.

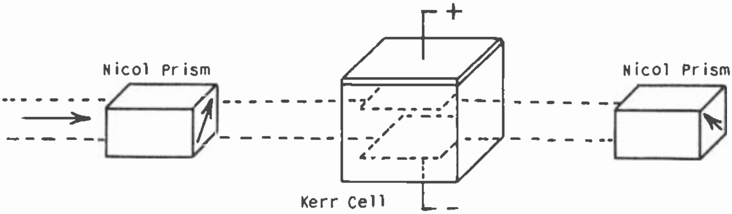


Fig.28 Light valve using a Kerr cell.

Fig. 29 is a graph showing the relation between the magnitude of the light flux passing through the second Nicol prism or analyzer, and the DC potential applied to the plates of the Kerr cell in Fig. 28. The rotation of the plane of polarization from 0 to 90 degrees requires a potential change from 0 to about 600 volts. The rotation from 90 to 180 degrees requires a potential increase from 600 to about 900 volts. It is evident that the voltage increase required to produce each additional rotation of 90 degrees becomes progressively less. For each 90 degree rotation, complete modulation of the light beam from black to white is obtained. However, the maximum output light flux is only fifty per cent of the input.

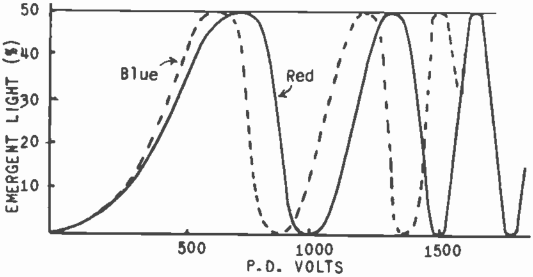


Fig.29 Characteristic curve of Kerr cell light valve.

The graph also shows that some dispersion takes place in the Kerr cell, and the magnitude of the dispersion becomes greater as the potential is increased.

It is evident from Fig. 29 that there are several operating ranges on the characteristic curve of the Kerr cell. Also, it is evident that the voltage change required to obtain complete modulation decreases with the higher applied potentials. Any of these operating ranges can be used by applying the proper DC bias voltage across the plates of the cell. Then modulation can be obtained by superimposing a picture signal of the necessary peak to peak amplitude on the DC bias, as for a conventional cathode ray tube. However, in actual use, the first operating range from 0 to about 600 volts must be used because of the dispersion occurring on the higher operating ranges.

The light valve shown in Fig. 28 in conjunction with a suitable light source, lens system, and mechanical scanner, can be used to project a large scale television picture.

As far as television is concerned, the Kerr cell is not as efficient as the supersonic light valve. With the Kerr cell, only one picture element can be projected at a time, instead of the 150 to 200 for the supersonic cell. Also, a much larger percentage of the transmitted light flux is lost in the Kerr cell than in the other. The capacity between the plates of a Kerr cell is of the order of .001 to .008 mfd., depending on the liquid and construction of the cell. Therefore, a very large amount of power is required to modulate a Kerr cell from black to white over the range of frequencies present in a modern high-definition television picture. At present, no modern commercial television receivers have been constructed using the Kerr cell in a light valve. It is not likely that such receivers will be built.

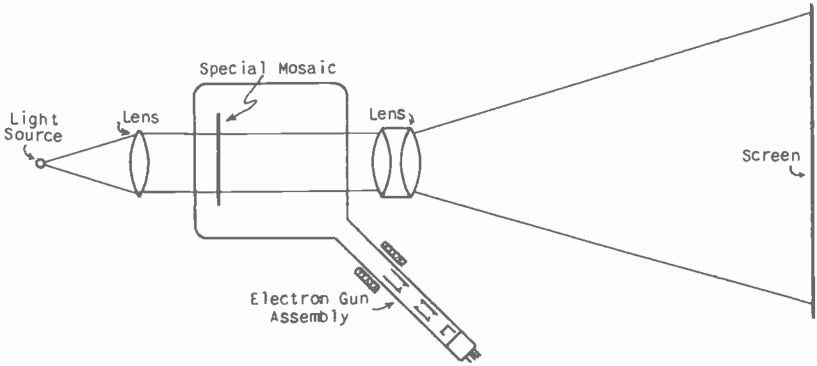


Fig.30 Ideal solution to problem of producing a large television picture.

11. THE IDEAL PROJECTION SYSTEM. The ideal projection system is one that makes use of the brilliance obtained from a high intensity source, such as an electric arc, and electronic scanning. Several systems have been developed with these criteria in mind, but up to the present, none of them have proved satisfactory. Fig. 30

shows the basic idea for most of them. The light from a high-intensity source is projected on a mosaic in a cathode ray tube. The mosaic is scanned by an electron beam. The mosaic is constructed so that it will transmit light through the section that is under the electron beam. All the other sections, except the one under the electron beam, are opaque. The amount of light transmitted by any elemental area of the mosaic is proportional to the intensity of the electron beam incident on that element. The light passing through the mosaic is projected on a screen.

Such a system, when and if it is perfected, will undoubtedly be the answer to the problem of obtaining a large scale television picture.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. What are the two methods that have been developed to produce a large television picture?

2. In what two ways can the brilliance of a cathode ray picture be increased?

3. What is diffraction?

4. What is a supersonic wave?

5. What is the difference between a spherical converging lens and a cylindrical converging lens?

6. Describe briefly the operation of a supersonic light valve.

7. What is plane-polarized light?

8. State briefly the principle of operation of a light valve using a Kerr cell.

9. Give two reasons why the Kerr cell is inferior to the supersonic light valve.

10. What are your ideas concerning methods of producing large scale television pictures?

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UNIT
NO.
7

FACSIMILE

LESSON
NO.
10

FACSIMILE

.....radio pictures and newspapers.

If you will stop and give serious thought to the many truly amazing things that we accept as commonplace, you cannot help but recognize the tremendous scope of present and future opportunities in the field of Radio and Television.

Many men, some young and others not so young, fail to recognize the extent of the opportunities that exist today. And the reason is just this: they do not keep abreast of the times. Their reasoning is based on old industries that have reached their peak or are on the decline.

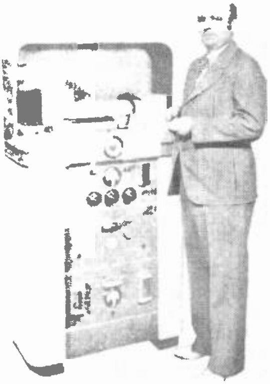
In this fascinating lesson, you will take up the study of Radio Facsimile----electrical impulses flashed through the air, that draw pictures and print newspapers at the receiving end.

Photographs of important events taking place in distant lands today, appear in our newspapers tomorrow. Just a few years ago, photos of importance were rushed to the newspapers by airplane. Today, Facsimile does the job in a flash. An attachment plugged into your radio set at night when you retire, reproduces a newspaper----headlines, print, and photos, while you sleep. What would your great grandfather think and say, were you to tell him about Facsimile?

Opportunities for the youth of the world, particularly America, are richer and greater than ever before. You realize this, and that is why you are a Midland student. So when you encounter anyone who may claim that opportunity is dead; do him a big favor. Put him wise to the great things that are taking place today.

Lesson Ten

FACSIMILE



"Although facsimile differs from television in many respects, still a great deal of the knowledge you have gained about television can readily be applied to facsimile.

"This method of transmitting information by wire or radio has been in use for a number of years. However, its popularity and number of applications is growing rapidly, and no engineer's knowledge would be complete without a very thorough understanding of its principles. This lesson has been designed to supply that need."

1. FACSIMILE AND TELEVISION. Television is the electrical transmission of several complete pictures per second. Facsimile is the electrical transmission of a single picture over a period of several minutes. Television is used for the transmission of scenes involving action. Facsimile is used for the transmission of photographs, letters, newspapers, etc.

In both systems the picture must be scanned or resolved into a large number of elemental areas. The light energy from each area is converted into an electrical impulse of proportional magnitude. Each of these electrical impulses is transmitted separately. At the receiver the electrical impulses are reconverted into elemental areas which radiate (television) or reflect (facsimile) light intensities proportional to the magnitude of the electrical impulses. These areas are combined in the proper order to reproduce the original picture.

The detail transmitted by the average facsimile system is much greater than modern television systems. Present day television systems transmit pictures with 441-line definition. Therefore a picture with an aspect ratio of 4:3 will have 259,308 picture elements. One of the present day facsimile systems transmits an 8×12 inch picture with 125 lines per inch definition. Therefore the total number of picture elements is 1,500,000. Thus there are approximately six times as many picture elements transmitted for a facsimile picture than there are for a television picture.

Since the time of transmission is not an important factor in facsimile, the range of frequencies produced in scanning a picture is just a fraction of the range generated in a modern television system. The formula for calculating the maximum frequency produced in scanning a picture is:

$$F = \tfrac{1}{2}A^2RN$$

where F is the maximum frequency generated, A is the number of lines, R is the aspect ratio, and N is the frame repitition frequency (number of pictures transmitted per second). From the formula we see that the maximum frequency generated is directly proportional to the frame frequency. For a modern television system, N is 30 cycles, and the maximum frequency generated is approximately 4 megacycles. The range of frequencies generated is from 0 to 4 megacycles.

The time of transmission for one picture in a facsimile system may range from ten to twenty minutes. A transmission time of ten minutes is equal to:

$$\frac{1}{10 \times 60} \text{ or } .00167 \text{ frames per second.}$$

Therefore the frame frequency is .00167 cycles per second. The maximum frequency generated in scanning a picture having the same detail as a modern television picture at the rate of .00167 frame per second is approximately 216 cycles. Therefore the range of frequencies generated is from 0 to 216 cycles. Since the frequency range is very small, a picture of much greater detail can be transmitted by facsimile than by television. The maximum frequency generated for a facsimile picture containing 1,500,000 elements at the rate of .00167 picture per second is 1250 cycles. The range of frequencies generated is from 0 to 1250 cycles.

The transmission of DC is required in both facsimile and television in order to transmit the average light level of the picture. The lowest AC frequency that must be transmitted by a television system is equal to the frame frequency for straight scanning and to the field frequency for interlaced scanning. Therefore the lowest AC generated in a modern television system is 60 cycles. The frequencies greater than 0 and up to 60 cycles are not transmitted. The minimum frequency that television amplifiers must pass without attenuation and phase shift is 60 cycles. The DC component can be inserted wherever desired by establishing a constant blanking level through the use of one of the methods described in Lesson 9 of Unit 5.

The lowest AC frequency that must be transmitted by a facsimile system is also equal to the frame frequency. For a ten minute transmission time per picture, the frame frequency is .00167 cycles per second. Since this frequency is practically DC, all the frequencies included between 0 and the maximum must be transmitted by a facsimile system. Therefore in a facsimile system, one problem is to obtain a linear frequency response for a range from 0 cycles to the maximum generated. There is a simple solution to this problem and it will be discussed in a later section of the lesson.

Since the frequency band required by facsimile is very narrow, it is a simple problem to find an electrical communications channel that will pass a facsimile signal without distortion. Expensive coaxial transmission lines are needed to transmit a video signal when wire transmission is desired. For radio transmission, the carrier frequency must be above 40 megacycles. An ordinary telephone line will transmit a facsimile signal without distortion. Likewise, an ordinary broadcast station can transmit facsimile, as the frequency requirements are well within the seven kilocycle sidebands that a broadcast station is allowed to radiate. There will be described in a later section of this lesson a method of using the facsimile signal to key a telegraph transmitter and transmit a picture.

The scanning systems used in facsimile are much simpler than those used in television. Since the scanning rate is so low, mechanical scanners will give very satisfactory service.

The amplitude of the picture signal generated in scanning a facsimile picture is many times greater than that for a television picture. The amount of electrical energy obtained per picture element is much greater because of the slower scanning rate. Since the subject matter consists of printed material and photographs, the optical systems can be designed for maximum efficiency.

The scanners at the transmitter and receiver must be synchronized in facsimile as in television. Since the scanning rate is much lower in facsimile than in television, synchronization is not as critical. A television picture with an aspect ratio of 4:3 and 441-line definition contains 259,308 picture elements. The number of picture elements transmitted per second will be 30 times 259,308, or 7,779,240. The time of transmission of one picture element is $\frac{1}{7,779,240}$ second, or approximately .13 microseconds. Therefore, the synchronization should be maintained accurately within .13 microseconds in order to prevent a displacement of picture elements and the resulting loss of detail.

For a facsimile system transmitting a picture containing 1,500,000 picture elements at the rate of .00167 frame per second, there will be .00167 times 1,500,000, or approximately 2500 picture elements transmitted per second. The time of transmission of one element is $\frac{1}{2500}$ second, or 400 microseconds. Thus the required accuracy of synchronization in a facsimile system is much less than that of a television system.

2. TYPES OF FACSIMILE SIGNALS.

There are three types of signals generated by facsimile systems. These are: modulated audio frequency tone, keyed audio frequency tone, and direct current signals. These signals may be transmitted over a line to the receiver or used to modulate or key the carrier of a radio transmitter.

The picture signal generated by the scanning of the facsimile picture covers a continuous range of frequencies from zero to the maximum determined by the number of picture elements. To build up such a signal having a magnitude suitable for transmission over a telephone line or to modulate a transmitter, DC-coupled amplifiers would have to be used. A stable multi-stage DC amplifier is very difficult to build and maintain in adjustment. Also, telephone lines usually will not pass frequencies below 30 cycles. In order to modulate a transmitter with extremely low frequencies, the modulator would have to be DC-coupled to the modulated stage, as in television. To eliminate these difficulties, the output of the scanner is used to modulate an audio carrier frequency. Then conventional AC-coupled amplifiers can be used to build up the magnitude of the carrier and sidebands.

The output of the facsimile system consists of an amplitude modulated audio frequency tone. Either positive or negative modulation can be used. This signal can be transmitted to the receiver over a telephone line or used to modulate the carrier of a broadcast transmitter.

The amplitude of the modulation of the audio carrier is proportional to the shade between black and white of the transmitted picture element. With positive modulation, 100% modulation will occur for white, and with negative modulation, 100% modulation will occur for black. If the received picture is to be a true replica of the transmitted picture, the carrier amplitude must remain constant during the transmission period of the picture. Constant carrier amplitude can be maintained over telephone lines and in the primary coverage area of broadcast stations.

Long distance radio communication is accompanied by fading. Variations in the RF carrier will produce corresponding variations in the audio carrier, and the shading of the reproduced picture will not correspond to that of the transmitted picture. The effect of fading can be reduced by using three receivers with separate antenna systems. The antennas are spaced about 1000 feet apart and the outputs of the three receivers are combined. With this arrangement, it is improbable that the amplitude of the received signal will be zero in more than one of the three receivers at once. With the use of automatic volume control and diversified reception, it is possible to maintain a signal of far more constant amplitude than with a single receiver and antenna. However, even with "diversified reception" it is impossible to maintain a facsimile signal with a carrier amplitude sufficiently constant to reproduce a satisfactory picture.

From the foregoing, it is evident that amplitude modulation is not satisfactory for transmitting a picture over long distances by radio. The system that has been developed for long distance radio transmission of facsimile is an electrical equivalent of the newspaper or magazine half-tone. A newspaper half-tone consists of rows of dots having the same distance between centers (See Fig. 1). Each alternate row is staggered so that its dots lie beneath the center of the space between the dots of the preceding line. The shade of all the dots is identical. The light and dark shades of the picture are obtained by varying the size of the dots, the larger dots occurring in the darker sections of the picture.

The electrical equivalent consists of pulses having a constant frequency and amplitude but having a variable duration. This is known as the CFVD (constant frequency, variable dot) system. These pulses may be used to key an RF carrier, or used to modulate an audio tone for telephone line transmission. Fading will have no effect on the shades of the reproduced picture as the shade between

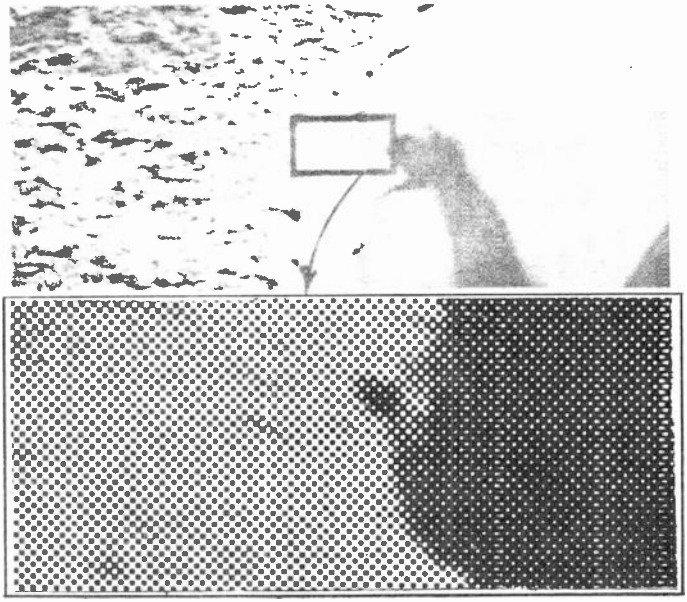


Fig.1 Illustrating the dot formation of a newspaper half-tone.

black and white depends on the duration of the dot and not its amplitude. To prevent the received signal from fading out completely, diversified reception is used. The method of producing the CFVD type of signal will be explained in a later section of the lesson.

There are many systems of facsimile in use today. However the general methods are the same in all of them. Therefore instead of discussing each system in detail, all of the important features of the various systems will be covered in a general discussion of facsimile circuits and apparatus.

3. FACSIMILE TRANSMISSION. This section of the lesson will include a description of the apparatus and circuits involved in the scanning of a picture or printed material, generation of the picture signal, carrier tone modulation, amplifiers, and the specialized circuits required to produce the CFVD type of signal.

All the scanning systems used in facsimile are mechanical. There are two general types; those designed to scan continuous copy, and those designed to scan a single picture or printed page. The latter are most common, as their construction and operation is much simpler. There is another type of scanner that will scan continuous copy in the form of a narrow tape. This method can be used only for the transmission of messages. The prepared tape is similar to printed tape from a stock ticker or telegraph tape printer. Fig. 2 shows schematically the simplest form of facsimile scanner. The picture or copy to be transmitted is wrapped around the drum. An optical system projects perpendicularly, a small spot of light of rectangular or square cross section on the picture or copy. The drum is rotated and the scanning light spot is moved parallel to the axis of the drum the width of a scanning line for each revolution of the drum. Thus as the drum rotates, the entire copy is scanned by the light spot. The light reflected from the copy is picked up by a photocell which moves along with the light source. The amount of light picked up by the photocell and the corresponding photocell current will vary with the light and dark shades of the picture.

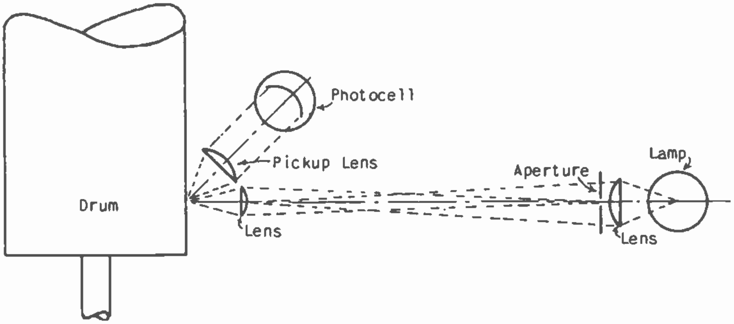


Fig.2 Simple drum scanner using reflected light.

The photocell is so placed that it will pick up only diffused light and not the direct reflection from the scanned spot. The direct reflection from a smooth and glossy surfaced copy will be fairly constant and not vary linearly with the shading of the picture. The diffused light from the scanned spot will depend on the shades of light or dark.

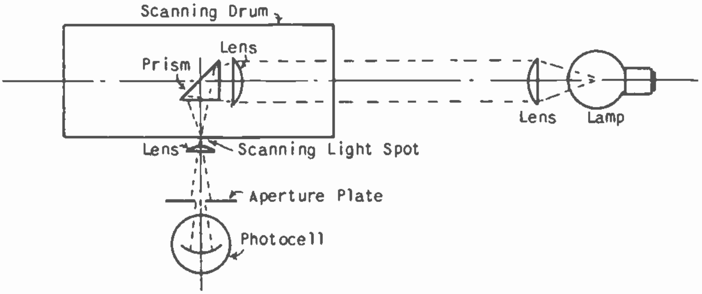


Fig.3 Simple drum scanner using transmitted light.

The size of the copy used on this simple drum scanner is determined by the usable area of the drum. Therefore, copy must be printed on a sheet of paper that will fit the drum.

Fig. 3 shows another form of drum scanner. The drum is transparent, and a light is projected through the copy wrapped around the drum. The copy used must be in the form of a photographic negative. The light source is at one end of the drum and the light is reflected through the drum by means of a 90° prism. The light from the source is focused into a parallel beam by a lens. This parallel beam is focused to a small spot on the copy by means of another lens. The light passing through the copy is focused on the aperture plate. The aperture is the size of a picture element. The light passing through the aperture is picked up by a photocell.

This system is less flexible than the preceding, because the copy must be in the form of a photographic negative. The amplitude of the picture signal generated is greater, as considerably more light is transmitted through a negative than that reflected from a printed surface.

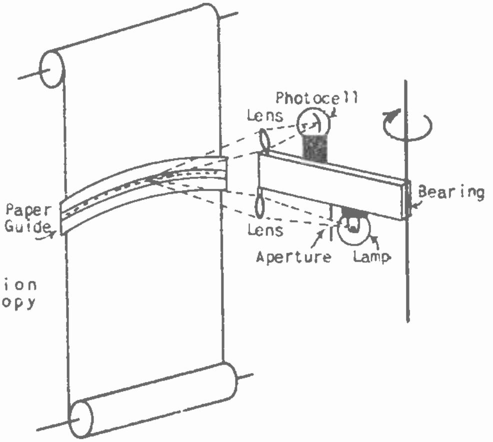


Fig. 4 Principle of operation of the Finch continuous copy scanner.

Fig. 4 shows the principle of operation of a continuous copy scanner. The copy to be scanned is arranged on a long narrow sheet of paper. The light source and photocell are mounted on an arm. One end of the arm is mounted on a bearing. Scanning is accomplished by sweeping the arm back and forth through a short arc. The spot of light sweeps across the paper from side to side. During the right to left swing of the light spot, the output of the photocell is used. During the left to right swing, the paper is displaced the width of a scanning line, and the output of the photocell is disconnected from the line. In order to reduce distortion, there is a guide to hold the surface of the paper along the line being scanned in the form of an arc which has the same center as the bearing supporting the arm carrying the photocell and light source.

The tape type of scanner, designed for continuous copy, is quite extensively used. Fig. 5 shows schematically the operation of a tape scanner. Fig. 6 shows how the rotating glass prism causes the light beam to scan a line across the tape and return instantaneously to the other side. You recall from your study of Lesson 3 in Unit 5, that a light ray is bent or refracted when it passes from one substance into another. It is bent toward the perpendicular through the boundary when entering a denser substance, and is bent away from the perpendicular when entering a less dense substance. Since the opposite surfaces of the prism in Fig. 6 are parallel, the paths of the ray entering and leaving the prism are parallel. The displacement of the beam is proportional to the angle of incidence. The scanning prism and movement of the tape is synchronized so that the tape moves the width of a line during the time the light beam scans a line. A photocell collects the light reflected from the tape.

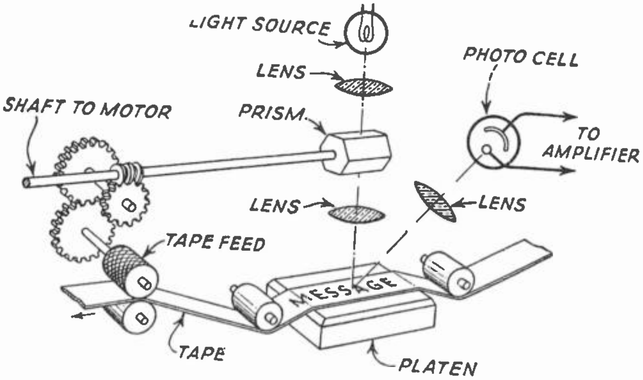


Fig.5 Principle of operation of a tape scanner.

4. RESOLUTION AND APERTURE DISTORTION. In most facsimile scanners, the length of the scanning light spot is slightly narrower than the distance between the centers of the scanned lines. This is necessary in order to obtain sufficient resolution at right angles to the scanning lines. The width of the scanning spot is considerably less than the length. This increases resolution along the line and helps correct for aperture distortion.

Since the resolution at right angles to the scanning lines, and aperture distortion are two important factors controlling the amount of detail transmitted by either facsimile or television, it will be advisable to review the nature of both briefly. The number of picture elements in a picture is equal to A^2R, where A is the number of lines, and R is the aspect ratio. The smallest detail that can be resolved is of the same size as a picture element. With these two points in mind let us examine Figs. 7 and 8. Fig. 7 is a pattern consisting of horizontal black and white bars that have the same width as a scanning line. Then a picture element will be a square whose side has the same length as the width of a scanning line. In Fig. 7A the pattern is arranged so that the scanning lines fall directly on top of the horizontal bars. Then the scanning aperture which has the same size as a picture element will scan alternately the black and white bars. The reproduced picture will consist of horizontal black and white bars. In other words the resolution is the maximum that can be obtained at right angles to the scanning lines.

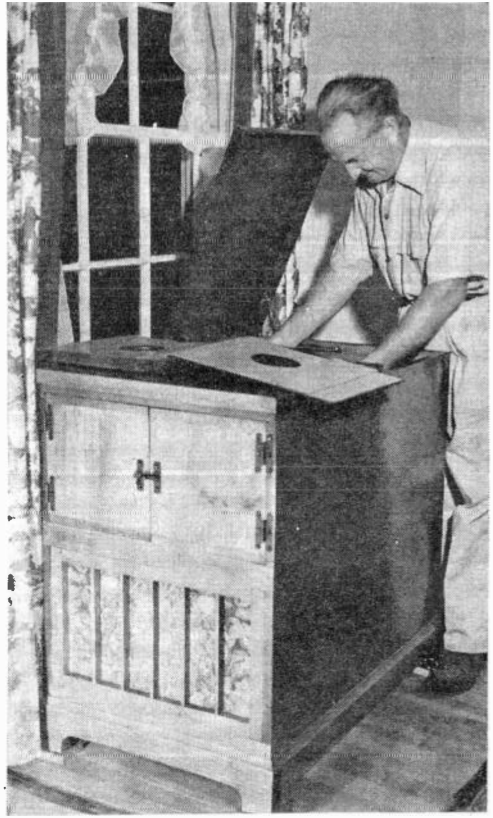
In Fig. 7B the pattern is shifted so that the scanning lines fall half on a black bar and half on a white bar. The light collected by the photocell will be the same for every line, as the scanning aperture is always covering half of a black bar and half of a white bar. The reproduced picture will not contain any bars and will have a shade half way between black and white. The resolution at right angles to the scanning lines is zero. Then the

---

Bring Out High Notes with a Tweeter

IS THE loudspeaker bottlenecking your radio-phonograph? It is if you have a good audio system feeding into a single speaker. For no matter how much sound quality reaches the plate of the output tube, you'll only hear what the speaker can reproduce.

Single speakers are found in the overwhelming bulk of all radios and phonographs. For the most part they give adequate sound reproduction, but in many cases they do not deliver the full potential



there will be times when the scanning aperture will pass over half of a black and half of a white bar when the bars have the width of a scanning line. Thus, it is a matter of chance that the resolution at right angles to the scanning lines will be equal to the number of lines.

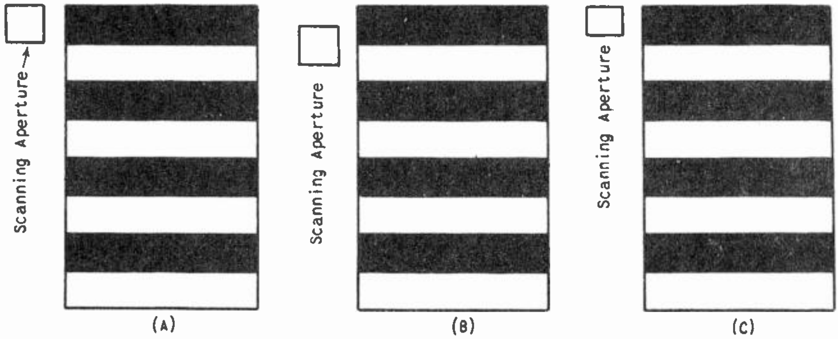


Fig.7 Conditions affecting vertical resolution.

Fig. 8 illustrates what is meant by aperture distortion. The pattern consists of vertical black and white bars that have the same width as a picture element. With a scanning aperture that is the same size as a picture element, this is the smallest detail that can be resolved. The waveform labeled A is the ideal voltage waveform that should result when one line of the pattern is scanned. Since the scanning aperture has a definite size and requires a certain amount of time to cross the boundary between the black and

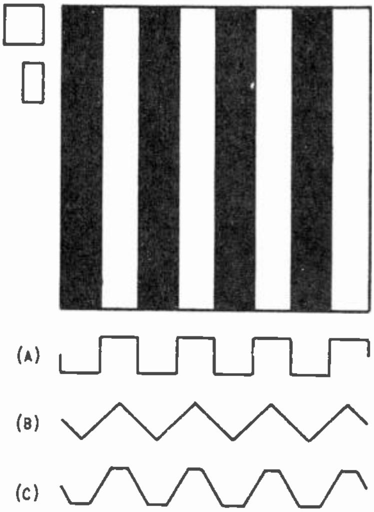


Fig.8 Cause of aperture distortion.

white bars, the actual voltage waveform that is generated will have the form shown in Fig. 8B. It is only when the scanning aperture is directly over a bar that the light collected by the photocell corresponds to the light reflected by the bar. At all other times the scanning aperture is covering part of both a black and a white bar. The deviation of the actual waveform from the ideal waveform is termed aperture distortion.

If the scanning aperture is made narrower than the width of a picture element, the aperture distortion is reduced as shown by the voltage waveform labeled C in Fig. 8. Of course, narrowing the scanning aperture improves the resolution along a scanning line, as it has the same effect as increasing the number of picture elements per line.

If the highest frequency passed by the amplifiers of a television or facsimile system is that determined by the formula:

$$F = \tfrac{1}{2}A^2RN$$

Then the voltage waveform appearing at the receiver for the condition illustrated in Fig. 8 when the scanning aperture is the same size as a picture element will be a sine wave and not the triangular wave labeled B in Fig. 8. To transmit the triangular wave, all amplifiers would have to pass at least the second and third harmonics of the maximum frequency. This is impractical in a television system because the transmitter would have to radiate sidebands three times as wide as they are at present. It is also unnecessary, since the observer cannot tell whether the maximum detail obtainable is produced by a sine wave or a triangular wave. In facsimile, where the maximum frequencies involved do not present amplifier or transmission problems, the third harmonic of the calculated maximum is usually transmitted. Without the transmission of the third harmonic of the maximum frequency, decreasing the width of the scanning aperture would not partially correct for aperture distortion as illustrated by the waveform labeled C in Fig. 8. If the maximum frequency transmitted was limited to the calculated value, the reproduced waveform would be a sine wave regardless of the generated waveform.

The past few paragraphs are a digression from the subject of facsimile, but they review ideas that are pertinent to an understanding of resolution.

5. **MODULATION.** Since the scanning methods used in facsimile systems have been covered, our next problem is to amplify and change the photocell output voltage so that it will be suitable for transmission to the receiver. It was stated previously that the output of the photocell contained a continuous range of frequencies from 0 to the calculated maximum or to the third harmonic of the calculated maximum. It was also stated that it was impractical to amplify and transmit such a range of frequencies directly. Therefore, the output of the photocell was used to amplitude modulate an audio frequency carrier. Then the modulated AF carrier can be amplified by conventional AC-coupled amplifiers and transmitted over telephone lines.

The selection of a carrier frequency is determined by two factors. These are the number of picture elements transmitted per second, and whether or not the second and third harmonics of the calculated maximum frequency are to be transmitted. It is a generally accepted standard that the carrier frequency should be at least three times the highest modulating frequency. If a lower carrier frequency were used, it would be difficult at the reproducer to filter the carrier frequency from the modulating frequencies. Since the transmission rates of the various facsimile systems are different, the audio carrier frequencies used by them are also different.

In the beginning of this lesson a facsimile system transmitting 1,500,000 picture elements in a ten minute period was used to compare the frequency requirements of television and facsimile. The maximum calculated frequency was 1,250 cycles. If the third harmonic of this frequency is to be transmitted, the carrier audio frequency should be 11,250 cycles. If the maximum calculated frequency only is transmitted, the carrier frequency should be 3,750 cycles. The carrier audio frequencies used in commercial facsimile systems range from around 2,000 cycles to 20,000 cycles.

Several methods have been used to modulate the carrier frequency with the output of the photocell. The photocells available several years ago, when used in facsimile scanners, developed an output of approximately .05 volt across a ten megohm load resistance. It is very difficult to obtain a high percentage of modulation with conventional systems when the modulating voltage is so low.

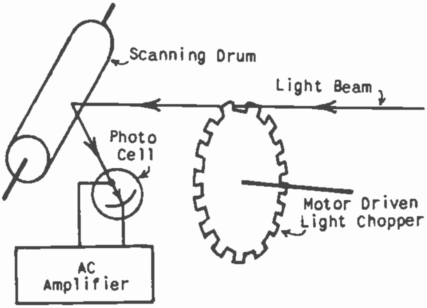


Fig.9 Introduction of the carrier by interrupting the light beam.

In early facsimile systems and in a few modern systems, the carrier frequency is developed in the output of the photocell by interrupting the scanning light beam at the carrier frequency rate. This can be done by directing the light beam through slots in the edge of a rotating disc (See Fig. 9). The carrier frequency is equal to the product of the number of slots and the speed of the disc in revolutions per second. The output of the photocell consists of an amplitude modulated audio frequency carrier, and can be amplified by conventional AC-coupled amplifiers. The modulation of the interrupted light beam is caused by the light and dark areas of the picture being transmitted. Since the light beam is shut off half the time, the amplitude of the generated signal is smaller than it would be if the light beam were not interrupted. Also the carrier amplitude is maximum for white. It is best to have the maximum amplitude for black as the maximum white level will be somewhat dependent on the extraneous light collected by the photocell.

These light chopper discs were satisfactory when the scanning rate was low and carrier frequencies of less than 500 cycles could be used. Light choppers suitable for high speed facsimile are bulky and are difficult to construct with the necessary physical precision. In modern systems using this method of developing a modulated carrier, light valves are used to interrupt the scanning beam. A light valve consists of a reed vibrating across a narrow slit type aperture. The light beam is directed through the slit and is interrupted as the reed vibrates. The light beam can be interrupted once or twice per complete vibration of the reed depending upon the location of the slit in reference to the path of the reed. If the slit is in the center of the reed swing, the beam will be interrupted twice per vibration; if it is at the end of the reed swing, the beam will be interrupted once per vibration. The reed is driven by an oscillating magnetic field energized by an audio frequency. The natural vibration frequency of the reed must be the same as the driving frequency. The carrier frequency is equal to the reed frequency or equal to twice the reed frequency, depending upon the number of light beam interruptions per reed cycle. Fig. 10 shows the construction of two RCA light valves. A is a valve designed for frequencies below 1500 cycles, and B is for frequencies above 1500 cycles.

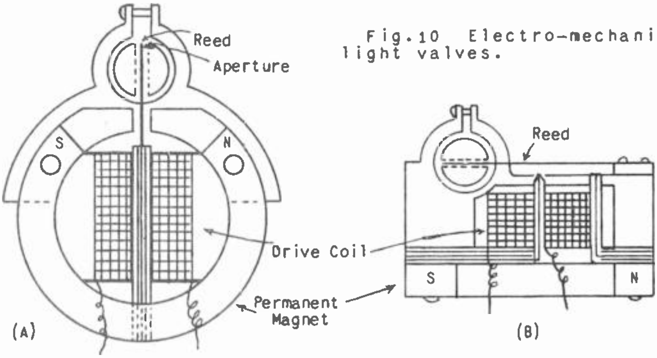


Fig.10 Electro-mechanical light valves.

A much larger output can be obtained with modern photocells, and more conventional systems of modulation can be used. A modern gas photocell, when used in a facsimile scanner, will develop one volt across a ten-megohm load resistance. Gas photocells can be used, as the facsimile frequencies are sufficiently low that the lag due to ionization does not cause trouble.

In order to obtain a high percentage of modulation, it is convenient to use a balanced modulator where the modulating frequency is used to control the amplitude of the carrier fed into the amplifier or line. This is superior to brute force modulators where the amplitude of the modulating frequencies must be of the same magnitude as the carrier in order to produce 100% modulation. Fig. 11 is the circuit of a balanced modulator. The modulated stage consists of two pentodes with their grid circuits in parallel, and their screen and plate circuits in push-pull. One stage of DC amplification is placed between the photocell and the modulated stage. The carrier frequency, generated by a separate source, is applied to the screens. When the *photocell is illuminated*, the output biases the grid of the first tube positive and the grids of the modulated stage negative. Since a negative modulated signal is preferable, the fixed bias on the modulated stage is adjusted for plate current cutoff when the photocell is receiving illumination from the white section of the copy. At this time there will be no carrier frequency appearing in the plate circuit of the modulated stage (tubes biased to cutoff). When the light collected by the photocell comes from the darker sections of the copy, the negative bias applied to the grids of the modulated stage will be reduced proportionately, and plate current will flow. Since the audio frequency carrier is applied to the screens, the output voltage will be the carrier frequency, amplitude modulated by the photocell output; that is, the output will consist of the carrier and both of its sidebands. The modulation will be negative, as the carrier amplitude is zero for white. Also, since white is represented by a definite carrier level (zero voltage), the DC component in the facsimile signal will be present. The modulating frequencies will not appear in the output, as the grids of the modulated stage are in parallel and the plates are in push-pull.

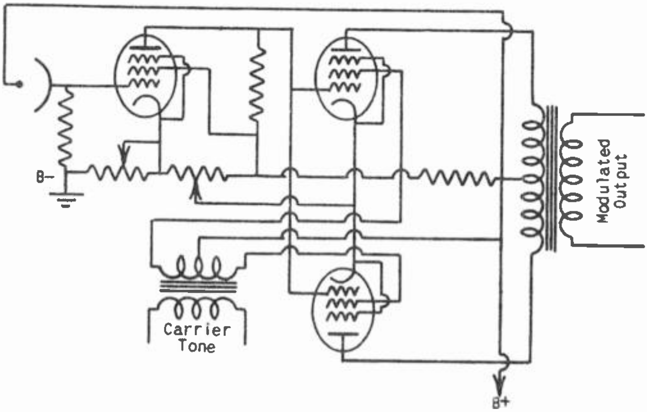


Fig. 11 Balanced Modulator.

If a single rather than a balanced modulator is used, the operation will be the same, but the modulating frequencies will not be balanced out in the plate circuit of the modulated stage, and a filter will have to be added to the output of the modulator to remove the modulating frequencies. If the carrier frequency is made relatively high, such as 20,000 cycles, and the highest modulating frequency is two or three thousand cycles, then the carrier and its sidebands will cover a rather narrow band of frequencies. If the output transformer has very poor low frequency response, the modulating frequencies will be effectively removed.

The main types of carrier tone modulators used in facsimile systems have been described in the past few paragraphs. In the next few paragraphs we shall study some of the specialized circuits used to modify the character of the facsimile signal.

6. **THRESHOLD-LIMITER AMPLIFIER.** When the scanned copy consists of printing or black and white drawings, a "threshold-limiter amplifier" is used to remove background noise and to correct for aperture distortion. For black and white copy, the audio frequency carrier has zero amplitude for white and maximum for black (negative modulation). Also, in order to reproduce the sharp boundaries between black and white, required for printing and drawings, the sloping leading and trailing edges of the wave, caused by aperture distortion, must be squared up.

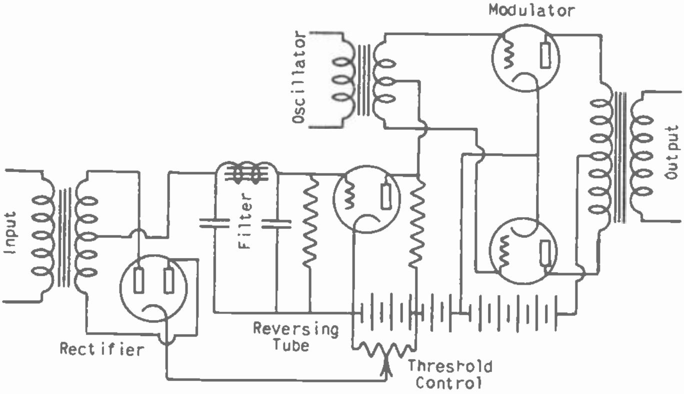
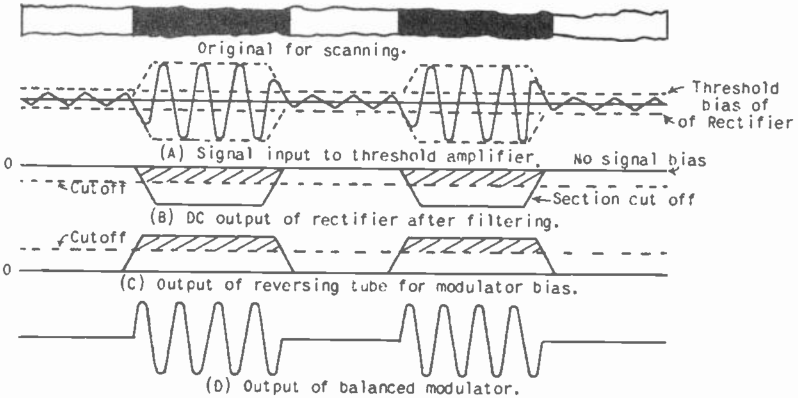


Fig.12 Threshold-limiter Amplifier.

Fig. 12 is a schematic of a threshold-limiter amplifier. Fig. 13 shows the modification of the facsimile signal as it passes through the threshold-limiter amplifier. Fig. 13A shows the output signal for black and white copy from a modulator unit such as shown in Fig. 11, which is connected to the input of the threshold-limiter amplifier. The amplitude is maximum for black, and zero for white. There is considerable noise present from shot effect and thermal agitation occurring in the photocell and its coupling resistor. The interference caused by the noise is greatest at zero carrier or white. The pulses representing black have the sloping edges caused by aperture distortion.

This signal is fed into a full wave rectifier in which the plates are biased a few volts negative with respect to the cathode. Therefore, the rectifier will not conduct unless the voltage applied to the plates exceeds this negative bias. By applying a negative bias to the plates of the rectifier that is greater than the peak noise present in the signal, the noise will be completely eliminated. The bias control on the rectifier is called the "threshold control" as it determines the conduction point of the rectifier.



(A) Signal input to threshold amplifier.

(B) DC output of rectifier after filtering.

(C) Output of reversing tube for modulator bias.

(D) Output of balanced modulator.

Fig.13 Waveforms produced as signal is modified by threshold-limiter amplifier.

The carrier frequency is removed from the output of the rectifier by means of a low-pass filter. The waveform developed across the load resistor of the rectifier has the waveform shown in Fig. 13B. These pulses apply a negative voltage to the following tube. The edges of these voltage pulses still have the slope caused by aperture distortion. The top of the pulses are considerably shorter than the bottom. If large sections were sliced off the top and bottom, the resultant pulses would be rectangular and free of the slope caused by aperture distortion.

The tube following the rectifier and filter is called a reversing tube. It is operated at zero bias and the plate current is maximum for zero input to the rectifier. The pulses developed across the rectifier load resistor apply a negative bias to the grid of the reversing tube. The amplitude of these pulses is great enough to drive the grid of the reversing tube to considerably beyond cutoff. Thus, only the top section of these pulses will cause a plate current flow in the reversing tube and the bottom section is effectively sliced off. The difference in length between the top and bottom of the pulses existing in the plate circuit of the reversing tube is considerably less than that for the grid pulses. The phase of the waveform in the plate circuit of the reversing tube is opposite to the grid voltage waveform. This waveform is shown in Fig. 13C.

The output of the reversing tube is applied in phase to the grids of the balanced modulator. The purpose of the modulator is to reinsert the carrier which was removed by the rectifier and low-pass filter. A new carrier is applied to the grids of the modulators in push-pull. The plates are also connected in push-pull. The negative bias applied to grids of the push-pull stage is the sum of the fixed battery bias and the drop across the plate resistor of the reversing tube. This bias will be maximum when the input to the rectifier is zero. The fixed bias is adjusted so that normal plate current will flow in the push-pull stage when the plate current of the reversing tube is cut off. When the rectifier input is zero, the bias on the push-pull stage is considerably greater than cutoff, and so the output of the push-pull stage is also zero. When a signal is applied to the rectifier, only the top of the pulses developed in the plate circuit of the reversing tube have sufficient magnitude to overcome the cutoff bias of the modulators. Thus the bottom of the pulses appearing in the plate circuit of the reversing tube are sliced off. Since the phase difference between the plate and grid circuit of the reversing tube is 180°, both the top and bottom of the output pulses from the rectifier have been sliced off. Thus, the push-pull stage does two things; reinserts the carrier and helps to eliminate the slope caused by aperture distortion.

Fig. 13D shows the form of the output pulses from the push-pull stage. The effect of aperture distortion is gone. These pulses are slightly shorter than the original pulses, but that will not interfere with reproduction of the copy.

A tape facsimile system deals with just black and white as only printed copy is transmitted via tape facsimile. The circuits in a tape facsimile transmitter must produce sharp impulses free of noise and aperture distortion. The methods used in obtaining the desired type of output signals are similar to the threshold-limiter amplifier previously described. For that reason the actual circuits used will be omitted.

7. CFVD CONVERSION. When half-tone pictures are sent, the type of signal transmitted depends on the communications channel used. The amplitude modulated carrier is satisfactory for telephone lines and the primary coverage area of broadcast stations. When long distance radio communication is used, the facsimile signal must be converted into an electrical equivalent of the newspaper half-tone. This is called the CFVD system.

Before going into the circuit used to convert an amplitude modulated carrier into a CFVD signal, it will be well to review again the nature of a newspaper half-tone. A newspaper half-tone consists of rows of dots with a fixed spacing between centers. The dots of each alternate row are displaced so that their centers lie half way between the centers of the dots in the preceding row. All the dots are of the same shade of black. Shading is obtained by varying the size of the dots. The CFVD signal consists of dots of constant amplitude and repetition frequency, but with variable duration. The duration of the dots is proportional to the shade of black in the original.

Fig. 14 shows a circuit that will convert the amplitude modulated carrier into a CFVD signal. Fig. 15 shows the voltage waveforms occurring in various parts of the circuit for two consecutive lines. Fig. 15 also shows the original pattern scanned and the reproduced pattern.

The signal applied to the plates of the rectifier T1, is the amplitude modulated carrier, shown in Fig. 15B. This is the same for both lines. After rectification and filtering, the voltage developed across the rectifier load resistor has the form shown in Fig. 15C. This is also the same for both lines. Part of this signal is applied to the grid of T2. A triangular waveform, shown in Fig. 15D, is also applied to the grid of T2. This triangular waveform has the same frequency as the dot frequency, and it is approximately equal to the maximum calculated frequency generated in scanning the pattern. The fixed bias on the grid of T2 is made greater than cutoff by the peak amplitude of the triangular waveform. Therefore, the positive peaks of the triangular waveform will just swing the grid of T2 up to cutoff. The voltage applied to T2 by the rectifier will have its maximum positive value when the scanned section of the copy is black. The output of the rectifier is adjusted so that the maximum amplitude occurring on black will swing the grid of T2 more positive than the cutoff point by a voltage equal to the peak amplitude of the triangular wave. For this condition, the tube will be conducting over the entire triangular cycle and the negative peak of the triangular voltage will just swing the grid of T2 to cutoff.

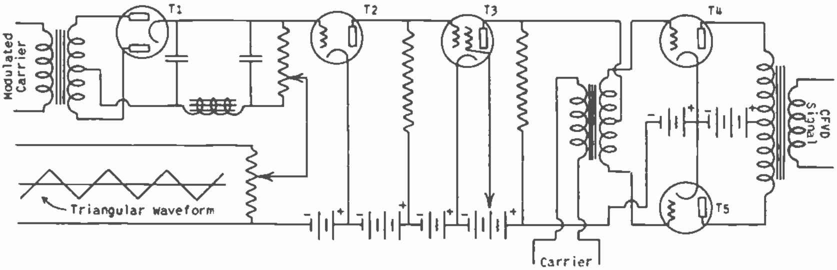
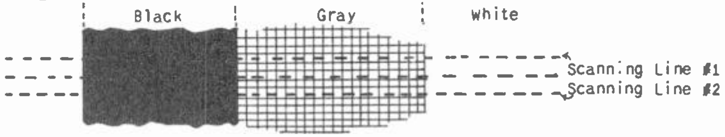
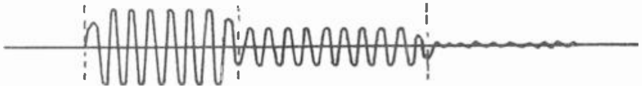


Fig. 14 Circuit to convert amplitude modulated carrier into a CFVD signal.

Fig. 15D shows how the output of the rectifier and the triangular waveform are combined in the mixer tube T2. Note that the output of the rectifier is identical for the two scanning lines, but the phase of the triangular wave differs by 180° for the two lines. The method of shifting the phase of the triangular wave by 180° for each alternate line will be discussed a little later. Fig. 15E shows the waveforms and their polarity for the two consecutive lines developed across the plate load of T2. For the black section of the original, the entire triangular wave appears in the plate circuit of T2; for gray, only the peaks of the triangular wave appear; and for white, the output is zero. The number of cycles of the triangular wave is independent of the shading of the original. The width of the peak of the triangular wave appearing in the plate circuit of T_2 is proportional to the blackness of the original pattern. For black, the width is maximum, and the peaks touch. For white, the width is zero. The signal appearing in the plate circuit of T_2 has a large amplitude.

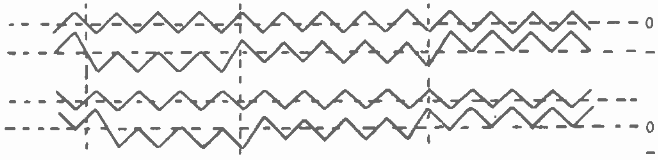


(A) Original for half-tone scanning.

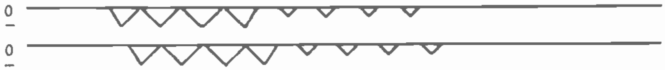


(B) Amplitude modulated half-tone signal.

(C) Wave "B", after rectification.

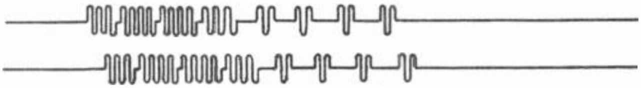


(D) Wave "C" added to triangular wave.

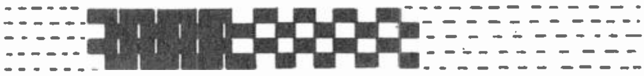


(E) Plate voltage of mixer for alternate lines.

(F) Plate Voltage of T3



(G) Output signal of variable dot amplifier.



(H) Appearance of recording--4 scanning lines.

Fig.15 Waveforms produced as amplitude modulated signal is converted into CFVD signal.

The voltage developed across the load resistor T_2 is zero for white, and swings negative for gray and black. The grid of T_3 is biased almost to cutoff and the voltages developed by T_2 for gray and black will swing the grid of T_3 considerably beyond cutoff, and only a small section of the peaks will appear in the plate circuit of T_3. Fig. 15F shows the waveform and polarity for two consecutive lines developed across the plate load of T_3. It consists of rectangular pulses of constant amplitude but variable duration. However, for white, the amplitude is zero. The shade between black and white is determined by the duration of the pulse.

The voltage developed across the load resistor for T_3 is zero for white, and swings positive for gray and black. T_4 and T_5 constitute a balanced modulator for reinserting the audio carrier frequency. For zero input to the circuit or white, the bias on the grids of T_4 and T_5 is at cutoff. This bias is the sum of the fixed bias and the negative drop across the plate load of T_3. For this condition there will be no output from T_4 and T_5. When the plate of T_3 swings positive for gray and black, T_4 and T_5 draw plate current, and a voltage is produced in the output. Fig. 15G shows the output from T_4 and T_5 for two consecutive lines of the picture.

When Fig. 15G is compared to Fig. 15B, we see the shades of black are no longer represented by the amplitude of the modulation of the carrier. The shading of the picture is now represented by dots of fixed amplitude and frequency but having variable duration.

Fig. 15H shows how four lines of the reproduced picture appears. As in a newspaper, half-tone shading is obtained by varying the size of the component black dots.

In order to stagger the dots of each alternate line, there must be a definite relation between the frequency of the triangular wave and the number of lines scanned per second. In order to shift the phase by 180° for each alternate line, the phase at the beginning and end of a line must differ by 180°. Then the beginning of a line will differ by 180° from the beginning of the preceding line as far as the triangular wave is concerned. In other words, there must be an odd number of half-cycles of the triangular frequency per line. As stated before, the triangular frequency is approximately equal to the maximum frequency generated when the copy is scanned.

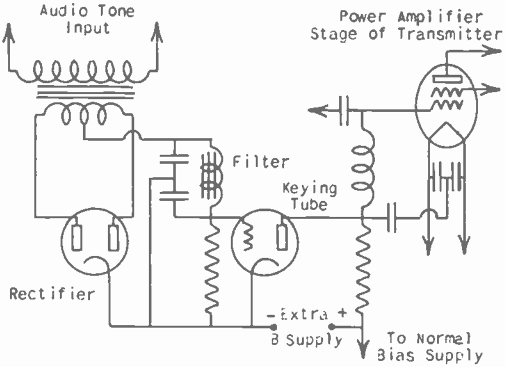


Fig.16 Keyer for telegraph transmitter.

The CFVD signal can be sent to a distant point by telephone lines, by a broadcast transmitter, or by a telegraph transmitter. Fig. 16 shows a circuit for keying a telegraph transmitter with the output of a CFVD system. The signals are first rectified and filtered, and the resulting DC pulses are used to vary the bias on one of the transmitter stages.

The amplifiers used in facsimile systems must have a uniform response over the frequency spectrum included by the audio carrier and sidebands. This will be true for both amplitude modulation and CFVD systems. Also, the time delay in the amplifiers and transmission lines must be constant for all frequencies. This is necessary in order to prevent displacement of the frequencies composing an element in the picture.

The past few paragraphs have covered the essential details of most of the methods of facsimile transmission in use today in the United States. We will next study the circuits for facsimile reception and the methods of converting the received electrical impulses into a replica of the transmitted copy.

8. **FACSIMILE RECORDERS.** The facsimile receiver makes a permanent record of the transmitted copy. There are many methods of recording used. The four most popular are: first, photographic recorders; second, ink recorders; third, electrolytic recorders; and fourth, carbon paper recorders.

In order for the reproduced copy and the transmitted copy to have the same proportions, it is necessary that the "facsimile indices" of the transmitter and receiver be the same. Facsimile index has the same meaning as aspect ratio in television, but it is expressed a little differently. The facsimile index is defined as the product of the length of the scanned line in inches, and the number of lines per inch of copy. If the transmitter scans 100 lines per inch with a line length of ten inches, the facsimile index is 10 times 100, or 1000. The recorder must have the same facsimile index. If the length of the line at the recorder is 5 inches, then there must be 200 lines scanned per inch, and the facsimile index will be 5 times 200, or 1000. The "scanning line rate" at the recorder and transmitter must be the same if the recorder is to reproduce the transmitted copy. The scanning line rate is equal to the number of lines scanned per second. Other factors, such as the method of synchronization and line return time at the receiver and transmitter, determine whether or not a recorder will reproduce the copy.

The recording scanners, like the transmitting scanners, are of two general types, those that handle continuous copy, and those that handle a single sheet of copy at a time. The continuous scanners do not have to be reloaded each time that a page of copy is reproduced.

The single sheet recorders use the simple drum scanner similar to those used for transmission. Recorders built for facsimile reception in the home use the continuous copy type of scanner.

The best reproduced copy is made by photographic recorders. Also, the photographic recorders are faster than the other types. Fig. 17 shows schematically the essentials of a photographic recorder. The simple drum type scanner is used. A sheet of photosensitized paper or film is wrapped around the cylinder. It is scanned by a spot of light with a cross sectional area equal to a picture element. The scanning process is identical to that of transmitting type drum scanners. The intensity of the light spot is varied according to the light and dark shades of the copy scanned at the transmitter. The transmitter and receiver drums rotate in synchronism. After transmission is completed, the exposed film or paper is removed from the scanner and the latent image is made permanent by developing and fixing.

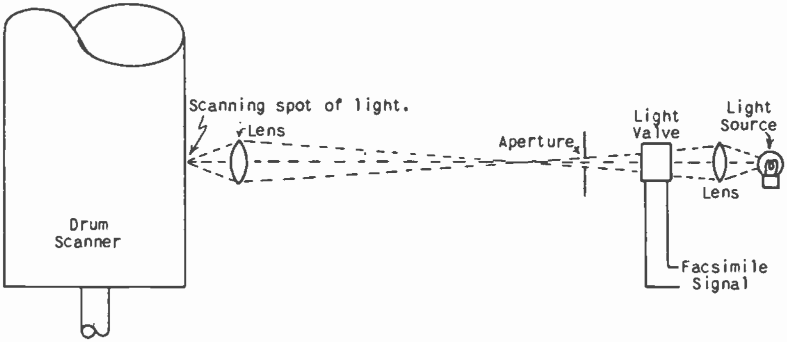


Fig.17 Principle of operation of a photographic recorder.

The intensity of the light beam can be varied electrically or mechanically. In mechanical systems, the intensity of the light source is maintained constant and the amount of light reaching the photosensitized film or paper is varied by a light valve. This method is shown in the figure. There are several types of light valves. One of the light valves used in this country is somewhat similar to the light valve previously described in the section of this lesson on modulating systems. The amount of light passing through an aperture is controlled by moving a shutter over more or less of the aperture. This shutter consists of a light metallic ribbon moved by a magnetic field. The magnetic field is supplied by the rectified facsimile signal (carrier frequency has been removed). The ribbon and its associated circuits must have a uniform frequency response over the range of frequencies from zero to the maximum present in the signal. There are many other types of light valves such as the Kerr cell and the supersonic cell. A description of these cells was given in a preceding lesson of this unit.

In the electrical systems, the intensity of the light source is varied with the changes in amplitude of the facsimile signal. An ordinary incandescent lamp cannot be used for this purpose, as it is too sluggish in following changes in the applied voltage. Neon lamps and similar lamps which radiate light by luminescence are used. The light intensity of such light sources can follow the rapid changes of the applied voltage. Those suitable for facsimile recording are usually designed to produce a point source of very intense illumination.

The photographic recorders, as mentioned before, are the most rapid recorders and produce the best replica of the original. However, the operator has no way of checking his adjustments until the picture or copy has been transmitted and the exposed film or paper developed and fixed.

The other forms of recorders reproduce the copy as it is received. The operator is able to correct his adjustments whenever necessary. Fig. 18 shows schematically the operation of one form of ink recorder developed by RCA. The simple drum type scanner is used. A sheet of paper is wrapped around the drum. It is scanned by a fine ink spray projected from a nozzle. The ink spray is modulated by moving a small vane to deflect more or less of the ink issuing from the nozzle. For white, all of the ink coming from the nozzle is deflected so that none reaches the paper; and for black, all of the ink reaches the paper. The deflected ink is collected. The unit moving the vane is similar in construction to the balanced armature type of magnetic speaker. There are two driving coils. One has maximum current on white and none on black; the other has maximum current on black and no current on white. The fields of the two coils are opposite. In this way the vane is made to follow the signal very faithfully. The coils that operate the vane are fed from a special push-pull circuit that will be described later.

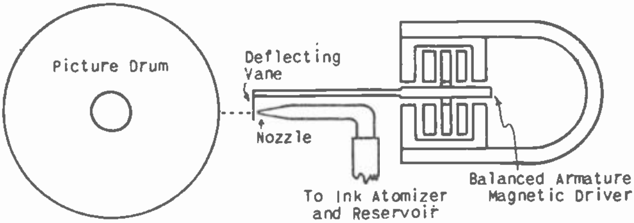


Fig.18 Principle of operation of an ink recorder.

The ink recorder produces an excellent replica of the original. Ink recorders are used on the RCA long distance facsimile circuits.

The electrolytic recorder makes use of the decomposition and change of color of some chemicals when an electric current is passed through them. Chemicals like starch iodide, and some silver and iron salts turn dark when under the influence of an electric current. Paper that has been treated with one of these special chemicals is used with the recorder. Some of the types have to be dampened before using; others function when dry.

Electrolytic recorders are made in both the simple drum scanner type and the continuous copy type. In the drum type scanner a sheet of the specially prepared paper is wrapped around the drum. The drum is a conductor. The paper is scanned by a stylus which rides on the paper as the drum rotates. The rectified facsimile signal is applied between the stylus and drum. The intensity of the change in color of the paper is proportional to the current density flowing between the stylus and drum. Thus a replica of the original is reproduced.

Fig. 19 shows schematically a continuous copy electrolytic recorder. The specially prepared paper is in a roll, and the recorder prints continuously without reloading after each page. The paper is fed between a printer bar and a non-conducting rotating drum. On the surface of the drum is a one-turn raised helix. The single turn begins at one end of the drum and ends at the other. The printer bar presses the paper against the helix. As the drum rotates in the direction shown, the point of pressure between the printer bar and helix moves across the paper from right to left. If the rectified facsimile signal is fed between the printer bar and helix, a line of the original copy will be reproduced each time the drum makes one rotation. The paper is pulled forward the width of a scanning line for each rotation of the drum.

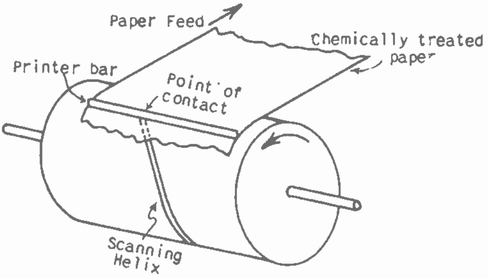


Fig.19 Principle of operation for an electrolytic recorder.

Finch has developed an electrolytic recorder which uses the same scanning system as used in his transmitter shown in Fig. 4. The scanning arm has a stylus that rides on the paper instead of a light source and photocell as on the transmitter. The paper is black with a white or orange surface coating. The electrolytic action removes the surface coating. The amount removed is proportional to the current density. The reproduced copy appears as black on a white or orange background.

In Fig. 20 is shown a schematic of the RCA carbon recorder. This is a continuous copy type of recorder. Scanning is accomplished by moving a sheet of paper between a printer·bar and a raised helix on a rotating drum. As described in a previous system, the paper is moved forward past the printer bar the width of a scanning line for each rotation of the drum. A sheet of carbon paper is also moved between the printer bar and the helix. The transfer side of the carbon paper is next to the recording paper. In this recorder the pressure of the printer bar on the helix is varied with the facsimile signal. The pressure is maximum for black and minimum for white. The pressure of the printing bar on the helix prints a black dot on the paper. The density of the dot will be proportional to the pressure. The carbon paper and the white paper are in close contact only when under the printer bar. The two are fed in together, but they are separated immediately after printing. The used carbon paper is rolled up. The carbon paper is moved through the printer one-fourth as rapidly as the white paper. In this way a roll of carbon paper is utilized completely on one passage through the printer.

The action of the printer bar is controlled by a push-pull magnetic driver similar to that used on the ink recorder previously described. Since considerable pressure is required on the printer bar, two or three balanced driving units are used on one printer bar. Each driving unit has two coils; one has maximum current on black, and no current on white; the other has maximum on white, and no current on black. The fields of the two coils are opposite. On a black signal, all the coils carrying current force the printer bar against the helix, and on white the remaining coils lift the printer bar from the helix.

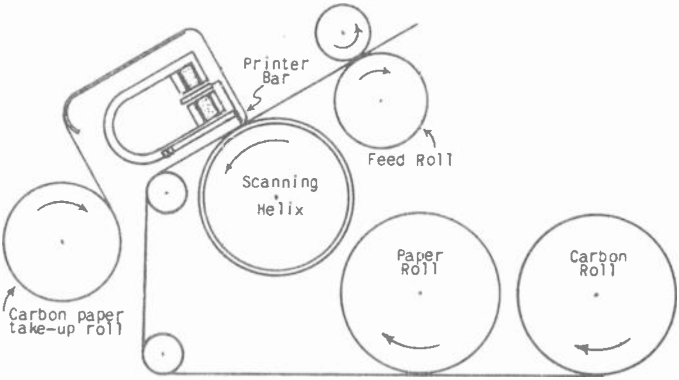


Fig.20 Principle of operation of a carbon paper recorder.

Most tape recorders also use carbon paper. In these, the carbon paper and record paper are about one-half inch wide. The printer bar is very light, and consequently the recorder is quite fast.

In order to operate any of the recorders described, the input facsimile signal must be free of the audio carrier. The carrier and its sidebands must be rectified and the carrier filtered from the signal before it is suitable for operating the light modulator or printer of the recorder.

The circuits used in the reception of a facsimile signal are usually quite simple. In the systems where synchronizing impulses are sent along with the signal, the circuits are a little more elaborate. If wire transmission is used, the received signal will be either an amplitude modulated audio frequency carrier or a CFVD signal. If radio transmission is used, the received signal will be an RF carrier modulated with the audio carrier and its sidebands either amplitude modulated or CFVD, or it can be an RF carrier, keyed by a rectified and filtered CFVD signal. In the case of the former, the output of the second detector of the receiver will have the same form as a signal transmitted via wire. In the case of the latter, the output of the second detector is in the correct form for operating a light valve or electromagnetic printer.

The output of the telephone line or second detector is very small and must be amplified several times before its magnitude is sufficient to operate a recorder. When the output consists of an audio carrier and sidebands, conventional AC-coupled amplifiers can be used. If a keyed RF carrier has been used as the transmission medium, it is best to use DC amplification after the second detector or use the output of the detector to operate a balanced modulator similar to those previously described. This latter method will be required if the recorder is located many miles from the radio receiver and the output of the receiver is sent to the recorder via wire lines.

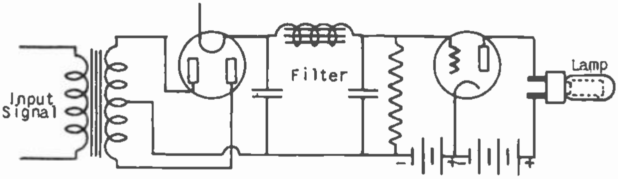


Fig.21 Circuit to operate an electrolytic or photographic recorder.

Fig. 21 shows a rectifier-filter unit suitable for operating a light valve or neon lamp photographic recorder, or an electrolytic recorder. The output of the filter must contain all frequencies from zero to the maximum present in the facsimile signal. However, the carrier frequency must be completely eliminated. Therefore, the filter must have a uniform pass band from zero to the maximum frequency in the facsimile signal and it must have a sharp cutoff in order to remove all traces of the audio carrier. The output of the filter is direct-coupled to the grid of the output tube. A neon lamp, light valve, or the input terminals of an electrolytic recorder is connected in series with the plate of the output tube.

The audio carrier is negatively modulated; that is, its amplitude is maximum for black. In the circuit shown in Fig. 21, a maximum positive voltage will be applied to the grid of the output tube for black. If the output tube is biased to cutoff for zero input, there will be no plate current for white and maximum plate current for black. These conditions will produce a positive picture for electrolytic recording, a positive or negative picture for a light valve photographic recorder (depending on the adjustment of the light valve), and a negative picture for a neon lamp photographic recorder. The reproduced picture can be reversed by reversing the connections from the rectifier unit to the output tube.

In Fig. 22 is shown a schematic of the circuit designed to operate a balanced armature magnetic printer such as used on the ink and carbon paper recorders. In these recorders there are two magnetic fields acting on the armature which controls the movement of the deflecting vane in the ink recorder and the printer bar in the carbon paper recorder. One field is maximum for black and zero for white, and the other is maximum for white and zero for black.

The voltage developed across the load resistance for the rectifier in Fig. 22 is zero for white and maximum for black. The top end of the resistor will be negative with respect to the bottom end for a black signal. With no input, there is zero bias on the upper tube of the push-pull stage, and its plate current is maximum. For the same condition, the fixed negative bias on the grid of the other tube in the push-pull stage is sufficient to bias it to cutoff. Therefore, for no input, or white, the current through the upper coil is maximum, and the current through the lower coil is zero. With full input, or black, the negative voltage applied to the grid of the upper tube is sufficient to bias it to cutoff; the positive voltage applied to the grid of the lower tube is equal to the negative fixed bias, and the plate current is maximum. Thus, the current through the upper coil will pull the printer bar away from the paper in the carbon printer recorder and deflect the ink away from the paper in the ink recorder. The converse will happen when the current is maximum in the lower coil.

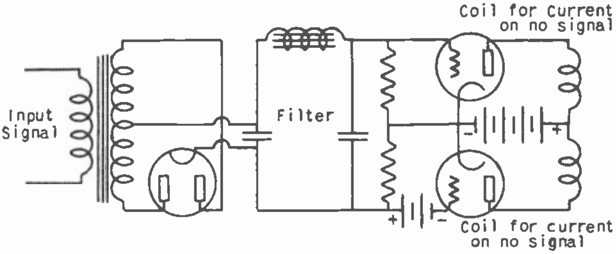


Fig.22 Circuit to operate printer bar type recorder.

When black and white drawings, printing, or CFVD half-tones are transmitted via radio, it is often necessary to reshape the received signals before satisfactory recordings can be made. This can be done very conveniently by using a threshold-limiter amplifier like the one shown in Fig. 12. The receiver circuits are designed so that the threshold-limiter amplifier can be switched in or out as desired.

8. FACSIMILE SYNCHRONIZATION. There are two problems involved in synchronizing the recorder with the transmitter. The first is to maintain the same scanning rate (the number of lines scanned per second) at both the receiver and the transmitter. The other is to keep the recorder in frame with the transmitter; that is, the recorder must start a new line at the same instant that the transmitter starts to scan a new line.

The first problem is the most easily solved. When the transmitter and recorder are in an area supplied by a common AC power system or by interconnected systems, the recorder and transmitter are operated by synchronous motors. The scanning speeds of the transmitter and recorder will always stay in step. When long distance transmission is used, either by radio or by telephone lines, the driving motors at the transmitter and recorder are controlled by local oscillators whose frequency is maintained within very close limits. Another method that is sometimes used is to transmit the motor control frequency right along with the facsimile signal. This complicates the design of the equipment. A fourth method that is used, is called start-stop synchronization. In this system there is a fairly large interval between the end of one line and the beginning of the next line. The recorder stops at the end of each line and begins the next line when a starting impulse is received from the transmitter. The recorder motor is governor-controlled and runs continuously during the transmission. The recorder and transmitter start each line together, and slight differences in scanning speeds will not destroy the detail of the picture.

Framing is done automatically in the last method given for maintaining the same scanning rates at the transmitter and receiver. In the others, the problem of maintaining proper framing is separate. When the scanning rates or drum speeds are kept constant, the framing will be correct if the receiver and transmitter are in step for the first line of the transmitted copy. In commercial facsimile, there is an experienced operator present at each receiving station and he can make sure that the receiver is in step with the transmitter at the start of a transmission. For home facsimile, the framing must be automatic.

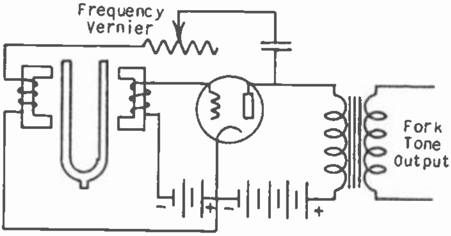


Fig. 23 Tuning fork controlled oscillator.

The most common method of getting the receiver in frame in commercial facsimile systems is for the operator to slow the drum speed slightly until the receiver is in frame. Usually there are commutators on the transmitting and receiver scanners which produce a sharp voltage pulse whenever the scanning aperture passes over the clamp that holds the paper on the drum. This clamp is parallel to the axis of the drum. The impulse from the transmitter is sent with the facsimile signal. These impulses occur from 60 to 75 times per minute, and the operator slows the receiving drum until the impulse from the receiver is in step with the one from the transmitter. There is usually a clutch connecting the drum to the driving motor, and the operator allows the clutch to slip slightly until the receiver is in step with the transmitter.

In other commercial systems the clutch is automatic and can be controlled by an impulse from the transmitter. The receiver operator loads the receiver drum and sets the receiver scanning aperture over the clamp on the drum. The transmitter operator does the same. The driving motors are started a few minutes ahead of the transmission so that they can get properly synchronized with their local frequency standard. When the operator at the transmitter is ready to start transmission he closes a switch and sends an impulse which causes the clutches at both the receiver and transmitter to engage at the same time. The two are in frame and stay that way for the transmission if the motor speeds remain the same.

In commercial systems that maintain local frequency standards, temperature-controlled tuning forks are the usual form of standard used. The tuning fork is used as the feed-back element in a vacuum tube oscillator. The grid and plate of the oscillator are coupled together by means of the tuning fork. Fig. 23 shows the simplified circuit diagram of such an oscillator. The variable resistance and condenser between the grid and plate serve as a variable feed-back control. The resistance can be used to control the amplitude of the oscillator and the vibration of the tuning fork. It will also have a small control over the frequency. The stability of such a standard is greatest when the amplitude of the tuning fork vibration is least. The tuning fork is kept in a temperature-controlled oven. The frequency variation of such a standard can be kept within 1 part in 100,000.

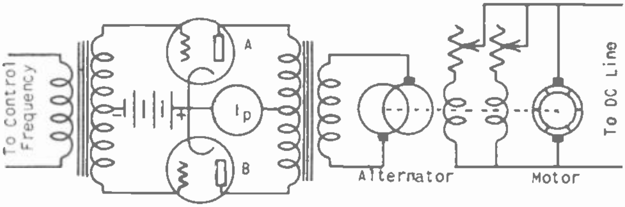


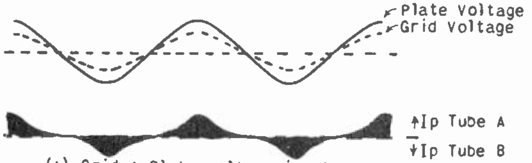
Fig. 24 Circuit of automatic brake.

There are two systems in general use for controlling the speed of the scanning motor by the frequency standard. In one, the scanning motor is DC-operated and its speed is varied by an automatic brake, controlled by the frequency standard. The other uses a high power oscillator, controlled by the frequency standard. The output of the oscillator is sufficient to operate a synchronous motor.

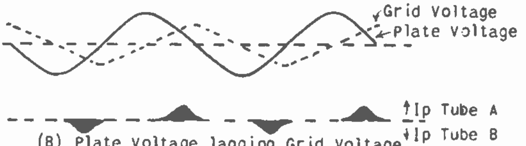
Fig. 24 shows the principle of the automatic brake. The driving motor has an AC alternator on the same shaft. The motor has poor regulation, and variations in load will change its speed. The output of the alternator is impressed on the plates of the push-pull stage. The alternator is the only source of plate voltage. The output of the frequency standard is impressed on the grid of the push-pull stage.

When the motor is running at the correct speed, the frequency of the alternator is the same as the standard frequency. The fixed bias on the grids of the push-pull stage biases the tubes to cutoff when there is no input from the frequency standard. Then the plate current and the load on the alternator and motor will depend on the phase between the plate and grid voltages. The power taken will be maximum when the voltages are in phase and zero when they are completely out of phase. Fig. 25 shows the variation in plate current of the tubes with changes in phase of the applied grid and plate voltages. In Fig. 25A, the plate and grid voltages are in phase and the plate current and load on the motor is maximum. In Fig. 25B, the plate voltage is lagging behind the grid voltage, and the plate current and load on the motor is less.

For normal line voltage and scanner operation, the apparatus can be adjusted so that the plate voltage lags the grid voltage by 90°. If the motor would start to slow down slightly because of decreased line voltage or some other cause, the frequency of the alternator will become slightly slower than the control frequency, and the plate voltage will begin to lag more and more behind the grid voltage. As the plate voltage lag starts to increase the plate current and load on the alternator will become less. This will continue until the load is reduced sufficiently so that the motor can run at its normal speed, and the plate voltage will lag the grid voltage by more than 90°.



(A) Grid & Plate voltage in phase for maximum loading.



(B) Plate Voltage lagging Grid Voltage by 90°.

Fig.25 Plate current waveforms for two different phase relations between grid and plate voltages.

If the speed of the motor should tend to increase because of increased line voltage or other cause, the frequency of the alternator will become slightly higher than the standard frequency, and the plate voltage will start to lag less and less behind the grid voltage. As the plate voltage lag starts to decrease, the plate current and load on the motor will become greater. This will continue until the load is increased sufficiently so that the motor will run at its normal speed, and the plate voltage will lag the grid voltage by less than 90°.

Thus, the load on the motor can be varied to compensate for line voltage variations and other causes. This system will maintain the speed of the motor very accurately.

In the other system, a push-pull gas triode oscillator is used to supply the power to operate the synchronous motor driving the scanning drum. (The operation of the gas triode was discussed in Lesson 5 of Unit 5.) Fig. 26 is a diagram of a gas triode oscillator. The circuit is similar to a conventional self excited push-pull oscillator.

Its operation can be explained as follows: Let us assume that the left hand tube fires through some voltage transient. The plate current of that tube will start to rise to the peak value determined by the resistance in the external plate circuit. The rise will not be abrupt because of the inductance of the plate reactor and plate transformer. The plate voltage will immediately drop to the minimum value necessary to maintain ionization in the gas triode. Since the plate of one tube is at the plate supply voltage and the other is several volts lower, the condenser in the plate circuit will charge to the difference in the voltage applied to the plates of the two tubes. This charging current, while flowing through the grid transformer, will induce a voltage in the secondary of the transformer that will swing the grid of the right tube more positive. This positive voltage swing will cause the right tube to fire. Its plate voltage will immediately drop to the minimum value necessary to maintain ionization of the tube. Since the condenser cannot discharge instantaneously, the voltage across the condenser will immediately swing the plate voltage of the left tube negative for a fraction of a second. During this interval, the left tube will de-ionize and its grid will regain control. The condenser will now charge in the opposite direction to its previous charge and the

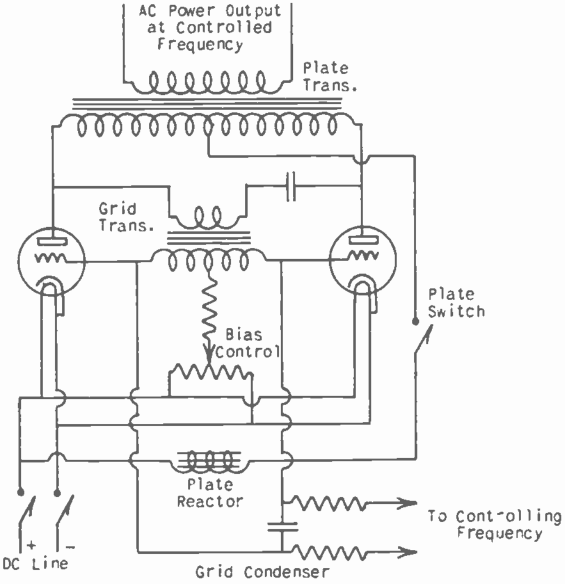


Fig. 26 Gas triode oscillator.

flow of charging current will swing the grid of the left tube in the positive direction and cause it to fire. Thus, the tubes are made to fire alternately. The frequency of the firing is determined by the inductance and capacitance in the plate circuit. By careful adjustment, the output waveform can be made sinusoidal; however, the plate current of the tubes is in the form of rectangular pulses.

By injecting into the grid circuit a voltage of a slightly higher frequency than the frequency of the oscillator, the oscillator will lock in with the control frequency with perfect synchronism.

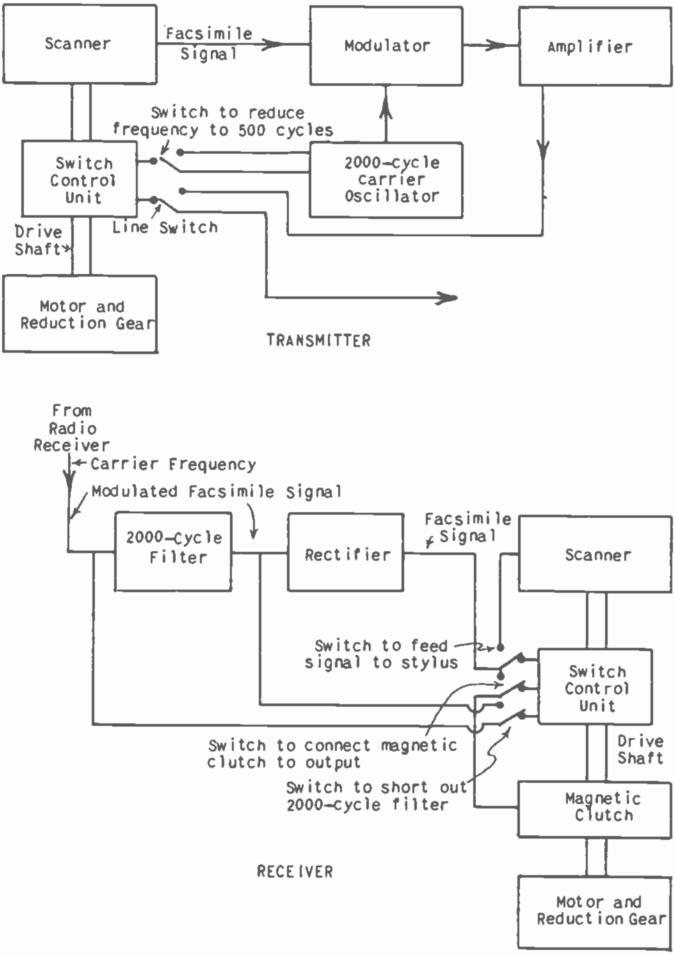


Fig.27 Block diagram of Finch transmitter and receiver.

The output power of this oscillator is great enough to operate a synchronous motor which in turn drives the scanner.

Sometimes the control frequency is transmitted along with the facsimile signal. This is usually not satisfactory, as sharp bursts of static will destroy synchronism for a few cycles.

In this section, the principal methods of maintaining synchronization between the transmitter and receiver have been discussed. In the next section, a short description of the principal facsimile systems in operation in the United States will be given.

9. FACSIMILE SYSTEMS. The main facsimile systems of general interest are those developed for: (1) home broadcast service, (2) Wirephoto service for newspapers, and (3) long distance radio communication. At this writing, there are two systems in general use for home broadcast service; the Finch and RCA systems. The wirephoto system is a developnent of the Bell Laboratories. The system for long distance radio communication that is described in this section is a product of RCA.

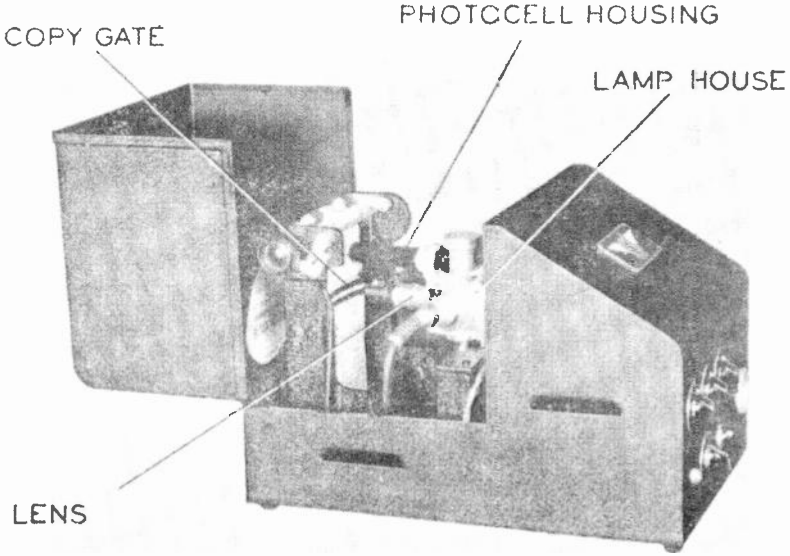


Fig.28 Finch Transmitter.

The Finch system is suitable for the transmission of half-tones and black and white copy over wire lines or in the primary coverage area of a broadcast station. Half-tone effects are obtained by varying the amplitude of modulation of a 2000-cycle audio carrier. Copy is scanned with a detail of 100 lines per inch and the scanning lines are four inches long. The copy is scanned at the rate of one line per second. The scanning and return time are each one-half second. During the return time, the copy is advanced the width of a scanning line at both the transmitter and receiver. Start-stop synchronization is used.

Since a complete scanning cycle requires one second, and the actual scanning time is one-half second, there are 400 picture elements transmitted in one-half second. Thus, the rate of scanning is 800 picture elements per second and the maximum frequency is 400 cycles.

Fig. 27 shows a block diagram of a Finch transmitter and recorder. The scanners are the same as the type illustrated in Fig. 4. Synchronous motors operating at a speed of 1800 r.p.m. are used to drive the scanners. This speed is reduced to 60 r.p.m. through a 30 to 1 reduction gear. The motor drives the scanning arms, advances the paper at the proper time, and operates switches which control the transmission of the signal and synchronizing impulses.

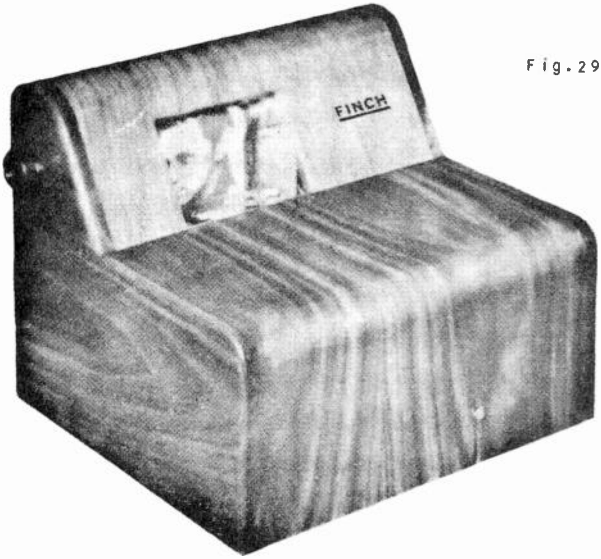


Fig.29 Finch Recorder.

When a scanning cycle starts at the transmitter, the switch feeding the signal to the line is closed and another switch shunts a condenser across the inductance of the 2000-cycle carrier oscillator for one-fiftieth of a second. During this interval, the output frequency of the oscillator is 500 cycles. This short 500-cycle note is used to synchronize the recorder with the transmitter. At the end of the scanning stroke, the line switch is opened. When the return stroke is half completed, a cam operates a ratchet which advances the copy the width of a scanning line or .01 inch. At the start of the next scanning line, the cycle is repeated.

The recorder has a magnetic clutch which couples the scanner mechanism to the driving motor. When there is no signal being received from the transmitter, the clutch is not engaged, and the motor runs free. Also, for the no-signal condition, the output of the rectifier is connected to the magnetic clutch, and a 2000-cycle filter is inserted in the line from the radio receiver to the rectifier. When a signal is being received from the transmitter, no signal will reach the rectifier until the short 500-cycle pulse generated at the start of a scanning line at the transmitter is received. The 2000-cycle filter prevents the carrier frequency from reaching the rectifier. When the 500-cycle pulse is received, it is rectified and the resulting DC is applied to the magnetic clutch. The clutch engages, and the scanning arm starts to sweep from left to right. At the same time, a switch connected to the drive shaft switches the output of the rectifier from the magnetic clutch to the stylus on the recorder. Another switch connected to the same drive shaft shorts the 2000-cycle filter out of the line to the rectifier. The signal is applied to the stylus for the remainder of the line and the facsimile signal is recorded electrolytically on the specially prepared paper (the nature of this paper was described in the description of this type of recorder given earlier in the lesson). At the end of the scanning stroke, the line switch at the transmitter opens the circuit and the printing stops. At the center

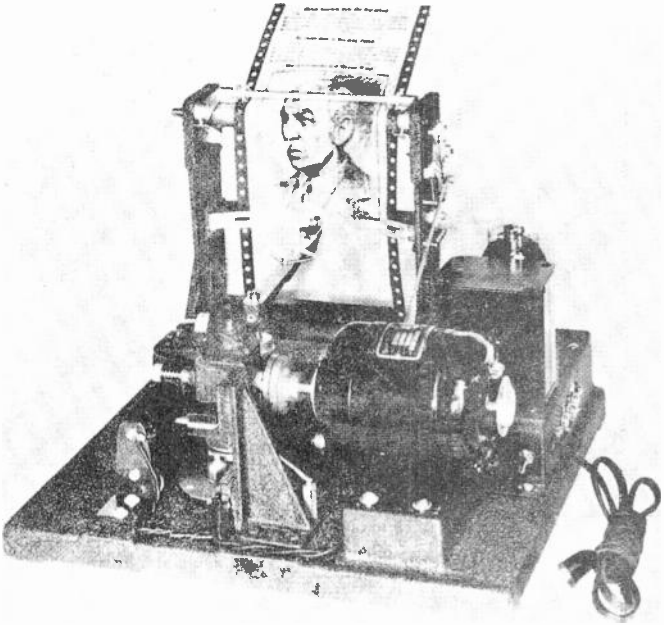


Fig.30 Finch recorder with cover removed.

of the return stroke, the paper in the recorder is shifted the width of a scanning line, the 2000-cycle filter is reinserted in the line between the receiver and rectifier, and the output of the rectifier is switched from the stylus to the magnetic clutch. At the end of the return stroke, the clutch disengages and the recorder stops. The clutch will not re-engage until another 500-cycle pulse is received from the transmitter. If the recorder is running slightly slower than the transmitter, it will synchronize perfectly with the transmitter; if the recorder is running faster than the transmitter, it will not synchronize with the transmitter.

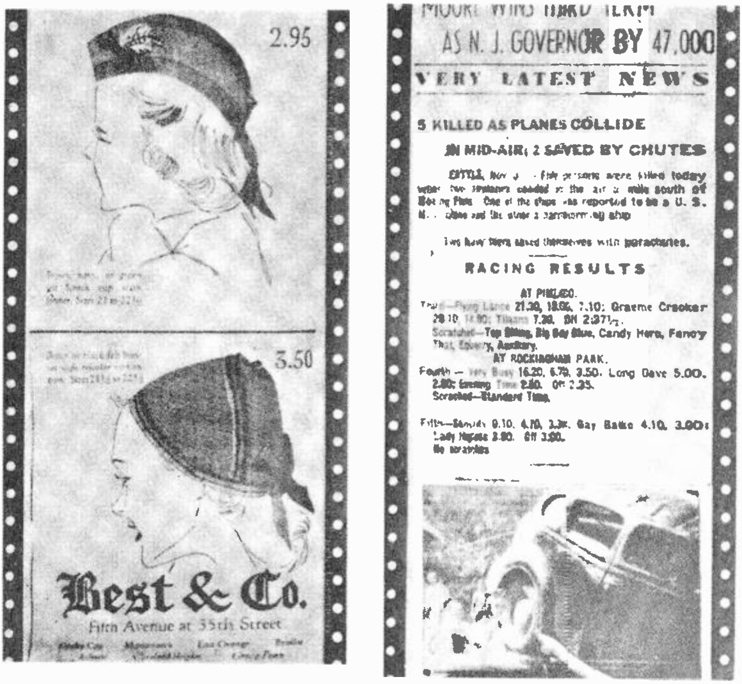

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Fig.31 Sample of a Finch recording.

Fig. 28 is a picture of a Finch facsimile transmitter. The output will feed a 500-ohm line at zero decibels. Fig. 29 is a picture of a Finch recorder that is used with an ordinary broadcast receiver. Fig. 30 is the same recorder with the cover removed. Fig. 31 is a photograph of part of a recording made by a Finch recorder.

The RCA system for home use produces an amplitude modulated half-tone suitable for transmission over telephone lines and broadcast stations. Copy is scanned with a detail of 125 lines to the inch. The length of a scanning line is 8½ inches. The copy is scanned at the rate of 75 lines per inch, or 1¼ lines per second. The number of picture elements scanned per second is approximately

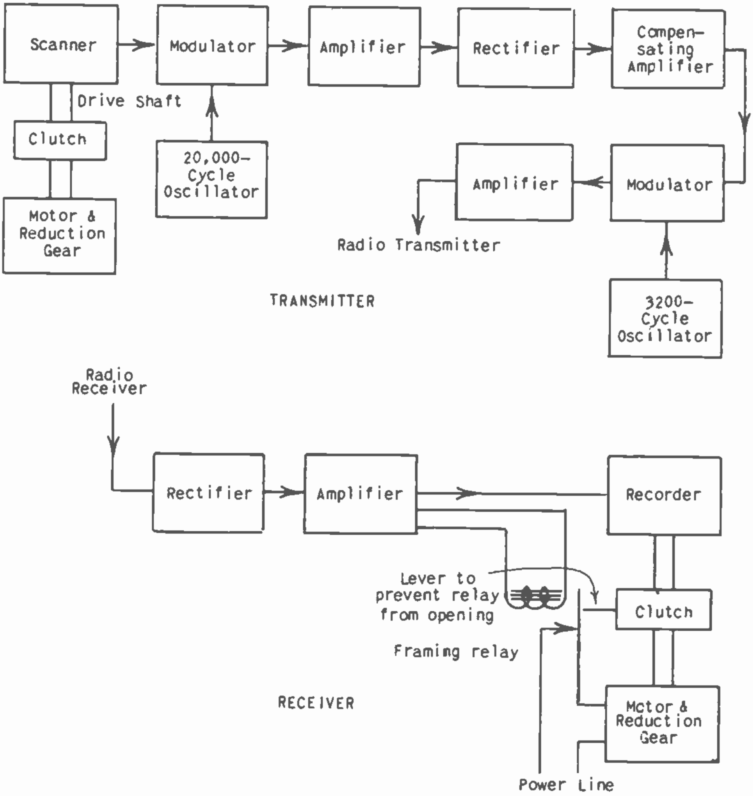

Fig.32 Block diagram of RCA transmitter and receiver.

1500, and the maximum frequency is 750 cycles. Seven and one-half inches of the scanning line can be used for copy. During the remaining inch, the framing impulse is transmitted. Synchronization is maintained through the use of synchronous motors operating from the same power system or interconnected systems. The carrier frequency is 3200 cycles.

A drum scanner is used at the transmitter. The scanner will handle a 12 inch sheet of copy. A continuous copy recorder is used at the receiver. The recorder uses carbon paper for recording, and is of the type shown in Fig. 20.

Fig. 32 shows a block diagram of the RCA transmitter and receiver for broadcast use. The scanning drum is driven by a synchronous motor through a 24 to 1 reduction gear. There is a single position clutch between the drum and driving motor. The relation between the position of the scanning aperture and phase of the supply voltage will always be the same when the clutch is engaged. Thus, the scanning drum can be stopped and started without getting out of frame with the recorder.

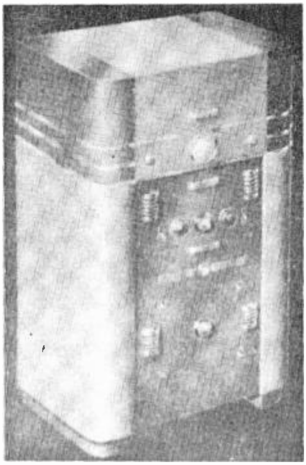

Fig.33 RCA Transmitter.

The output of the photocell is amplified by a single stage DC amplifier, and the output of the amplifier is used to modulate a 20,000-cycle carrier. The modulated carrier is amplified and is then rectified. The rectified output which contains the same frequencies originally produced in the photocell is fed into a compensating amplifier. The purpose of the compensating amplifier is to modify the facsimile signal in order to compensate for the lack of linearity in the carbon paper recorder over the entire range from black to white. The output of the compensating amplifier is

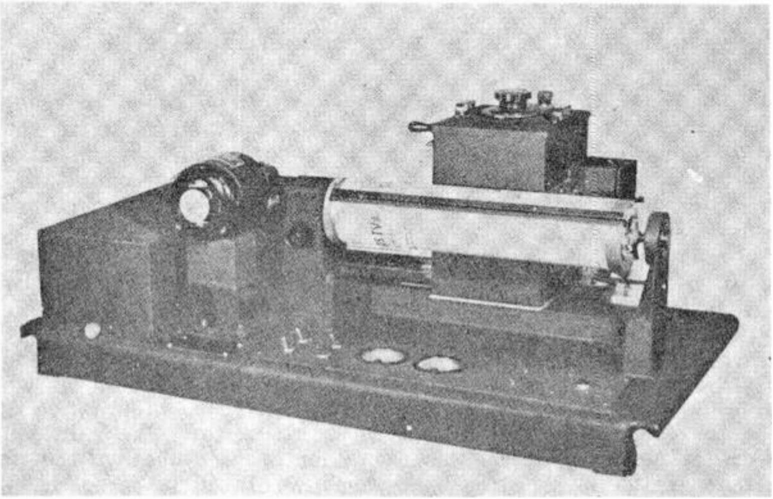

Fig.34 Scanner in RCA transmitter.

used to modulate a 3200-cycle carrier. The 3200-cycle carrier and its sidebands are sent out via a wire line or a broadcast transmitter.

The radiated signal is picked up by an ordinary broadcast receiver. The output of the second detector is amplified and applied to a rectifier and filter. The output of the rectifier is applied through a push-pull output stage to the printer bar on the recorder. Since the recorder and scanner are driven by synchronous motors operating on the same power system, the scanning rate at the transmitter and receiver are the same. Framing is accomplished by the operation of a framing relay on the recorder. This relay breaks the power circuit to the motor when an impulse is received from the transmitter as the scanning aperture passes over the clamp holding the copy to the drum. If the recorder is in frame with the transmitter, the relay is prevented from opening the motor circuit by a lever which clamps the armature during the interval between the end of one line and the start of the next on the recorder. If the recorder is out of frame, the motor circuit is opened momentarily during the interval between lines on the recorder and the motor loses a few revolutions. This is repeated at the end of every line until the recorder is in frame with the transmitter. Framing takes place only at the start of a transmission schedule and does not have to be repeated unless there is a power failure or loss of signal through fading.

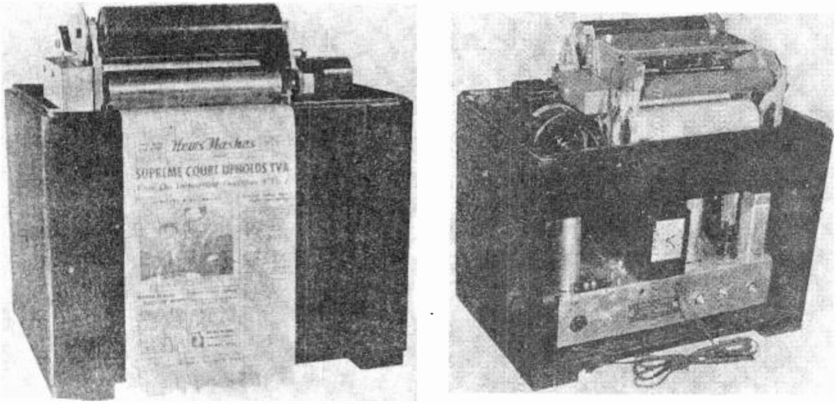

Fig.35 Front and back views of RCA recorder.

Fig. 33 is a picture of the complete transmitter unit. Fig. 34 is a picture of the scanner. The output can be fed into the main amplifiers in the control room of a radio station. Fig. 35 shows a front and back view of the recorder. Fig. 36 is a picture of copy reproduced by the RCA recorder. The recorder includes a fixed tuned radio receiver and a time clock. It is preferable to use a separate receiver with the recorder rather than a regular broadcast receiver because the recorder receiver is always properly tuned for the station carrying the facsimile signal.

Broadcast News CARTOON SECTION Page 3

"Accordin' to my diary, I didn't rob no bank on the fifteenth of July!"

Fig.36 Sample of RCA recording.

The facsimile system used for the transmission of newspaper pictures is a development of the Bell Laboratories. It is called "Wirephoto". The system utilizes telephone lines for the transmission of amplitude modulated half-tones and black and white copy. Drum scanners are used at both the transmitter and receiver. Photographic recording is used. The scanners will handle copy up to 17 by 14 inches. The copy is scanned with a 100 lines per inch definition. The scanning line is 11 inches long. The scanning rate is 100 lines per minute. The maximum frequency generated is 1000 cycles. The carrier frequency is 2400 cycles. Photographic recorders are used. The scanning rates at the transmitter and receiver are synchronized by the use of temperature-controlled tuning fork frequency standards. Framing is accomplished by starting the receiver drums and transmitter drums simultaneously by an impulse generated at the transmitter.

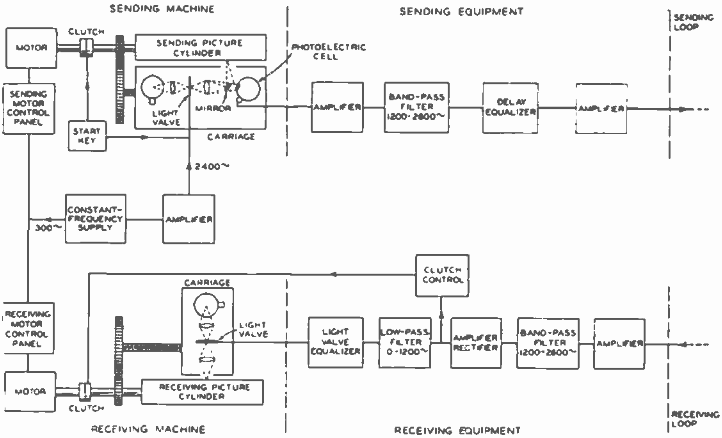

Fig.37 Block diagram of transmitter and receiver used for transmission of newspaper pictures.

Fig. 37 shows a block diagram of the transmitter and receiver. Since this system is used for the transmission of newspaper photographs, there is a transmitter and receiver at each station. The frequency standard is a 300-cycle temperature-controlled tuning fork oscillator. The oscillator circuit is similar to the basic circuit described previously in the lesson. The driving motors operate on DC, and their speed is controlled by the frequency standard by means of an automatic brake similar to that shown in Fig. 24. There is a magnetic clutch coupling the driving motor and the drums. The carrier frequency of 2400 cycles is obtained by multiplying the 300-cycle standard. The carrier frequency is introduced by modulating the scanning light beam with a light valve similar in design to those shown in Fig. 10. The output of the photocell is a 2400-cycle carrier, amplitude modulated by the picture signal. The output of the photocell is amplified and is passed into a filter which removes the upper sideband. This is necessary, as ordinary phone lines will not pass a wide enough range to include both sidebands. The output of the filter must be passed through a delay equalizer which corrects phase shifts caused by the sideband filter. The output of the equalizer is amplified and is transmitted to the receiver over telephone lines.

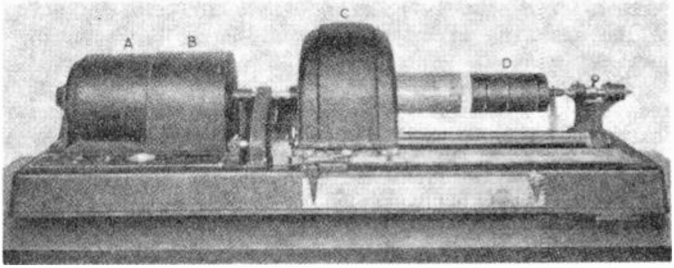

Fig.38 Scanner.

The received signal from the telephone lines is amplified and passed through a filter to remove line noise with frequencies outside of the required transmission band. The output of the filter is amplified and rectified. The rectified output is passed through a low-pass filter to remove the carrier. The resultant signal modulates the scanning light beam of the recorder by means of a ribbon light valve. In order to make the response of the light valve uniform over the entire range of frequencies in the signal, a light valve equalizer must be used. The equalizer distorts the signal so that the output of the light valve will be uniform over the frequency band.

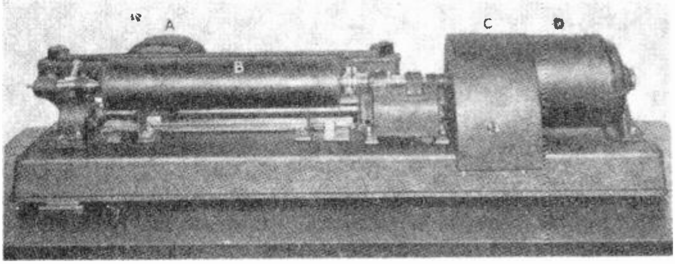

Fig.39 Recorder.

In order to obtain framing, all the scanning drums at the transmitter and receivers are set in position so that the scanning apertures are at the beginning of the first line. Before a transmission starts, all the receiving stations are notified of the fact. The operator at the transmitter presses a key which generates an impulse which is transmitted over the line to all the receivers. This impulse engages the clutches at the transmitter and all the receivers simultaneously. The recorders remain in frame for the duration of the transmission.

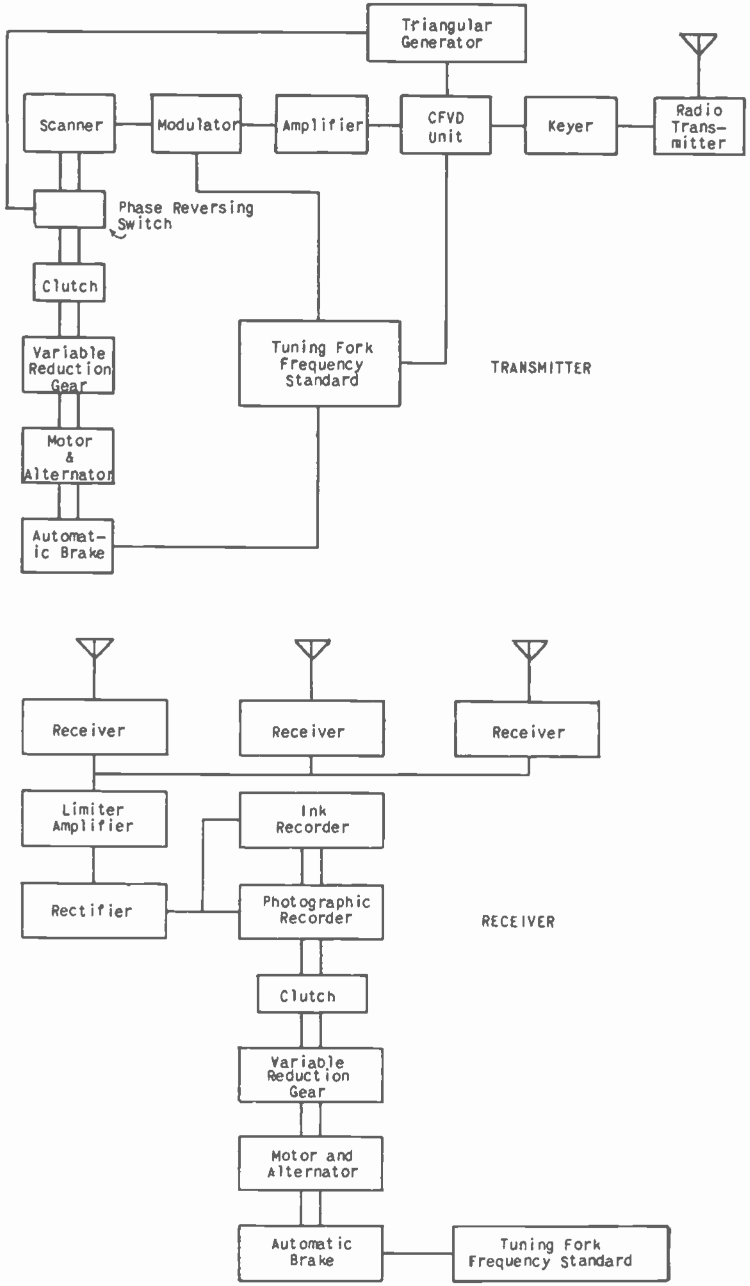

Fig.40 Block diagram of RCA transmitter and receiver for transoceanic work.

Fig. 38 is a picture of the scanner at the transmitter, and Fig. 39 is a picture of the recorder. Examples of the pictures transmitted by this system can be found in most newspapers carrying Associated Press news. These pictures are labeled "Wirephoto".

The last facsimile system to be described is that developed by RCA for trans-oceanic transmission of pictures by radio. The CFVD system is used for the transmission of half-tones. Drum scanners and recorders are used. At the receiving point, both photographic and ink recorders are used. The ink recorder is used so that the picture can be viewed during reception and the required adjustments can be made to compensate for changes occurring during the period of transmission. The photographic recorder produces the best picture.

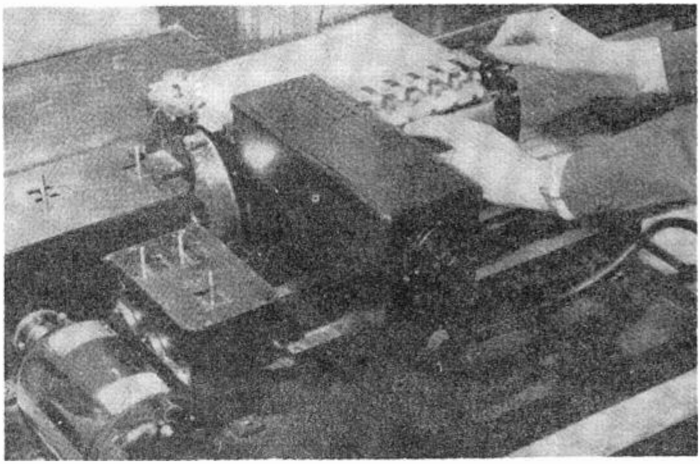

Fig.41 RCA Scanner.

The definition can be varied from 40 to 300 lines per inch in twelve steps. The scanning rate can be varied from 20 to 60 lines per minute in four steps. This variation is necessary in order to handle different transmission conditions. The scanning rates at the transmitter and receivers are synchronized by the use of local frequency standards controlled by temperature-compensated tuning forks. Automatic vacuum tube brakes similar to those used in the Bell systems are used to lock the driving motors in with the frequency standards. Framing is done by letting the clutch of the recorder slip slightly until the recorder is in step with the transmitter. At the beginning of the transmission, impulses are transmitted every time the scanning aperture at the transmitter passes over the clamp holding the picture on the drum. The receiver operator adjusts the phase of the recorder until these impulses occur at the same instant that the recorder aperture passes over the clamp on the recorder drum.

Fig. 40 is a block diagram of a transmitter and receiver. The output of the photocell is used to amplitude modulate a carrier.

This amplitude-modulated carrier is amplified and then converted into the CFVD signal by the method previously described. The output of the CFVD unit is used to key a radio telegraph transmitter by a keyer circuit similar to the one previously described in the lesson. Since the scanning speed is variable, the triangular wave frequency must be varied to fit the scanning speed. The harmonic relation between the triangular wave frequency and the scanning speed in lines per minute is kept constant. In order to insure that the phase of the triangular wave is changed by 180° for alternate lines, a phase reversing switch is incorporated on the scanner drive shaft.

At the receiver, diversified reception is used to minimize effects of fading. The outputs of the three receivers are combined and fed into a limiter amplifier like the one previously described. The purpose of this amplifier is to square up the received impulses and reduce the effect of noise. The output of the limiter amplifier is applied to the two recorders.

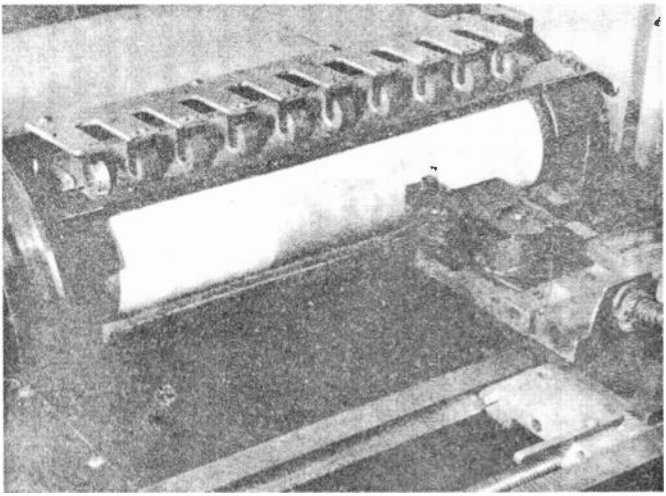

Fig.42 RCA Ink Recorder.

Usually at the start of a transmission, a standard graduated wedge of all the shades between black and white is transmitted. Then the circuits of the transmitter are adjusted so that the CFVD signal at the receiver reproduces all the shades from black to white distinctly. The duration of the received dots can be changed to meet transmission conditions between the receiving station and the transmitting station.

Fig. 41 is a picture of the transmitting scanner. Fig. 42 is a picture of the ink recorder. Newspapers often contain pictures that have been transmitted via radio by this system. They are labeled "Photoradio" or some other similar term.

EXAMINATION QUESTIONS

INSTRUCTIONS. Before starting to answer these examination questions, you should have studied the lesson material at least three times. Be sure that you understand each question--then proceed to write the best answer you can. Make all answers complete and in detail. Print your name, address, and file number on each page and be neat in your work. Your paper must be easily legible; otherwise, it will be returned ungraded. Finish this examination before starting your study of the next lesson. However, send in at least three examinations at a time.

1. What is the difference between facsimile and television?

2. Why is it necessary to amplify a facsimile signal as an amplitude modulated audio tone?

3. (a) What are the two methods of transmitting half-tones? (b) Which of these can be used on long distance radio circuits?

4. What is aperture distortion?

5. What is the value of a threshold-limiter amplifier?

6. What is meant by "facsimile index"?

7. Diagram a circuit that will operate a photographic recorder from the output of an ordinary radio receiver. Explain briefly its operation.

8. Name three methods for maintaining the same scanning rate at the transmitter and receiver.

9. What is the method used to synchronize the Finch receiver and transmitter?

10. In your opinion, which type of service, facsimile or television, will become more popular? Give a reason for your answer.

Notes

(These extra pages are provided for your use in taking special notes)

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