COLLECTED TELEVISION PAPERS

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FOREWORD

This volume is a collection of selected articles on television. These papers are from the series of television volumes published by RCA Review, and have been chosen by instructors at RCA Institutes, Inc. as particularly pertinent for use in their classes.

This one-volume edition has been prepared for the convenience of the students. It presents in one book those of the many papers written by RCA authors on this subject that are essential to supplement textbooks currently available.

The staff at RCA Institutes wishes to extend its thanks to the Society of Motion Picture and Television Engineers and the Institute of Radio Engineers for granting permission to use papers which originally appeared in the publications of those organizations. This book was made possible through the courtesy of RCA Review.

GEORGE F. MAEDEL
July 1950

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TELEVISION TRANSMITTERS OPERATING AT HIGH POWERS AND ULTRA-HIGH FREQUENCIES

BY

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THE advent of high-definition television, involving modulation frequencies up to several million cycles, has necessitated the development of high-power, ultra-high-frequency transmitters. The unique tube and circuit problems encountered and the practicability of line sections as circuit elements has resulted in radical departures from conventions in transmitter design as may be seen from the accompanying illustrations, showing features of high-power, ultra-high-frequency television transmitters.

VACUUM TUBE PROBLEMS

In ultra-high frequency development the vacuum tubes have always been one of the major sources of difficulty. Vacuum tubes developed for lower frequencies have a number of limitations rendering them unsuitable for u-h-f applications. For low-power u-h-f transmitters and receivers, special tubes having low internal capacities, short leads, and other features are available, permitting conventional designs insofar as tubes are concerned. Television transmitters with carrier powers between five and ten kilowatts require tubes with dissipation capabilities of the order of thirty kilowatts. The Type 899 shown in Figure 2, is one of the tubes now used in these applications. Some of the problems of high-power u-h-f transmitters are due to the large physical size of tubes now available.

Water cooled tubes have glass envelopes to provide insulating supports for grid and filament structures in high-power tubes. For manufacturing reasons, these envelopes are made of considerable length since the resulting lengths of filament and grid leads do not present serious difficulties at low frequencies. At ultra-high frequencies the inductance of the leads, plus the loading effects of the inter-element capacities, result in potential and phase differences between the actual internal elements and their external terminals, which increase roughly with the square of the frequency. There exist, in effect, standing waves on the leads and as the frequency increases, a condition is reached

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where the voltage nodes move inside the envelope, i.e., the effective lengths of the internal leads are greater than a quarter wavelength. In common high-power, water-cooled tubes, this condition occurs at frequencies from 30 to 50 megacycles. As a result, at ultra-high frequency it is impossible directly to ground the filaments for radio

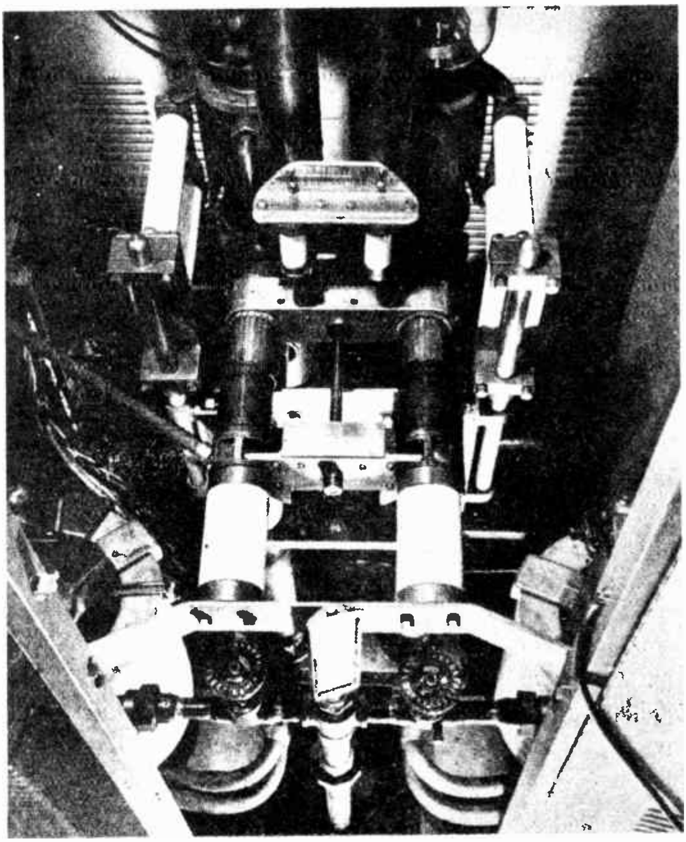

Fig. 1—Tank circuit of a fifty-megacycle television power amplifier. The inductive section is very short and tuned by an adjustable shorting bar. The two tubes dropping from the top carry the antenna coupling leads.

frequencies or achieve satisfactory neutralization, because of the length of the grid and filament leads.

In Figure 3-A are shown graphically the voltage gradients existing on the elements of Type 899 as observed in actual operation at 50 megacycles. Figure 3-B shows an approximate equivalent network representing the tube under these conditions. It will be noted that the external grid terminal is actually at a voltage nodal point.

In Figure 2 there are also shown two smaller water-cooled tubes which are used in ultra-high-frequency transmitters, Types 846 and 858. It will be noted that these tubes are of the single-ended type, that is, the filament and grid leads enter a common envelope, the opposite end of the anode being closed as contrasted with Type 899 which is double ended, that is, the grid and filament structures are supported by separate envelopes at opposite ends of the tube. The single-ended construction has generally been found easier to handle at ultra-high frequency as the excitation is normally introduced as a grid-filament potential. For this reason it is convenient to have the external terminals for these elements close together. Also in ultra-high frequency circuits it has been found convenient to form the tank-circuit inductance from straight tubings which are a continuance of the water-jacket assemblies. Examples of this type of tube mounting are shown in Figures 6 and 7. Type 846 tube, because of its small

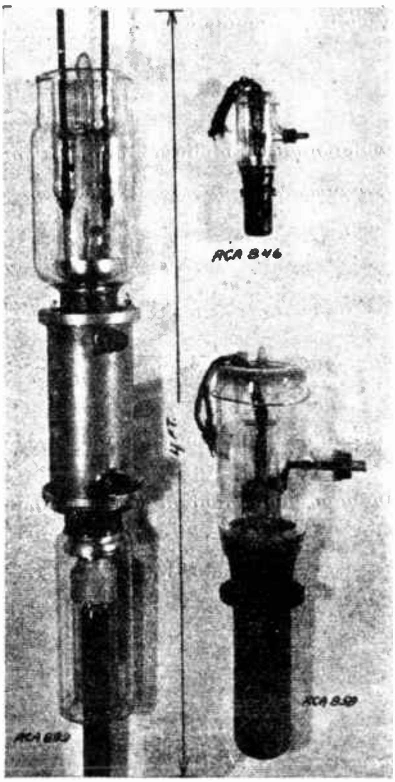

Fig. 2—Water-cooled vacuum tubes used in ultra-high-frequency television transmitters.

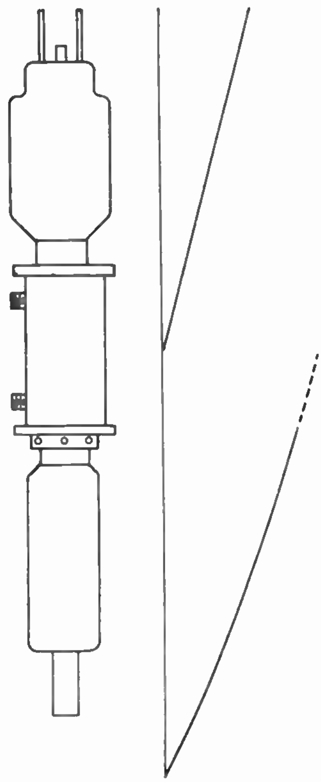

Fig. 3A—Voltage gradients in 899 Tube at 50 mc.

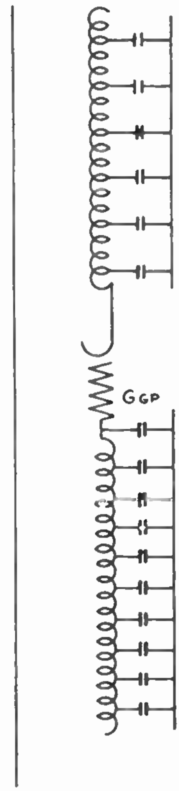

Fig. 3B—Approximate equivalent network.

physical size, functions satisfactorily in conventional circuits at frequencies as high as 100 megacycles. Type 858, which is considerably larger, has been found most useful for frequencies below 40 to 45 megacycles.

Reactance of Filament Leads

In operation the reactance of the filament leads is common to the plate and grid circuit, as shown in Figure 4-A, and in tubes of large physical size, the internal reactance is great enough to make satisfac-

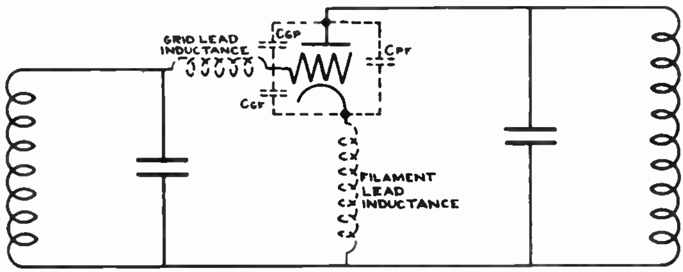

Fig. 4A—Circuit illustrating grid circuit-plate circuit coupling from filament-lead reactance.

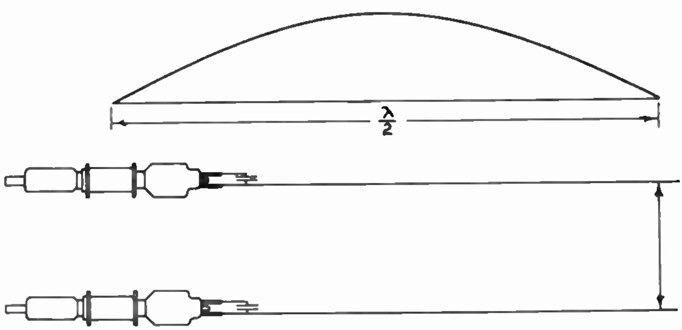

Fig. 4B—Showing how half-wavelength leads overcome internal filament-lead reactance.

tory neutralization difficult when the filaments are grounded directly. Even when satisfactory neutralization can be achieved, this filament-lead reactance prevents attainment of 100 per cent modulation of a modulated stage, as it permits radiation of the excitation power on negative-modulation peaks.

For several reasons to be discussed later, push-pull circuits are used almost entirely for large high-power u-h-f transmitter stages. This permits a simple method of overcoming the reactance of the filament leads by interconnecting the filaments of the opposing tubes through a pair of parallel conductors, as shown in Figure 4-B. These are inter-connected at a point effectively one-half wavelength from the actual cathodes, giving an effect substantially the same as a direct interconnection between filaments. In practice, the inter-connecting bar is made adjustable and the correct setting determined as a part of the neutralizing procedure. Figure 5 shows the installation of such filament lines. At 50 megacycles these filament lines are of the order of 10 feet in length and for convenience they are doubled back on themselves to reduce the size of the inclosure required.

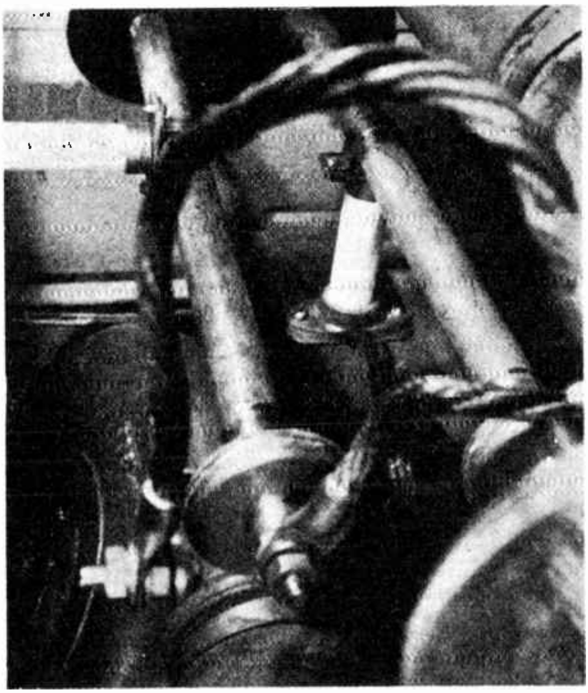

Fig. 5—Filament tuning lines used with Type 899 tubes on a 50-mc. transmitter. Note by-pass condensers which by-pass opposite side of filament to line. The heating current circuit is completed through an internal conductor.

Neutralization Problems

Long internal grid leads result in difficulties in cross-grid, cross-plate neutralizing, which may be further increased by necessarily long external neutralizing leads. In the case of Type 899 the internal grid lead is effectively a quarter wavelength at 50 megacycles and this makes neutralization difficult through connections to the external grid terminal. Fortunately, the grid end of this tube is of such construction that it was found feasible to form the cross-grid or cross-plate neutralizing capacity directly between the internal grid-lead and external concentric-sleeve fitting over the glass envelope. These neutralizing sleeves may be seen in Figure 8. Even with this arrangement it is not possible to form a true reactance bridge, because there is still left a considerable length of free grid reactance and the circuit is neutralized only over a small band near the operating frequency and has to be heavily loaded to prevent oscillation.

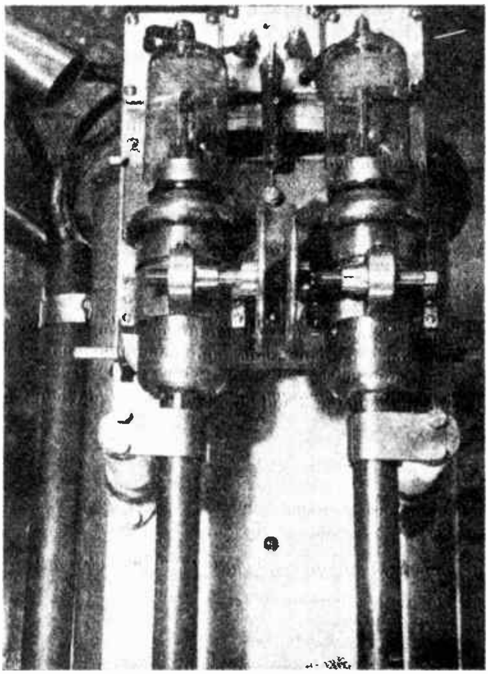

Fig. 6—A 50-megacycle amplifier. Tank circuit is of the parallel line type using a small disc condenser for fine adjustment. Compare the size of the tank conductors with the 90-mc. unit shown in Fig. 7.

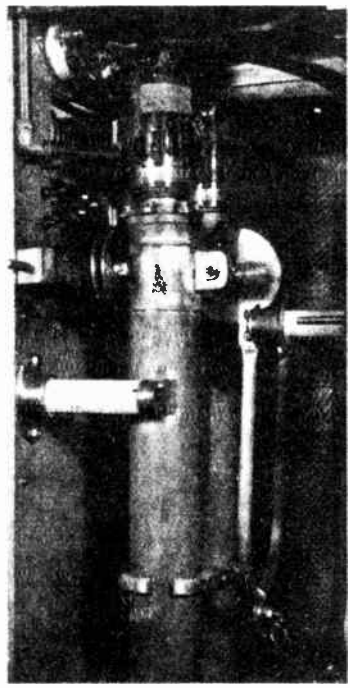

Fig. 7 — A five-kilowatt power-amplifier adjusted for operation at 90 megacycles. Note length of tank circuit as a result of using large diameter conductors.

INTER-ELECTRODE CAPACITIES

Inter-electrode tube capacities impose a number of u-h-f limitations on tube performance. In the case of high-power transmitting tubes, these limitations are of a different nature than those generally associated with low power and receiver applications. Large water-cooled tubes have high inter-electrode capacities, output capacities ranging from 25 to 50 $\mu\mu f$. These capacities do not impose serious tuning difficulties as the physical size of the tube makes it convenient to use large tank circuits having very low inductance. Figure 7 is a view of a tank circuit of the parallel-line type, using an adjustable shorting bar with a small vernier condenser to cover a tuning range from 40 to 90 megacycles. However, from the standpoint of neutralizing, high grid-plate capacities are awkward as the physical size of the neutralizing condensers necessitates long leads and increases stray capacity effects.

At ultra-high frequencies, the inter-electrode capacities have very low reactance values and, as a result, the circulating currents in the tubes become unusually high. These high currents cause excessive heating of the elements, leads, and seals and in general necessitate extra precautions in air cooling of all glass parts, particularly the seals. This heating is also increased at ultra-high frequencies by the increase of the radio-frequency resistance because of skin effect.

In television applications, inter-electrode capacities have a more serious effect particularly in r-f power amplifiers required to pass modulation side bands, which under present standards may be 2.5 megacycles from the carrier. It is not generally realized that for a desired tank-circuit frequency response, the tube, neutralizing and associated stray capacities, automatically determine the load resistance regardless of carrier frequency. In practical cases, this has generally necessitated operating tubes into load resistances considerably lower than normal with resulting poor plate efficiency.

In television power-amplifier applications, tube efficiency in one respect depends upon the ratio of tube capacity to plate conductance. Unfortunately, this ratio is a fundamental inherent relation in practical vacuum tubes of the triode type, and while tubes may be improved, it is doubtful if the present conception of a triode r-f power amplifier is the final answer for high-definition television applications.

The high side-band frequencies do not require 100 per cent modulation of the transmitter. It is thus possible to compensate partially for discrimination against high frequencies occurring in the radio-frequency circuits by equalizing at low levels in the video amplifier. However, excessive compensation in this type usually introduces objectionable phase shifts and transients. The problems of relaying a picture from the studio to the radiating transmitter and amplifying it to modulation power, is of itself a sufficient problem. It is therefore desirable that the radio-frequency circuits of the transmitter have a flat characteristic over the frequency band to be transmitted.

In one case of a 7.5-kilowatt, 50-megacycle television transmitter, in order to obtain a power-amplifier frequency response flat within 3 decibels over a 1.5-megacycle band, it was necessary to overload the power amplifier to a point where the plate efficiency was less than 15 per cent. The power amplifier used two Type 899 tubes in push-pull and the total plate input was approximately 60 kilowatts when delivering 7.5-kilowatts carrier-wave output.

New tube developments increasing the ratio of output conductance to output capacity may partially alleviate the poor power efficiency at

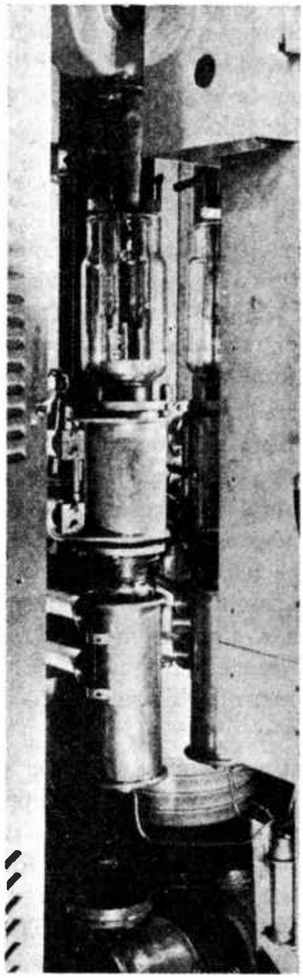

Fig. 8 — Showing neutralizing sleeves enclosing grid end of Type 899 tubes in a fifty-megacycle power amplifier. Note air blowers above and below tube to cool glass parts.

present obtainable in television transmitters. However, the difficulty is more or less fundamental with tubes of the triode type, and the ultimate solution will, more likely, be the development of entirely new types of power-amplifying tubes and modulation methods.

Because of the difficulties in producing high modulation power at the high side-band frequencies involved in television transmission, grid modulation in the power amplifier has been the most practical method of modulating high-power u-h-f television transmitters. Absorption modulation has been used successfully on low-power transmitters of one or two kilowatts carrier power. The principal advantage of absorption modulation as applied to television is that it removes the band-pass requirements from the power-amplifier circuits and consequently makes possible higher plate efficiency.

Circuits

At ultra-high frequencies, wavelengths reduce to a few feet and in high-power transmitters this fact introduces difficulties, but makes practicable circuits not adapted for lower-frequency design. In general, it becomes convenient to regard all circuit elements as sections of transmission lines and analyze them as such.

To begin with, at frequencies above 40 megacycles, it is found economical to use resonant line-controlled master oscillators as the primary frequency source. In such oscillators, the equivalent of a quarter-wavelength low-loss line resonator is used as the primary oscillatory circuit with power-oscillator tubes. Such resonators become of convenient size in the ultra-high frequency band and it has been found that the total transmitter tube complement is much less than would be required with a conventional frequency source such as a crystal oscillator and subsequent frequency multipliers and amplifiers.[1] Figure 10 shows a 50-mc. quarter-wave, line-controlled power oscillator.

Enclosures or mounting frames used for the high-power stages of u-h-f transmitters, because of their size approach major fractions of the operating wavelengths in dimensions. A true common r-f ground for the inclosed circuit is thus difficult to obtain, and considerable difficulty is experienced with single-tube circuits which necessarily are assymmetrical with respect to an enclosure. Troubles from this source largely disappear when push-pull circuits are used and mounted symmetrically in relation to a large plane-conducting surface. For these reasons, push-pull types of circuits are generally used in preference to single-tube circuits where the physical size is a major part of a wavelength.

At ultra-high frequencies quarter and half-wavelength line sections become reasonably short in length and it is practicable to take advan-

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[1] "Frequency Control by Low-Power-Factor Line Circuits" by P. S. Carter and C. W. Hansell. *Proc. I.R.E.*, April 1936.

tage of some of their particular properties. Thus in u-h-f transmitters quarter-wave line sections are used as impedance transformers, "metallic" insulators, and impedance inverters. In Figure 9 is shown an assemblage of quarter and one-half-wave, coaxial-line sections forming a cross-coupling filter to permit the operation of both the picture and sound transmitters into a common antenna without objectionable cross modulation. A U-shaped section of coaxial line serving as a transformer to couple the 72-ohm coaxial line to a 500-ohm, two-conductor, open-wire line is also shown. Short sections of lines having open or

Fig. 9—Quarter and half-wavelength line sections used to form high-"Q" filter elements preventing cross coupling of "sound" and "picture" transmitters separated a few percent in frequency and operated on a common antenna.

short-circuited terminations are conveniently used as efficient reactances at ultra-high frequency. Examples of this type of application of stub-line sections are the use of parallel tubular conductors having lengths less than a quarter wavelength to form the inductive component in high-power tank circuits. Several such assemblies are shown in the accompanying illustrations.

At ultra-high frequencies in circuits of large physical size all currents may be assumed to flow in the surfaces of the conductor, that is, constrained to a skin of less than a thousandth of an inch deep. This makes possible the construction of circuit members from inexpensive, easily fabricated materials such as steel which is subsequently plated with a highly conductive metal such as silver. A frequency-controlled resonator may be constructed entirely of cold-rolled steel and invar and silver plated. The actual conducting surface is thus formed of silver which has a very low electrical resistance, and at the same time, the structure is lighter and stronger and has a lower thermal

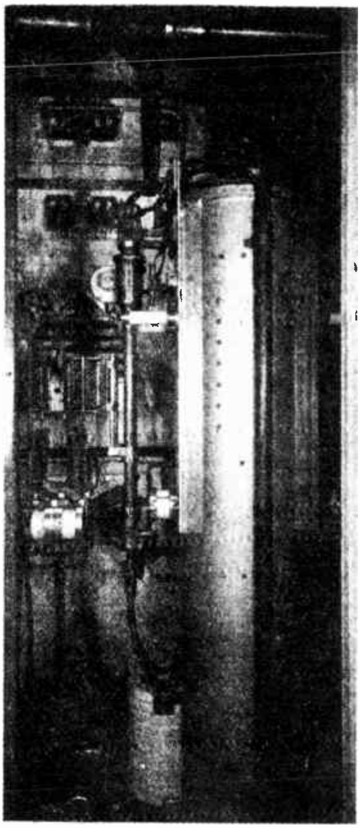

Fig. 10 — 50-megacycle "power" master oscillator. Frequency is stabilized by means of the quarter-wave coaxial line.

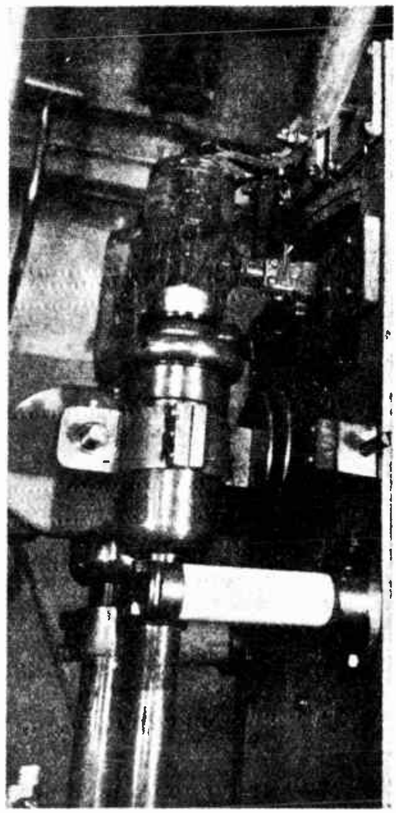

Fig. 11—A close-up of the oscillator shown in Fig. 10. Mechanical arrangement is simple and rugged and provides short electrical connections.

coefficient of expansion than copper, which formerly has been used for these devices. As the practical thickness of plating of this type is limited to a few thousandths of an inch, this type of construction could not be used at lower frequencies where the depth of penetration is greater. It is necessary to consider skin effect and current distribution in the design of u-h-f transmitter components, as these phenomena are of much more importance at these frequencies.

Auxiliary Apparatus

The difficulties encountered with tubes for u-h-f work have been previously discussed. Other apparatus such as condensers, resistors, meters, insulators, etc., also have serious limitations.

Condensers

Variable condensers of the conventional type cannot be used at ultra-high frequencies primarily because both minimum and ground capacity values are too high and insulation paths are not very long. For most u-h-f work two circular disks arranged so that the distance between them can be varied continuously have been found to be satisfactory and can be mounted directly on a tank circuit without requiring insulating mountings. In most u-h-f circuits the tube and neutralizing capacities form the major part of the tank capacity. External capacities are added only for tuning purposes. Suitable fixed condensers for by-passing, and coupling present serious difficulties. It is frequently desirable directly to couple the plate circuit of one stage to the grid circuit of the next. A coupling condenser is required to block the d-c plate voltage from the bias voltage of the next stage. In high-power u-h-f transmitters the radio frequency currents in this circuit may reach magnitudes of 30 to 40 amperes or more. At 50 megacycles, 1000 $\mu\mu f$. are required to obtain 3.2 ohms of reactance. A value as high as 15 to 20 ohms may be tolerated in coupling or by-passing, but a higher reactance will cause difficulties. Ultra-high-frequency circuits are usually constructed of low-reactance components, and higher-reactance blocking condensers will greatly disturb the circuit operation.

The condensers usually available for this service consist of a stack of copper sheets with mica insulation impregnated with wax. The dielectric losses in the wax and mica go up rapidly with frequency, resulting in excessive heating of the condenser at values much below its rated current. Another disadvantage with this type of construction is that it often results in having considerable inductance in series with the condenser proper. One alternative is to use high-current-rating condensers and operate them considerably below their rating. This is undesirable because of the bulkiness of the condensers, which is detrimental to good circuit design. Other dielectrics may have possibilities and a suitable condenser may be developed in the future.

Air has proven to be the most reliable dielectric, but has the disadvantage of having a dielectric constant of one, which results in bulky condensers for the conditions mentioned above; namely, 40 amperes r.f. at 50 megacycles, 10,000 volts d.c. and from 200 to 1000 $\mu\mu f$. Compressed air condensers may offer a solution to this problem since the spacing may be decreased approximately as the pressure is increased. However at ultra-high frequencies compressed air condensers present insulation difficulties that offset their advantages.

Vacuum condensers similar in construction to vacuum tubes have been tried in an effort to obtain high voltage rating in small physical space. These failed by going "gassy" as they do not have the "clean-up"

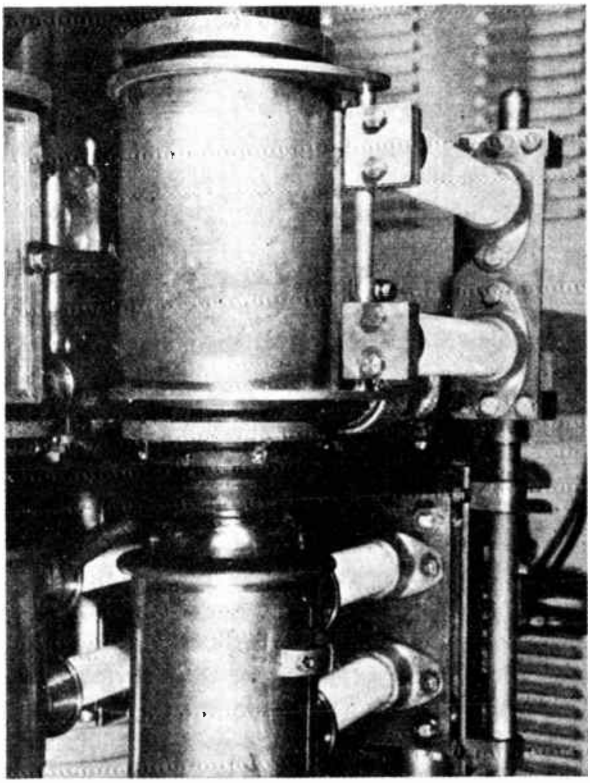

Fig. 12—A tube mounting including neutralizing sleeves and output coupling capacitors.

feature of vacuum tubes in operation. It is relatively easy to find standard condensers that will stand up for by-passing purposes, since for this condition the r-f current through the condenser is usually small. In many cases, however, the condenser will have considerable impedance to ground, because of its inductance. A case was encountered in which a parasitic oscillation existed with all types of standard condensers used for by-passing. A large parallel-plate condenser of extremely low inductance finally cured this condition.

Insulators

Closely associated with condenser problems is the problem of insulation. Any insulator is in a sense a capacity with the insulating material as the dielectric. For lower frequencies, the admittance of an insulator is so slight as to be negligible, but at u.h.f. there are many cases in which the radio-frequency currents flowing through the insulator are of such magnitude as to shatter the insulator, due to the internal heat produced. Points of contact with metal were found to be glowing at white heat. The above conditions as a rule are true only when a metal button or screw extends into the insulating material. This results in internal heating of the insulator, causing it to shatter. A simple remedy lies in the use of a corona shield. The corona shield tends to divert the path of the r-f currents along the outside of the insulator where cooling may take place.

No really suitable insulating material is available for u-h-f high-power transmitter work combining good insulating properties with mechanical strength. For this reason u-h-f transmitters must be constructed to eliminate insulation in high-frequency fields.

Meters

The measurement of u-h-f currents is a difficult problem. The ordinary calibration of thermocouple ammeters does not apply at u.h.f. because the skin effect in the couple causes the meter to read high. This, however, can be taken into account by applying a suitable correction factor.[3] A further difficulty, however, arises when the meter is actually placed in the circuit. In many cases the circuit is disturbed by the presence of the meter, resulting in erroneous readings. Lack of satisfactory voltage and current indicators increase the difficulties of studying problems in connection with u-h-f transmitters.

Resistors

Resistors are often desirable in u-h-f television circuits. It is difficult to build good non-inductive resistors at lower frequencies and at u.h.f. the problem is still more difficult. Types that are satisfactory at lower frequencies develop "hot spots" through the presence of standing waves. Carbon resistors become capacities at u.h.f. because of their granular structure. Metal coated resistors are satisfactory for low-power work, but no satisfactory resistors of this type have been

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[3] "Thermocouple Ammeters for Ultra-High Frequencies" by John H. Miller, *Proc. I.R.E.*, December 1936.

developed for high-power work. A pure resistance, free from reactance, is practically impossible to obtain at u.h.f. A possible exception to this statement may be an infinite line having no reflections.

Another method of obtaining a resistance free from reactance is to tune it out. For instance, load circuits have been constructed using a high-resistivity material as the inductance element of a tank circuit. This circuit may be tapped at any two symmetrical points and a pure resistance obtained, the value depending on the tapping points. This method may be used as an artificial load by circulating water through the conductor and measuring the temperature rise and water flow. It has been found desirable to arrange such loading circuits to avoid all coupling with associated circuits since the energy stored is extremely high and its field may interfere with the function of other circuits.

It has been the purpose of this article to give a general description of the problems encountered in the development of high-power television transmitters, and some of the methods used to overcome them. New vacuum tubes and equipment are now being developed with features intended to simplify these problems.

TELEVISION RADIO RELAY

BY

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Summary.—A general description of the 177 Mc television radio link between the RCA Building and the Empire State Building in New York City is given. The transmitter and receiver are described in detail along with results of tests on the circuit.

WITH the installation of the new television transmitter in the Empire State Building, it became necessary to provide a connecting link to carry the video frequencies from the studios at Radio City to the transmitter. Both a coaxial cable and radio circuit are used for this link. This paper is devoted entirely to the radio circuit and its terminals.

The radio circuit is operating on a carrier frequency of 177 Mc which was chosen to be clear from harmonics of the picture and sound transmitters operating in close proximity to the relay receiver. A high frequency was chosen to be free from interference on existing radio services, to allow directive antennas to be used in which space was a limiting factor, and to take advantage of the lower man-made noise level encountered from sources such as elevator contractors, motors, etc. Vacuum tubes now available make operation above 200 Mc difficult. The air line distance from the transmitting antenna at Radio City to the receiving antenna at the Empire State Building is approximately 4600 feet. Ultra-high frequencies are particularly adaptable to distances of this sort, and to the wide modulation band required.

PROPAGATION TESTS

The video frequencies up to 1500 kc. to be transmitted require the radio circuit to carry a band of 3000 kc. with double side band transmission. Calculation showed that the combination of the direct and reflected rays at the receiving antenna could cause serious variations in transmission efficiency throughout the extremely wide band, depending upon the location of the points

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of reflection, and the intensity of the reflected ray or rays. To obtain more accurate information regarding this variation in transmission efficiency, propagation tests were carried out over the hand of 176 to 182 Mc. This work is described in a paper by P. S. Carter and G. S. Wickizer.[1] The results of these tests showed that a reasonably flat response could be obtained by using transmitting and receiving antennas having moderate horizontal directivity. Fig. 1, from the above paper, shows the response curve obtained with a directive transmitting antenna located at the 14th floor level of the RCA Building and a directive receiving

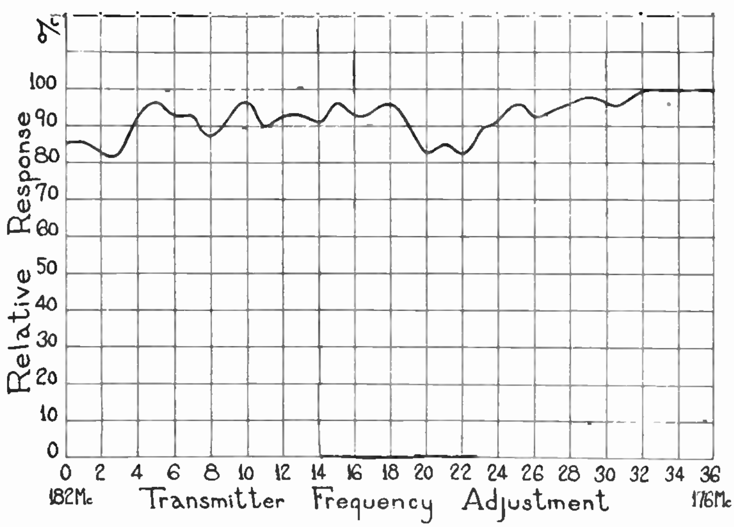

Fig. 1—Variation of received signal—horizontal one wavelength transmitting antenna at 14th floor level; receiving antenna, two horizontal half wave dipoles end to end spaced 1½ wavelengths between centers.

antenna at the 85th floor of the Empire State Building. The antennas now in use at each end of the circuit are electrically equivalent to each other and consist of a one wavelength horizontal radiator, fed at the middle, located in front of a metal reflector. Fig. 2 is a photograph of the transmitting antenna now in use. These antennas are sufficiently broad to pass, without appreciable attenuation, the 3000 kc. band.

Transmitter

The complete transmitter is mounted in a standard relay rack as shown in Fig. 3. The top unit contains the power amplifier, master oscillator, modulator and modulator amplifiers. This unit is mounted on rubber to protect the tubes and circuits from vibration. A peak voltmeter has been provided to measure the input level to the modulator amplifier and the output of the monitor rectifier. It is located just below the rubber mounted unit. Below the peak voltmeter panel is the d-c filament supply unit for the master oscillator. The bottom two units are plate supply rectifiers for the radio frequency stages and the video frequency stages. A schematic diagram for the transmitter is shown in Fig. 4.

The master oscillator, right hand compartment of Fig. 5, consists of two RCA-834 type tubes operating in push-pull at a

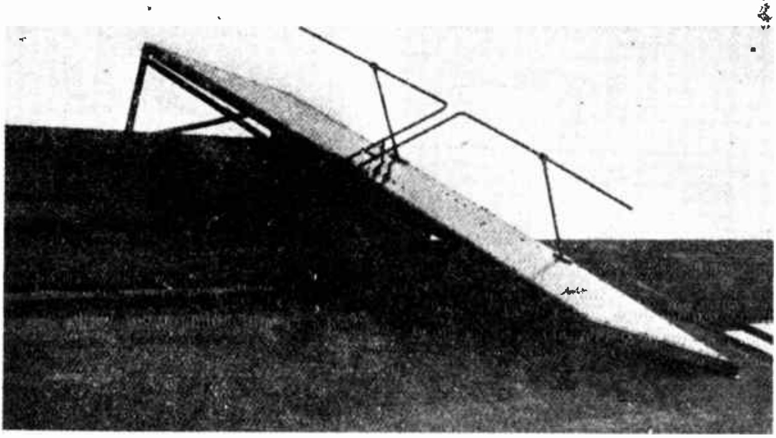

Fig. 2—Transmitting antenna at 14th floor level of RCA Building.

frequency of 177 Mc. The frequency of the oscillator is determined by a low power factor, concentric resonator to which the grids are inductively coupled. The grid loops which are coupled to the frequency control circuit are in opposite polarity so that the phase of the grid voltages differ by 180°.[2] The ratio of the diameters of the concentric conductors of the frequency controlling circuit is 3.5.[3] A theoretical Q of 11,370 is obtained with an inside conductor diameter of 2.25 inches.[2, 4] The inner member, which is .2 of a wavelength long, has one end silver soldered to an end plate of the outer sheath, and the other end connected to a four-inch diameter sylphon bellows one inch long. The free end of this bellows is screwed to an invar rod which is connected to the same end plate which supports the inner conductor. Since the temperature coefficient of expansion of invar is nearly zero the electrical length of the inner conductor is approximately constant with changes in temperature. Thus the resonant frequency

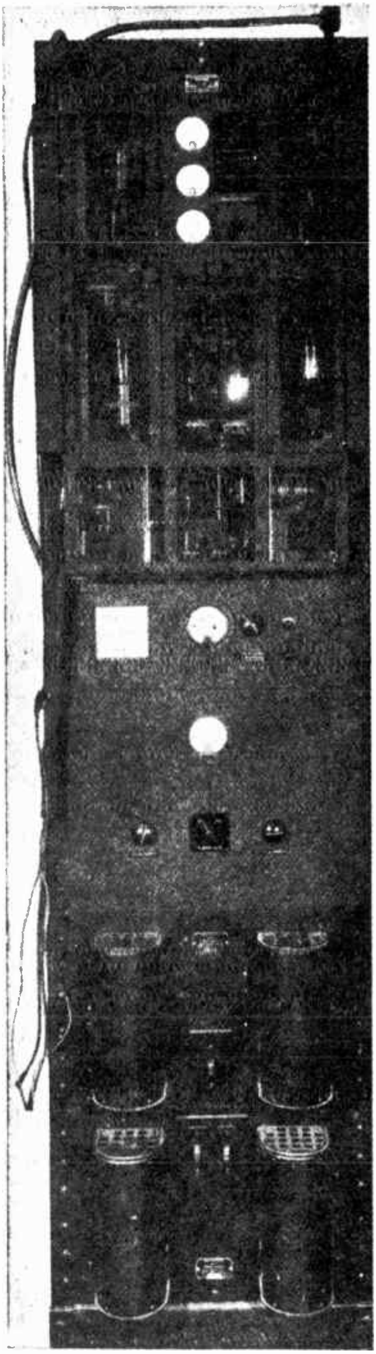

Fig. 3—177 Mc television radio relay transmitter.

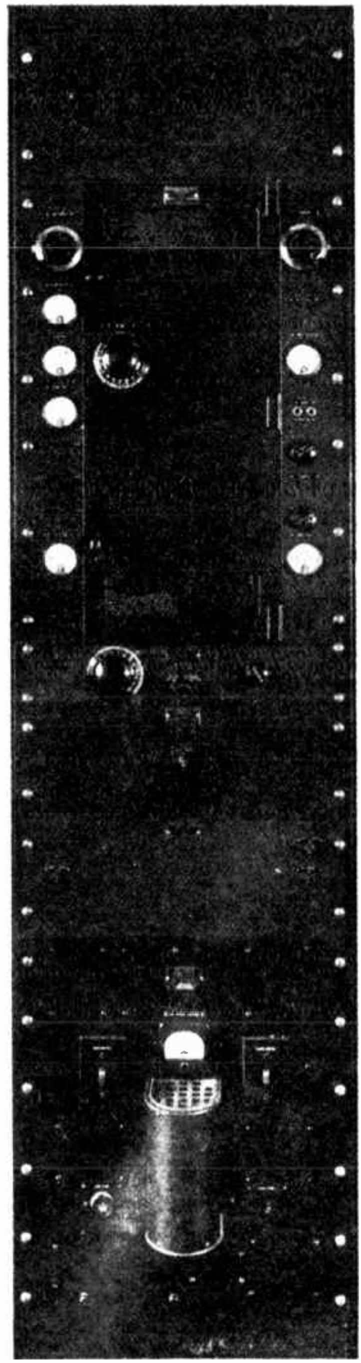

Fig. 7—177 Mc television radio relay receiver.

of the low power factor circuit is made substantially independent of temperature.

The master oscillator has adjustable impedances in its plate and filament circuits. The grid circuit reactance was adjusted to about the required value by a short wire connected from grid to grid. This wire is in parallel with the grid loops which couple to the low power factor circuit. The plate inductance is a concentric conductor line connected from plate to plate. At the neutral point on this line the inside conductor is exposed so the power amplifier grid coil may be inductively coupled to it. The photograph of Fig. 5 shows the master oscillator on the right

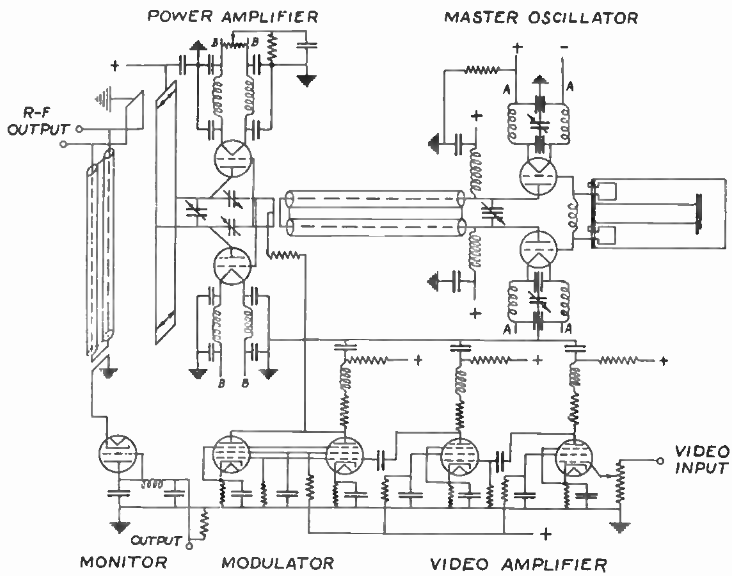

Fig. 4—Schematic diagram of television relay transmitter.

hand side, the modulator in the center, and the power amplifier on the left. With such an arrangement the connection from the modulator to the power amplifier grid circuit is short and the modulator output capacitance is reduced. However, this makes the link from the master oscillator to the power amplifier rather long. This link is the master oscillator plate inductance and its maximum reactance is fixed by the tube inter-electrode capacities. The correct inductance was obtained by the proper choice of the conductor diameters. A small balanced condenser connected from plate to plate is used for fine adjustment of the master oscillator circuit. Plate voltage is supplied to each tube through an r-f choke.

The r-f output stage is a conventional, push-pull, cross neutralized amplifier.[5] The tubes (two RCA-834's) are located as shown in Fig. 5 to make the length of the connections from the grid and plate tube prongs to the neutralizing condensers a minimum. This is necessary to prevent parasitic oscillations. The

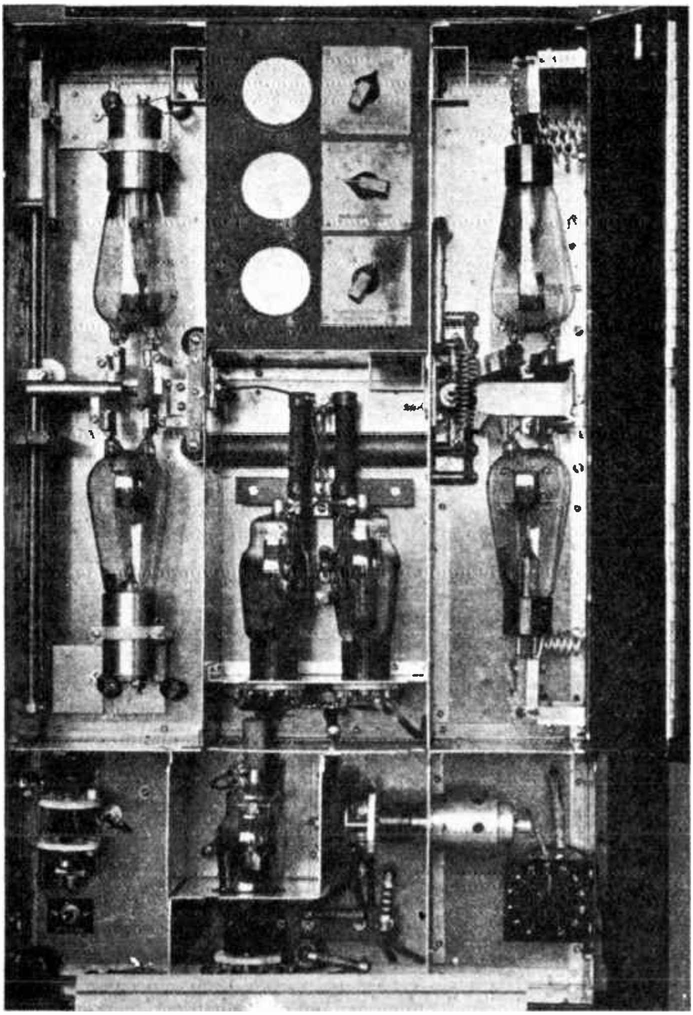

Fig. 5—Radio and video-frequency units of transmitter; left, power amplifier; right, master oscillator; center, modulator; lower left, monitor; lower right, modulator amplifiers.

neutralizing condensers are the horizontal concentric cylinders at the center of the power amplifier compartment. The outside cylinders are connected to the plates and are made up of two telescoping tubes for adjusting the neutralizing capacitance. The inside cylinders are connected to the grids. This arrangement reduces the stray capacity between the input and output circuits of the power amplifier.

The power amplifier grid circuit is an untuned inductance composed of a short brass strip connected from grid to grid, and closely coupled to the voltage nodal point of the master oscillator plate inductance. The center point of the grid inductance is directly connected to the modulator plates. By eliminating the blocking condenser between the modulator and the power amplifier the stray capacitance of the modulator output circuit to

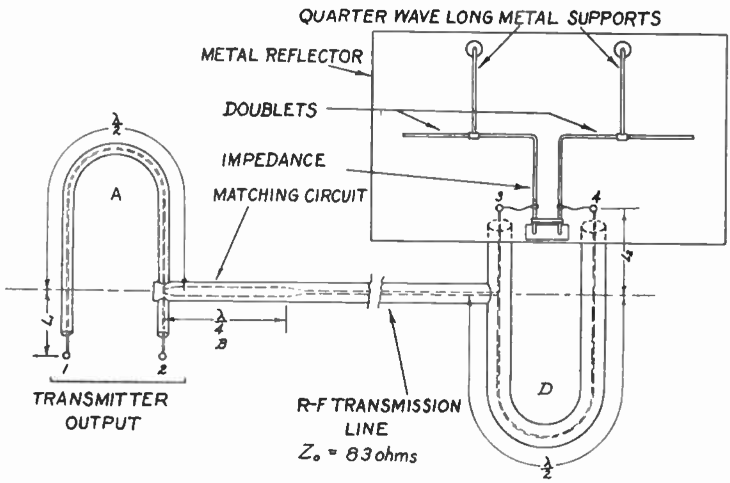

Fig. 6—Balanced to unbalanced transfer circuits, transmission line, and transmitting antenna.

ground is reduced. Since the power amplifier grids are connected directly to the modulator plates they are maintained at a plus potential of 250 volts. To give proper operating bias the filaments are maintained at a plus potential of 400 volts by a filament return resistor.

To maintain the symmetry of the power amplifier output circuit the plate inductance is made of two balanced lines in parallel. Slides on these two wire lines are provided for approximate tuning of the power amplifier plate circuit, and a small two-plate variable condenser for the fine adjustment. One branch of the plate circuit is inductively coupled to a balanced 150-ohm load. The power amplifier will deliver to this load a 15-watt carrier.

The monitor step-down transformer is a concentric conductor line one half wavelength long connected across the output terminals of the power amplifier. At the center or voltage nodal point on this line a loop is inductively coupled to the inside conductor. This loop is connected to a rectifier which is used to monitor the transmitter r-f output.

The modulator uses two RCA-802's in parallel. The frequency response curve for each video frequency stage was made flat by using the system described by Messrs. Kell, Bedford and Trainer.[6] If the total output capacitive reactance is X_c at the highest frequency (1.5 megacycles) it is desired to transmit, then the plate load impedance at this frequency is made $1.13X + j\ .5X_c$ ohms. The phase shift produced by the resultant plate impedance is proportional to frequency over the transmission band and, hence, does not produce phase distortion. The attenuation and phase shift produced at the low frequencies by the interstage coupling and cathode by-pass condensers are compensated for by choosing a suitable value of plate supply by-pass capacitance. The plate supply of each stage has a series resistance or damped reactor to isolate it from the power supply.

The modulator amplifier consists of two stages, an RCA-802 and RCA-6C6. An input level of .45 volts r.m.s. is required to modulate the transmitter 85 per cent.

The possibility of transferring the balanced output at the transmitter to an unbalanced load suitable for feeding a single coaxial line was considered. The circuit shown in Fig. 6 proved to be satisfactory. A coaxial line A with a characteristic impedance of 75 ohms was made $2L_1 + \lambda/2$ meters long. L_1 may be any convenient length. When there is a traveling wave on A the proper impedance of 150 ohms will be presented to the transmitter output terminals 1 and 2. The voltages on terminals 1 and 2 differ in phase by 180 degrees. The path from terminal 1 to the quarter wavelength section of line B is one half wavelength longer than the path from terminal 2 to this junction. Hence, the wave that leaves terminal 1 arrives at B in phase with the wave from terminal 2. The section of line B which has a characteristic impedance of 55.8 ohms steps the 83-ohm transmission line down to 37.5 ohms, which will match the two sections of the loop A.[7] The r-f transmission line is made of a one-inch inside diameter copper pipe and a quarter-inch outside diameter copper tube. The quarter-inch tube is held concentric with the one-inch pipe with low-loss insulators. These insulators are spaced about every quarter wave to reduce the reflections produced by them. The length of the line is 100 feet and the efficiency about 90 per cent. At the antenna the unbalanced feed is transformed to a balanced feed of 332 ohms. The wave that leaves the transmission line and travels over the longer branch of D arrives at terminal 4, 180 degrees out of phase with the wave which travels over the shorter branch of D to Terminal 3. The concentric conductor D has a characteristic impedance of 166 ohms. The two branches in parallel will match the transmission line if the Terminals 3 and 4 are connected to a load of 332 ohms.

The impedance matching circuit shown is adjusted to step the antenna input load down to this value at 177 Mc. This is necessary to give maximum efficiency and flat frequency response over the band used.

Receiver

The receiving antenna on the north wall of the 85th floor of the Empire State Building is approximately 100 feet from the receiver location. The antenna feed line is composed of two 76-ohm, 13-gauge, coaxial cables, located in a conduit running from the receiver rack to the back of the antenna reflector. The use of two cables gives in effect a balanced, shielded, 152-ohm transmission line. Special tests were carried out to properly match the antenna to the feed line at 177 Mc to get the greatest overall efficiency and flatest response with frequency. The transmission line loss was estimated to be not more than 1.9 db. With all adjustments made the cables were sealed off, evacuated, and filled with dry nitrogen under pressure. This process insures the removal of moisture from the cables, and gauges permanently installed show whether the pressure is maintained. The sealed cables are thus impervious to weather conditions.

Fig. 7 on Page 154 is a photograph of the front of the receiving rack as installed in the Empire State Building.

Fig. 8 shows a schematic diagram of the receiver circuits. A balanced concentric line type band-pass transformer, receives 177 Mc energy from the balanced coaxial feed line. The transformer in turn feeds a balanced heterodyne detector consisting of two RCA 954 acorn tubes whose cathodes are excited by a concentric line type of local oscillator operating at 156 Mc. The intermediate frequency of 21 Mc appears push-pull in the output of the balanced detector stage and is coupled to a single ended 6-stage, band-pass amplifier using coupled circuit transformers. The overall flat band width is A Mc. The i-f amplifier is fed to a linear diode rectifier (RCA 955), which in turn feeds the RCA

42 output tube. Video frequencies are carried from the receiver over a coaxial cable to the transmitter line amplifier.

Automatic gain control of the i-f amplifier is accomplished by means of a d-c amplifier driven from a voltage divider across the diode load resistance. This circuit is arranged to feed variable negative control voltages to two tubes in the i-f amplifier. A switch is provided to allow the gain to be controlled manually.

A set of switches on the front panel allows the plate currents of the i-f amplifier tubes to be checked on one meter without interrupting the operation of the receiver. The other plate currents with the exception of the automatic gain control tube are shown continuously on individual meters.

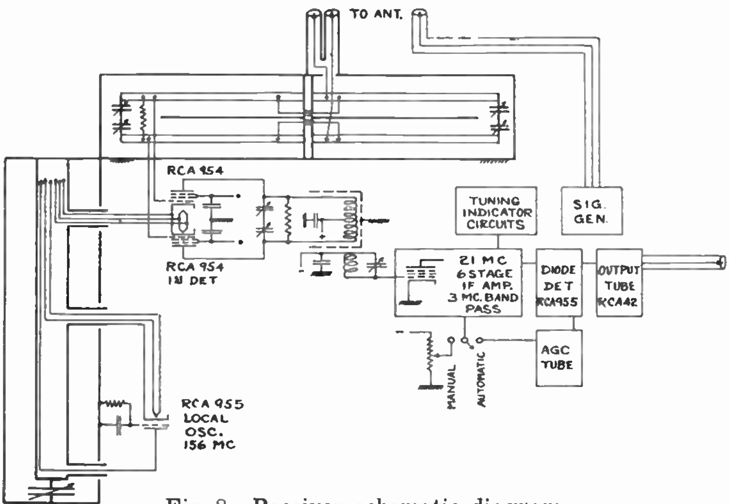

Fig. 8—Receiver schematic diagram.

Due to the extreme width of the i-f amplifier it is necessary to provide an indicator to show when the signal carrier is tuned to mid-band. This indicator allows the operator to easily find the correct setting of the local oscillator. A 0-1.5 ma. meter on the front panel shows the plate current of a biased triode detector, which is excited by a high C resonant circuit having fairly high Q. This resonant circuit is driven by an r-f pentode fed from the i-f amplifier. A push button is arranged to connect a small fixed capacity across part of the resonating inductor, and the resonant frequencies with the push button out and in are set to be equally spaced about the i-f mid-band. With such an arrangement the tuning indicator meter (0-1.5 ma) in the plate circuit of the biased detector will show no change with the push button out or in when the carrier is accurately tuned to mid-band.

A separate regulated power supply having an effective internal resistance of less than one ohm is used to supply power to the RCA 42 output tube. Thus, objectionable low frequency resonance often occurring in ordinary power filters is eliminated which permits a flat frequency response to be obtained down to 10 cycles or less with the output tube working into a load resistance of only 100 ohms.

In a receiver of this sort having a tuned band-pass input transformer at signal frequency, the problem of properly tuning these two circuits presents itself. The correct tuning to give a flat band pass is not necessarily that obtained by setting each dial for maximum response in the usual way. This problem is overcome by supplying in the receiver rack a shielded oscillator to supply energy over a single coaxial cable to a point on the

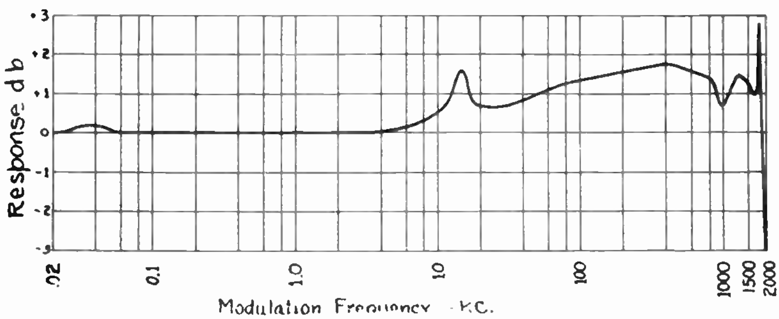

Fig. 9—Overall frequency characteristic of the radio relay.

antenna reflector. The cable termination near the antenna includes a damping resistance, a small radiating rod, and a shunt inductance all combining to produce a correct terminating impedance over the band of frequencies used. With this arrangement the operator can vary the oscillator frequency over the receiver pass band and observe the shape of the (antenna—feed line—receiver) characteristic on the output of the receiver. With a few trials the correct tuning of the input transformer can be obtained. These adjustments may be made when the 177 Mc television transmitter is off the air. These circuits once set require little attention thereafter.

Results of Tests

As might be expected a signal of considerable intensity is received at the Empire State Building. Calculation shows that the transmitter power and antenna directivity used should result in a direct ray field strength of approximately 30 millivolts per meter at the receiving antenna. A strong signal is necessary to override local disturbing noises from elevator machinery, etc., which become more bothersome the wider the receiver band width. It might be mentioned that lightning flashes in the immediate vicinity give only moderate clicks in the receiver output.

Signal strength observations made so far show variations in intensity of only a few per cent, showing that variations in rays reflected from the ground are unimportant.

Overall frequency characteristic measurements have been made from the transmitter input to the receiver output over a range of from 20 cycles to 2000 kc. Fig. 9 shows the result of these tests. The irregularities in the curve from 100 to 1700 kc. are mostly caused by the propagation path as mentioned earlier. The peak at 1800 kc. is produced by an equalizing circuit in the receiver output. It will be observed that the maximum deviation over the desired band of 20 cycles to 1500 kc., is 1.8 db which occurs at 400 kc.

Signal to noise level measurements were made at the receiver using a measuring amplifier having an effective band width of 20 kc. These measurements showed a signal to noise level of 44 db with 85 per cent. modulation at the transmitter. The noise level was almost entirely 60 and 120-cycle power supply hum.

Although the receiver is located near the high power television broadcast transmitters no trouble is experienced from interference. Pictures have been transmitted over this circuit without altering their quality.

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ANALYSIS AND DESIGN OF VIDEO AMPLIFIERS

BY

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NATURE OF THE PROBLEM

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

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

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

The necessity for constant time delay over the video band may be explained from consideration of the effect upon picture detail of using a video amplifier with phase characteristics which cause the high-frequency end of the video band to be delayed with respect to the low-frequency end in transmission through the amplifier. (This is generally the manner in which the time delay

* Trade Mark Registered U. S. Patent Office.

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

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

Generally the patterns scanned by the "Iconoscope"* beam are not as geometrically precise as that used here for discussion of the video amplifier requirements, but are made up of random

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* Trade Mark Registered U. S. Patent Office.

variations of light and dark shading. The necessity for constant time delay is no less important in this case, for the picture will be distorted in the event of non-uniform time delay, especially if the pattern contains considerable detail.

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

Analysis

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

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

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

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

R and L for a given value of f_o are determined by the load-circuit capacitance C. To fulfill this condition R must be made equal to the reactance of this load capacitance at the top frequency, f_o, that is, $R = \frac{1}{2\pi f_o C,}$, and the reactance of the compensating choke at f_o must be equal to half the load resistance,

i.e. $2\pi f_o L = \dfrac{R}{2}$

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

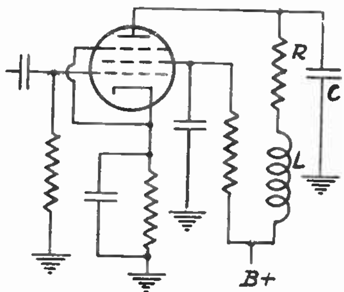

Fig. 1

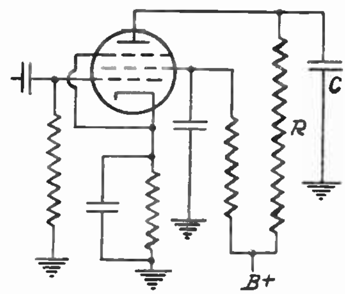

Fig. 2

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

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

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

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

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

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

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

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

Phase and Time Delay

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

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

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

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

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

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

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

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

The phase delay for a single compensated stage is

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

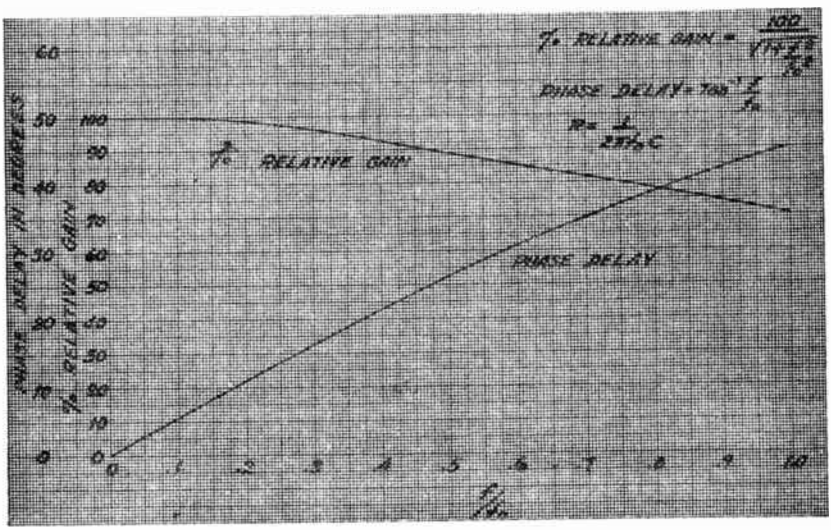

Fig. 3—Uncompensated stage of video amplifier.

and the time delay in seconds is

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

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

The phase delay of a compensated stage is plotted vs. f/f_o

in Figure 4 and it is seen that the non-linear phase characteristic will result in a non-uniform time-delay curve.

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

As a quantitative indication of the magnitude of anticipated time delay and its effect upon the displacement of picture ele-

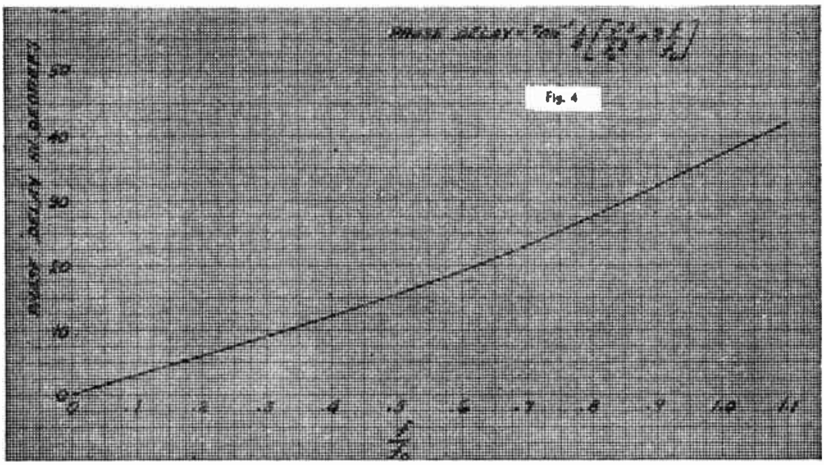

Fig. 4—One compensated stage of video amplifier.

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

Figure 5 shows that the delay increases with frequency. The total time delay is not significant, as it is the difference in time delays for the various frequencies in the video band with which we are concerned. These differences are the cause of the relative displacements of the various frequency components in the picture. Constant time delay would result in all the picture elements being displaced by the same amount, regardless of the video frequency with which they are associated, and no picture distortion would result. As it is, the curve of displacement vs. frequency shows that picture elements corresponding to the two extremes of the video band (60 cycles and 2.5 megacycles) will be displaced by approximately .019″ at the low end and .024″

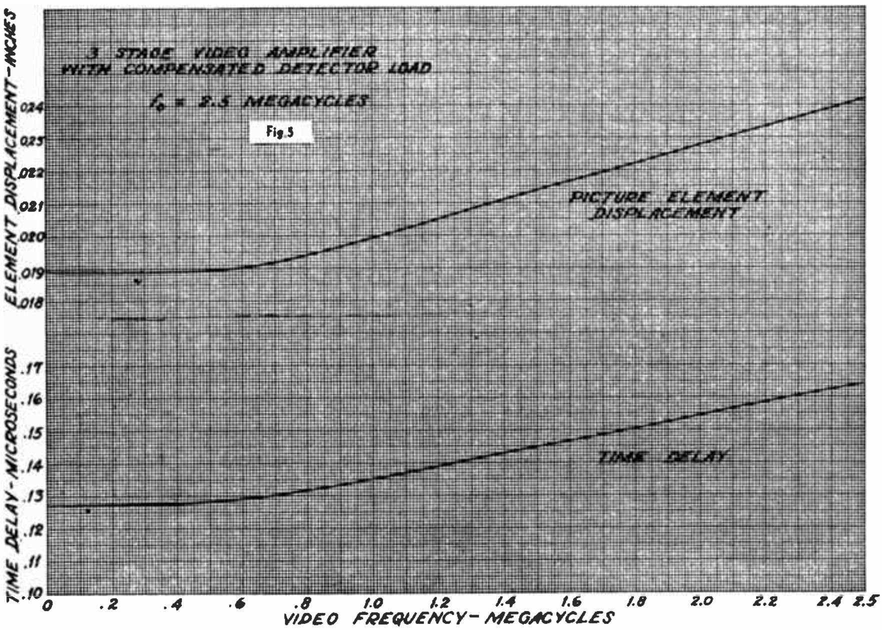

Fig. 5—Three-stage video amplifier with compensated detector load.

at the top end. The relative displacement, or the effect which causes the relative shift in picture elements and the corresponding distortion, is only .005″, and this is small in relation to the width of a scanning line.

Number of Stages

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

Low Frequency Considerations

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

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

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

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

Measurement of Gain and Phase Delay

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

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

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

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

The presence of capacitive reactance in the plate circuits of the video amplifier causes the output voltage to lag the input voltage in time phase; hence the *R-C* circuits used for phase shifting in the phase-angle measurements must be so arranged as to cause the phase of the input voltage to be delayed before

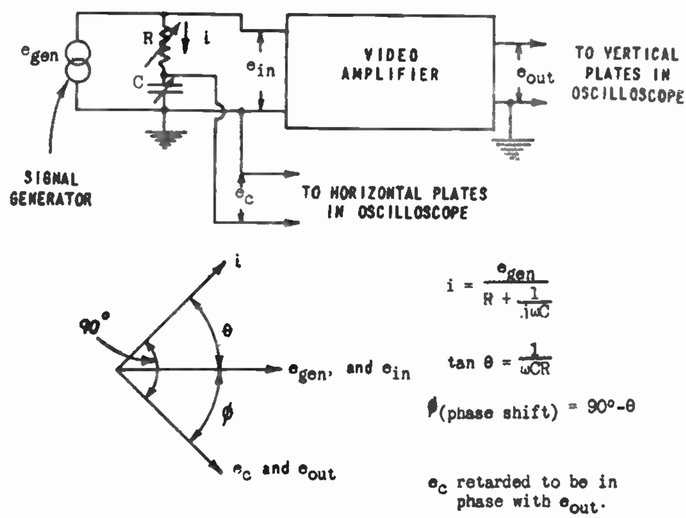

Fig. 6

it is applied to the oscilloscope, or, conversely to advance the phase of the output voltage. Figures 6 and 7 show typical circuits for use in this work. A linear trace is obtained when the phase shift in the *R-C* network is equal to the phase shift in the video amplifier. As noted above $\phi = (90° - \theta)$, where θ is the angle between e_{gen} and i through *R* and *C*. Note that *C* must include the additional capacitance occasioned by connection to the oscilloscope; i.e., either the input capacitance of the horizontal amplifier or the capacitance between horizontal deflecting plates, depending upon whether or not the voltage is applied directly to the plates. The input capacitance of the vertical-deflection system will also add to the output capacitance of the video amplifier, and this must be taken into account in determining overall performance.

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

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

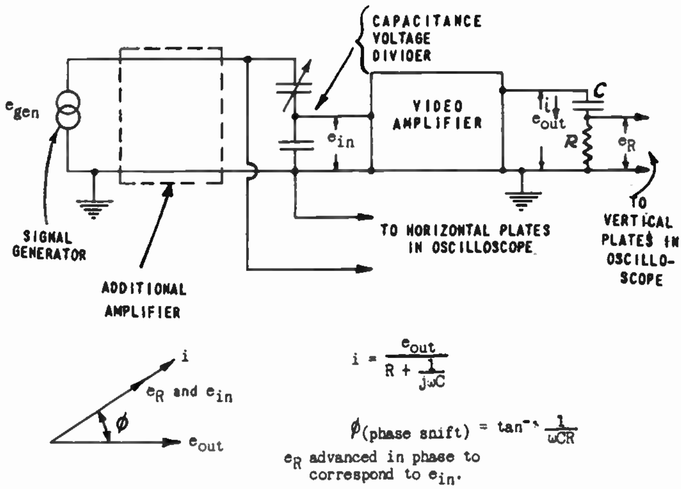

Fig. 7

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

APPENDIX I.

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

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

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

Substituting $R = 2L\omega_0 = \frac{1}{C\omega_0}$ as the condition for constant gain to $\frac{\omega_0}{2\pi}$ cycles.

$$Z_L = \frac{R\left[1 - \frac{j}{4}\left(\frac{f^3}{f_0^3} + 2\frac{f}{f_0}\right)\right]}{\left(\frac{f}{f_0}\right)^2 + \left(\frac{f^2}{2f_0^2} - 1\right)^2} \quad \text{gain} = \frac{g_m R \sqrt{1 + \left(\frac{1}{4}\frac{f^3}{f_0^3} + \frac{1}{2}\frac{f}{f_0}\right)^2}}{\left(\frac{f}{f_0}\right)^2 + \left(\frac{f^2}{2f_0^2} - 1\right)^2}$$

The phase delay in the stage is

$$\phi = +\tan^{-1}\tfrac{1}{4}\left(\frac{f^3}{f_0^3} + 2\frac{f}{f_0}\right)$$

and the time delay

$$\Delta t = \frac{\phi}{2\pi f} = +\frac{1}{2\pi f}\tan^{-1}\tfrac{1}{4}\left(\frac{f^3}{f_0^3} + 2\frac{f}{f_0}\right)$$

Appendix II.

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

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

$$Z_L = \frac{\frac{R}{S\omega C}}{R + \frac{1}{S\omega C}} = \frac{R\,(1 - SRC\omega)}{R^2C^2\omega^2 + 1} = \frac{R}{\sqrt{R^2C^2\omega^2 + 1}}$$

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

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

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

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

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

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

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

THE BRIGHTNESS OF OUTDOOR SCENES AND ITS RELATION TO TELEVISION TRANSMISSION

By

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Summary—*The average brightness of typical outdoor scenes has been determined by computation and by measurement. The average brightness of some scenes was found to be over 1000 candles per square foot, and of other scenes nearly zero. In many cases the average brightness lay between twenty and 200 candles per square foot. The sensitivity of a present-day television system using the Iconoscope has been found to be sufficient to permit the transmission of pictures with good quality when the average brightness of an average scene was greater than about fifteen candles per square foot. This sensitivity is sufficient for the transmission of parades, races, baseball games, and many other outdoor events. Football games, which last until near sunset, cannot always be satisfactory reproduced.*

Some of the Iconoscopes used in these tests are of added sensitivity, which has been achieved by means of a silver evaporation process, as well as by careful control of the purity of the materials.

During the early stages of television, both the transmitting and receiving systems were crude, and experimenters were glad to obtain a recognizable picture. The last few years have witnessed great improvement in the quality of the picture. The adoption of cathode-ray tubes has permitted a large increase in the number of scanning lines, and use of interlaced scanning and a greater number of frames per second has practically eliminated flicker. As the system improved, larger and brighter pictures became possible.

A comparable change has also taken place in the sensitivity of devices for converting light into television picture signals. The earliest apparatus required so much light that transmission was largely limited to films. At best, direct pickup could be obtained only when the scenes were in direct sunlight or under blinding artificial light. With the advent of such electronic devices as the Iconoscope,[1,2] direct transmission of outdoor scenes became practicable even on cloudy days.

Since the illumination requirements for television transmission now fall within practical limits, let us consider the relation between light available under average conditions and the sensitivity of the ap-

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[1] Registered trade-mark, RCA Manufacturing Company, Inc.

[2] V. K. Zworykin, "The Iconoscope—A modern version of the electric eye," Proc. I.R.E., vol. 22, pp. 16–32; January, (1934).

Reprinted from *Proc. I.R.E.*, August, 1937.

paratus. In the transmission of motion pictures or studio scenes, the amount of light used may be controlled; out-of-doors it is usually necessary to operate with whatever light the sun provides. In considering how much light is available for the illumination of a scene which is to be transmitted, we shall, therefore, largely limit the discussion to outdoor scenes in daylight.

The matter will be discussed from two angles: (1) the surface brightness and contrast of typical outdoor scenes, and (2) the brightness and contrast which are necessary for the transmission of a satisfactory television picture. Because of the complexity of the subject it will be necessary to make some approximations, but these approximations are relatively unimportant in view of the wide range of brightness encountered and the tolerance of the television apparatus.

Illumination of Outdoor Scenes

When commercial television broadcasting is well established, it is probable that the public will wish to see varied events, such as basegall games, boat races, parades, and political gatherings. The illumination encountered will vary over a tremendous range at different pickup points, and even change in a short period of time at a given place. A successful broadcast must make allowance for the variations that may occur, for, unlike motion pictures, the unsatisfactory scenes cannot be discarded. Neither would it be satisfactory to postpone or stop a broadcast because of adverse conditions.

It is desirable, therefore, to know how much light to expect from a given scene. From data already published the required information may be judged for a general case; this we have supplemented with data obtained by measurement of some specific subjects.

Since the intensity of the-light which the lens focuses on the apparatus depends on the surface brightness of the object, irrespective of the object distance, surface brightness (usually measured in candles per square foot)[3] is the best measure of the light available for the transmission of a scene.[4] This quantity can be directly measured for each scene that is to be transmitted, or a fair estimate may be made by a simple computation. If the intensity of illumination received from the sun and the sky under different conditions is known, the surface brightness of different diffusely reflecting objects can be calculated by the use of the relationship

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[3] Care should be taken to distinguish foot-candles from candles per square foot. The foot-candle is a measure of the illumination received at any given point from a source of light, while candles per square foot is a measure of the intrinsic brightness of a source of light.

[4] W. N. Goodwin, Jr., "The Photronic photographic exposure meter," *Jour. S.M.P.E.*, vol. 20, pp. 95–118; February, (1933).

$$B = RI/\pi \qquad (1)$$

where,

B is the surface brightness of the object in candles square feet
R is the reflection coefficient of the object
I is the illumination received by the object in foot-candles.

The illumination at any place and time can be computed from such factors as the sun's distance, its radiation, and the absorption and scattering of the atmosphere. However, since experimental data have already been taken by several observers, it is more convenient to use

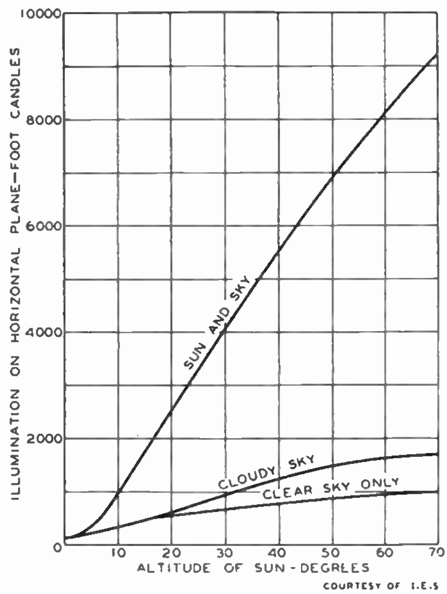

Fig. 1—Illumination from the sun and sky on a horizontal plane for different altitudes of the sun.

their results. For example, Kunerth and Miller,[5] using a MacBeth illuminometer have measured the illumination on a horizontal plane, as a function of the altitude measured in degrees of the sun above the horizon. Fig. 1, reproduced with permission of the Illuminating Engineering Society, illustrates some of their results. Although these data were taken at latitude 42° N and longitude $93\frac{1}{2}$° W, they may be used, with sufficient accuracy for the present purpose, in almost any locality. The altitude of the sun, as given by Kunerth and Miller, for different hours of the day during several illustrative days of the year is shown in

[5] W. Kunerth and R. D. Miller, "Variations of intensities of the visible and of the ultraviolet in sunlight and in skylight," *Trans. Ill. Eng. Soc.*, vol. 27, pp. 82–94; January, (1932); and vol. 28, pp. 347–353; April, (1933).

Fig. 2. These curves, unlike those of Fig. 1, must be corrected when they are applied to any other latitude or relative position in a time zone. As shown, they represent quite closely the situation in New York City or Philadelphia.

The curves of Fig. 1 show the average of data taken during a great many days. Wide variations from these average values may be expected on particular days. Nevertheless, a fair estimate can be made of the illumination to be expected at a given time by combining the information included in the two figures. For instance, on November 5 at 4:30 P.M., an illumination of about 350 foot-candles can be ex-

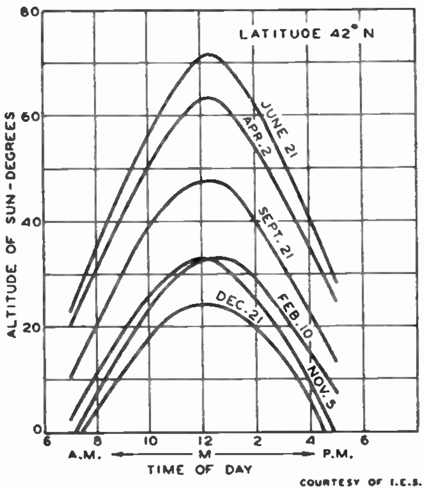

Fig. 2—Altitude of the sun for different times of the day and year at 42° N latitude and 93½° W longitude.

pected in sunlight if the day is clear, or about 200-foot candles if the scene is in the shade or the sky is cloudy. As another example, on June 21 at noon, an object in full sunshine will be illuminated with about 9100 foot-candles; in the shade, about 1000 foot-candles. In the latter example, it is interesting to notice that the illumination of an object in the shade will probably be greater on a cloudy day than on a perfectly clear one.

The reflection coefficients of many surfaces have already been measured; of the information available, the list in International Critical Tables is one of the most complete. Several illustrative items are given in Table I.

The surface brightness of a simple scene may be estimated by substituting in (1) values taken from Fig. 1 and Table I. A shady football field at 4:30 P.M. on November 5 can be expected to have a surface

TABLE I

Material	Reflection Coefficients	Material	Reflection Coefficients
Snow	0.93	Plaster	0.65
White paint	0.71	Brown soil	0.32
Light gray paint	0.49	Green leaves	0.25
Medium gray paint	0.30	Black velvet	0.01

brilliance of about $(0.25 \times 200)/\pi = 16$ candles per square foot (assuming R for grass to be the same as for green leaves). Near noon on June 21, the brown soil in the infield of a baseball diamond will show a surface brightness of about $(0.32 \times 9100)/\pi = 930$ candles per square foot.

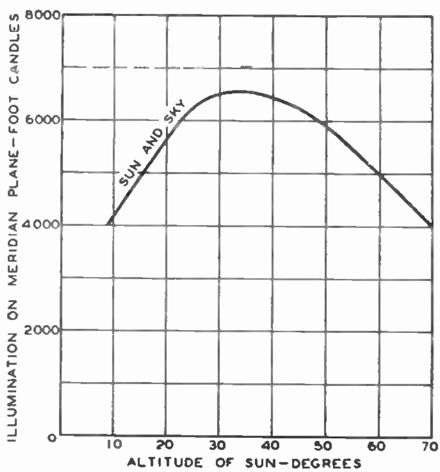

Fig. 3—Illumination from the sun and sky on a meridian plane for different altitudes of the sun.

Few scenes are as simple as those which have been assumed so far. For one thing, the subject of interest in a picture being transmitted is more likely to be vertical than horizontal. Therefore, from the data of Fig. 1, we have computed the illumination on a vertical surface facing the sun, as a function of the sun's altitude. Results are shown by the curve of Fig. 3, which is much flatter than the curve for a horizontal surface. When the sun is low and the light strikes the object squarely, much light is lost by atmospheric absorption; at midday, although atmospheric losses are small, the principal illumination is from the sky and reflection from near-by objects. It is interesting to observe that on clear summer days some scenes will be brighter in midmorning and midafternoon than at noon.

Reflected light must be taken into account. This is of particular importance in the case of vertical surfaces where reflection from nearby objects will change the expected illumination. In one test we found that the brightness of a vertical wall was reduced fifteen per cent when a large sheet of white cardboard on the ground was covered with black velvet. Snow may also cause a very considerable increase, and white billowy clouds in the sky have some effect.

Local conditions may still further affect the situation. These have been treated quite thoroughly by J. E. Ives and co-workers,[6] who simultaneously measured illumination in New York City where the air was very smoky, and at a point several miles distant where the air was comparatively free of smoke. The measured illumination in the city was usually less than at the outside station, and was sometimes as much as fifty per cent lower. In general, the loss was greatest when the sun was low in the sky, the sky cloudy, the relative humidity high, and the wind velocity low.

The color of light from the sky is usually unlike that from the sun. Due to scattering, skylight is stronger in the blue section of the spectrum than direct sunlight. Reflection surfaces which change the color of the illumination of the scenes, as well as the differential reflection of the scenes themselves should, for completeness, be taken into account. This will be treated more thoroughly later in the paper.

Although the data of Kunerth and Miller illustrate quite well what average illumination is to be expected at different times and under different conditions, they do not show the rapid fluctuations that may occur. Ives has illustrated, by means of several curves, these rapid variations. On a clear day the average illumination due to the sun may be as high as 10,000 foot-candles, but clouds passing over the sun may cause this to drop within one minute to 3000 foot-candles. One minute later the illumination may return to its original value. Such changes are not particularly noticeable to the eye, but may be quite bothersome in the use of a television transmitter. Smoke on a clear day can also cause variations which are, however, much smaller.

From the discussion so far given, the conclusion may be drawn that the probable brightness of a simple scene can be computed from information already available in the literature, but that the number of doubtful factors involved in an average scene is great enough to make the answer an approximation, at best. For this reason we have depended chiefly upon the results obtained by actually measuring the brightness of hundreds of typical subjects for a television broadcast. Because of its convenience, a Weston exposure meter was used for this survey. The spectral sensitivity of the meter is much the same as that

[6] "Studies in Illumination—Part III," Public Health Bulletin, No. 197, U. S. Treasury Department.

for the human eye, so that the readings are comparable with the surface brightness calculated by the method previously described.

An exhaustive investigation of the brightness of different views would involve measuring separately each object in the field of view. Because of the labor involved, and the fact that approximations are sufficiently accurate for the purpose, we have recorded only the average brightness in most cases. Table II gives several representative

TABLE II

Scene	Location	Time (E.S.T.)	Date	Weather	Surface brilliance (candles/sq. ft.)
Sixth Avenue	New York, N.Y.	9:30 A.M.	4–25–35	Clear	6½
Sixth Avenue	New York, N.Y.	1:15 P.M.	4–25–35	Overcast	40
Times Square	New York, N.Y.	1:30 P.M.	11–6–34	Light rain	40
Parade	East Orange, N.J.	10:30 A.M.	11–29–34	Light rain	40 to 60
Street	Rockland, Me.	1:15 P.M.	7–5–36	Overcast	100
Street	Warrenton, N.C.	3:15 P.M.	6–30–35	Clear	130
Street	Harrison, N.J.	3:30 P.M.	8–15–34	Hazy	130
Street	Harrison, N.J.	9:30 A.M.	8–15–34	Rain	16
River	New York, N.Y.	2:30 P.M.	10–24–35	Hazy	50
River	Pennsville, Del.	1:30 P.M.	6–29–35	Hazy	350
Bay	Cape Charles, Va.	10:00 A.M.	6–30–35	Clear	250
Beach	Atlantic City N.J.	2:00 P.M.	8–18–34	Hazy	500
Football game	New York, N.Y.	1st quarter	11–17–34	Clear	55
		2nd quarter		Hazy	50
		3rd quarter		Hazy	27
		4th quarter			16
		End			2
Baseball game	New York, N.Y.	1:00 to 3:40	9–8–35	Clear	70 to 100
Snow bank	Harrison, N.J.	10:00 A.M.	1–24–35	Bright sunshine	700
Open field	Bethel, N.C.	3:45 P.M.	7–1–35	Severe thunderstorm	2

readings out of the many which were taken under various weather conditions at different localities, times of day, and times of the year. The readings illustrate the tremendous differences that arise. Since television pickup devices are already known to be capable of operation under favorable conditions, it is of interest to consider the cases when the surface brightness of scenes is low. These occasions are most likely to be near sunrise or sunset, during severe storms, or when the light of the sun and sky is cut off by trees or tall buildings.

The unfavorable conditions were studied at greater length by recording the brightness of one particular subject in all kinds of weather, and by checking the variations of light in the neighborhood of tall buildings. Fig. 4 illustrates the brightness of a factory yard at Harrison, New Jersey, under many conditions. During rainy days the average values dropped to as low as seven candles per square foot, and on clear days rose to as high as 130 candles per square foot. Tremendous variations over short periods of time have been observed; in one case the brightness dropped from 100 to eight candles per square foot in less than two hours. Fig. 5 shows typical variations in brightness found in the shadows near tall buildings. At each location observations were made in several directions. In the morning, when it was clear, readings

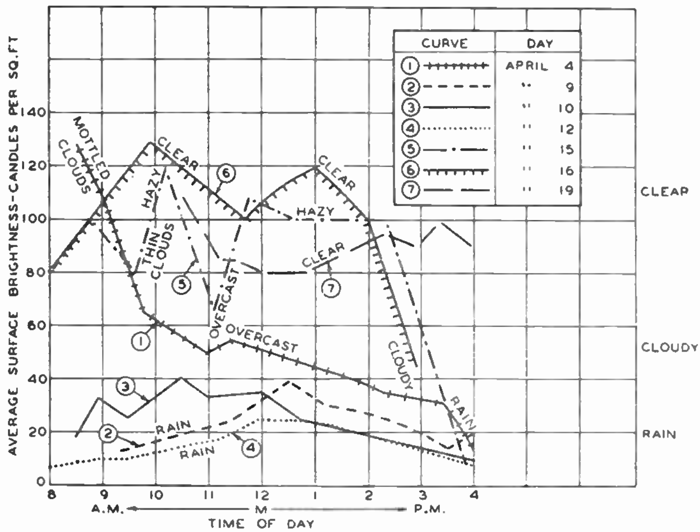

Fig. 4—Average surface brightness of a scene at Harrison, N. J., under various weather conditions during different times of the day.

ranged from two to 100 candles per square foot. Later in the day, when it was cloudy, the brightness ranged from twenty to 130 candles per square foot.

To summarize the matter of brightness of scenes encountered in

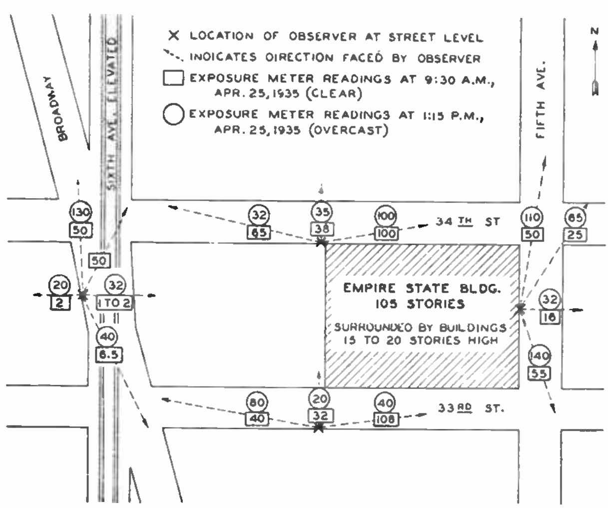

Fig. 5—Surface brightness at points around the Empire State building in clear and in cloudy weather.

nature, it may be said that almost any degree may be found; certainly from practically zero at night to over 1000 candles per square foot on a snow-covered mountain. However, during daylight hours most outdoor subjects fall in the range from twenty to 200 candles per square foot.

Brightness Necessary for the Transmission of a Television Picture

The discussion so far has disregarded the nature of the television equipment to be used for the transmission of a picture. We have been particularly interested in the performance of the Iconoscope as a television pickup device. Hence, we have compared its sensitivity with the brightness available in natural scenes.

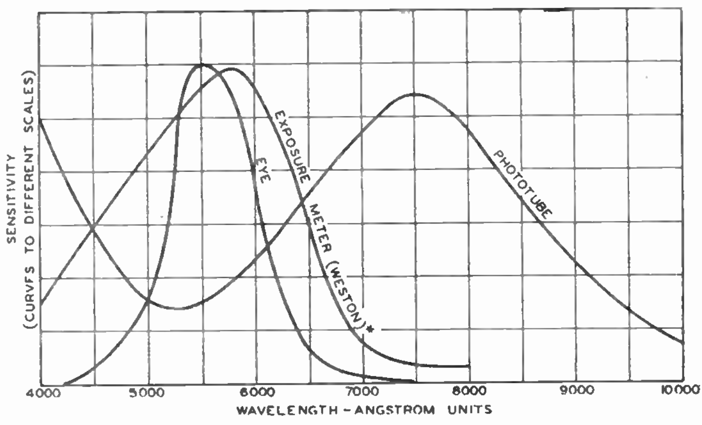

Fig. 6—Relative spectral sensitivities of the eye, the Weston exposure meter, and the usual caesium silver-oxide phototube.

One matter to be considered in this regard is the relative spectral sensitivities of the eye (upon which the discussion so far has been based) and the Iconoscope. Fig. 6 shows the relative spectral sensitivities of the eye and the Weston exposure meter while Fig. 7 shows that for the present-day Iconoscope. The curves for the eye and the Weston meter are substantially the same but that for the Iconoscope is quite different. In order to determine whether the surface brightness as measured can be applied to the Iconoscope the brightness of some subjects has been measured both with the Weston meter and a phototube photometer,[7] with a special phototube having, as Fig. 7 shows, a spectral sensitivity quite similar to the Iconoscope. These measurements show that shadows will appear relatively about twenty-five per cent brighter to

[7] Designed by T. B. Perkins of RCA Manufacturing Company, Inc.

the Iconoscope than they do to the eye. The explanation is that the Iconoscope has a lower relative sensitivity in the longer visible wave lengths (except the red). Therefore, in shadows, where blue skylight is used instead of direct sunlight, the Iconoscope will be affected less than the eye. Also, blue objects in sunlight appear about twenty-five per cent brighter to the Iconoscope than to the eye, while yellow objects appear about as much dimmer. In Fig. 6 is also given the spectral sensitivity curve for the usual caesium silver-oxide phototube. This has a peak in the red and a minimum in the blue while the present Iconoscope has no read peak and is rapidly rising in the blue. In case Iconoscopes are made in the future with a spectral sensitivity more like the usual

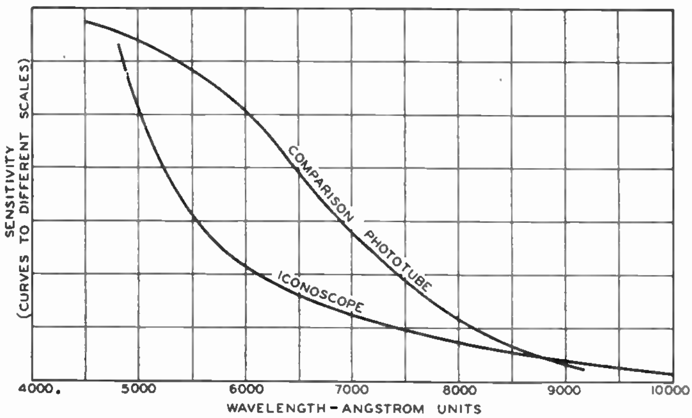

Fig. 7—Relative spectral sensitivities of the present Iconoscope and a comparison phototube.

phototube, readings have been taken with this type of phototube in the same phototube photometer. The results are about opposite from those found with the present Iconoscope. Shadows appear relatively darker to the phototube than they do to the eye. Also red objects appear about twenty-five per cent brighter to the phototube, while blue ones appear about twenty-five per cent dimmer. For highest accuracy, then, the differences in spectral sensitivity of the eye and the present Iconoscope or an Iconoscope with a red peak should be taken into account when the data of Table II are used. However, in view of the tremendous variations in natural illumination and the ability of the television transmitter to accommodate itself, these differences are, in practice, not very important.

Another matter which helps to determine the effective sensitivity of a television transmitter is the opening of the lens used to image the scene. It can be shown[4] that the illumination of the Iconoscope mosaic is given by the expression

$$I_m = \frac{0.54TB}{f^2} \text{ foot-candles}$$

where,

T is the transmission of the lens

B is the surface brightness of the scene in candles per square foot.

f is the ratio of the principal focus of the lens to its effective diaphragm opening.

For our work an f 4.5 lens with a six-inch focal length was available, and an Iconoscope having a mosaic slightly larger than four by five inches was used. In a commercial installation these factors, particularly the lens openings, may be changed to give still better results.

About two years ago a series of tests was made using an Iconoscope, one of the best then available, to find how much light was needed for the transmission of a "good" television picture. By "good" is meant one in which the sharpness of definition, contrast, and brightness are not so altered from their values in the scene transmitted as to be objectionable to the majority of observers. The amplification which could be used was found to be limited by noise from the first stages of amplification, and not from the Iconoscope. It was necessary, therefore, that the latter deliver enough signal to make the shot and thermal-agitation noises relatively very small. This signal delivered by an Iconoscope increases with the beam current used. An increase in beam current, however, was found to increase the spurious signal or so-called "dark spot" which is caused by redistribution of secondary electrons.[8] The signal output, therefore, has to be limited to a value which will allow compensation for the "dark spot" signal. Under these conditions the conclusion was reached that an average brightness of fifty candles per square foot was sufficient to give a reasonably satisfactory picture of an average scene on a cloudy day with an f 4.5 lens to focus the light on the Iconoscope mosaic. In this case, the brightest object in the field of view had a brightness of sixty-five candles per square foot, while the dimmest had fifteen candles per square foot. Another scene having an average brightness of thirty candles per square foot was generally agreed to be equally satisfactory for transmission. In the second test the brightest object had a brightness of 100 candles per square foot, and the darkest had five candles per square foot. The average brightness needed for satisfactory operation, then, is less when the contrast between objects of interest and the background is high. As a result of

[8] A more complete discussion of this spurious signal is given in a paper by V. K. Zworykin, "Iconoscopes and Kinescopes in television," *RCA Rev.*, vol. 1, p. 75; July, (1936).

these and other experiments, the conclusion was reached that an *f* 4.5 lens would allow pickup of most views brighter than fifty candles per square foot, or that if an *f* 2.7 lens were used an average brightness of twenty candles per square foot would be sufficient.

This performance is very good; five years ago it might have been called miraculous. Yet this performance is not sufficient to permit the transmission of every subject that might be of interest to the public, nor does it allow otherwise acceptable scenes to be transmitted with adequate depth of focus. Research and development work have, therefore, continued with the object of providing still higher sensitivity.

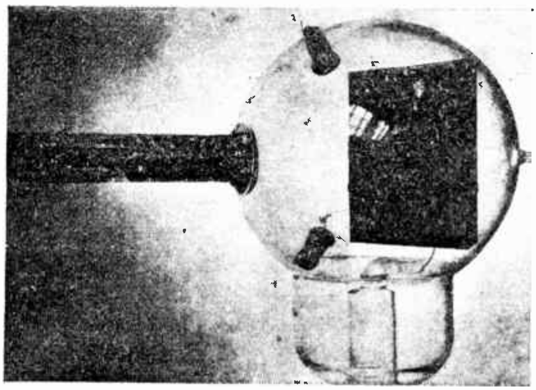

Fig. 8—Iconoscope with silver evaporation sensitization.

The problem of sensitizing an Iconoscope is more complicated than that of sensitizing a phototube, because it is necessary to maintain very high insulation between all of the tiny photosensitive particles on the mosaic, as well as to obtain good photoemission. In the past, the difficulty of maintaining good insulation has been a limitation to the photosensitivity attainable; the presence of enough caesium to provide optimum photoemission would cause too much conductivity. A film of cryolite evaporated on the mica sheet before the mosaic is formed has been found to give a marked improvement in the insulation, in fact, enough to make practical a silver-evaporation sensitization process.

According to this method, the mosaic is first sensitized as usual by oxidizing it, admitting caesium, and baking. Then, a very thin coating of silver is evaporated from filaments provided for the purpose, and the tube is baked again. A photograph of such an Iconoscope is given in Fig. 8; the side tubes in front of the mosaic contain the filaments for evaporating silver.

Not only has the sensitization process been improved, but also the quantity and purity of materials have been controlled more carefully than in the past. As a result, the best Iconoscopes today are more than three times as sensitive as the best of two years ago, and the average is more than correspondingly improved.

Quite faithful reproductions of scenes having an average brightness of fifteen candles per square foot may now be transmitted with an f 4.5 lens. The received pictures are not perfect, especially in regard to shading, but they are substantially the same as the original scene so that the entertainment value is little affected. An equally good picture, except for depth of focus, could, of course, be transmitted of a scene

Fig. 9—Photograph of a received image when the lighting of the transmitted scene was optimum for the mosaic of the Iconoscope.

having a surface brightness of five or six candles per square foot if an f 2.7 lens is used. When the light is still less, it is possible to identify familiar objects, though the quality of the picture is definitely impaired. Using an f 4.5 lens we have been able to reproduce scenes where the average brightness was as low as 2.5 candles per square foot. Such pictures, however, do not have much entertainment value.

When greater illumination is available at the transmitted scene, pictures of better quality are obtained. Fig. 9 is a photograph of a received picture which was taken when the lighting of the transmitted scene was optimum for the mosaic of the Iconoscope.[9] The use of the word optimum, of course, implies that the illumination on the mosaic can be too strong as well as too weak. This has been found to be the

[9] This photograph is shown through the courtesy of R. M. Morris of the National Broadcasting Company.

case. Because only a small electric field is available for drawing photoelectrons away from different parts of the mosaic, the photocurrent on the strongly illuminated portions of the mosaic can be saturated. This saturation will in turn lower the signal output from these strongly illuminated parts. In using the Iconoscope, it is found that saturation shows up as a reduction in contrast. Consequently, when plenty of light is available, the illumination on the mosaic is adjusted for maximum contrast.

As a result of the tests which have been made, it is possible to tell quite well what subjects would be available for transmission if television broadcasting should start now. Motion pictures and studio entertainment are, of course, to be expected. Outdoor scenes of parades, baseball games, and races can be handled until near sunset under almost any weather conditions. It is even possible that transmission of some night baseball games is technically possible; newspaper descriptions of Crosley Field, in Cincinnati, indicate that the lighting is sufficient. Many football games would be satisfactory subjects, although in some cases the light would be too dim at the end of the game unless the starting time were advanced by half an hour.

The mention of these possibilities does not necessarily mean that such sports events will be shown as soon as television broadcasting is started. Economic considerations, rather than technical ones, may determine which subjects are feasible. In the meantime, work of improving the Iconoscope is being continued; there is hope that its sensitivity may be increased still further. Some day it will be able to "see" anything that the human eye can see and some things that the human eye cannot observe.

ICONOSCOPES* AND KINESCOPES** IN TELEVISION

By

V. K. Zworykin

RCA Manufacturing Company, Inc., Camden, N. J.

TWO extremely important elements in any television system are the pickup device which converts the light image into electrical signals, and the viewing arrangement transforming the electrical signals back into visible images. In fact, the success or failure of a televison system depends perhaps more on these two links than on any other part of the chain. In its present project RCA Manufacturing Company is using the Iconoscope and Kinescope for dissecting and resynthesizing images, and it is the purpose of this discussion to explain the operation of these instruments and point out the reasons for selecting them over other devices designed to serve as pickup and viewing equipment.

Historically, the development of any form of television had to await a means of converting a light signal into a corresponding electrical impulse. This step became possible through the discovery of the photoconductive properties of selenium in 1873. Within two years after this discovery, Carey proposed to make use of the properties of selenium in the solution of the problem of televison. His suggestion was to construct a mosaic consisting of a great number of selenium cells, in a sense imitating the retina of the human eye. These cells were to be connected to shutters or lamps in corresponding positions on a viewing board. Although the suggestion was made in 1875, the device was not put into operation until 1906 when Rignoux and Fournier used this arrangement to transmit simple patterns and letters. Their mosaic consisted of a checker-board of sixty-four selenium cells. Each cell was connected to a shutter on a viewing screen which was also made up of sixty-four elements in positions correspond-

* From the Greek word "Icon" meaning an image, and "scope" signifying observation. Trade Mark Registered U. S. Patent Office.

** From Greek word "Kineo" meaning movement. Trade Mark Registered U. S. Patent Office.

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ing to those in the pick-up screen. When a picture was projected on the selenium cells the resistance of those illuminated decreased allowing an electric current to flow which opened corresponding shutters on the viewing screen. A light behind these shutters made the reproduced picture visible.

The idea of dividing the picture into elements, converting the illumination on each element into electric current and sending the signal from each over individual wires is practical for a small number of divisions or picture elements and for transmission over short distances, but is useless as a means of producing pictures of the standard required of television today.

The next step was proposed by Nipkow in 1884. Instead of using individual wires connecting each picture element, he suggested sending the information from one element at a time over a single communication channel and then reassembling this information again at the viewing screen. This process was to be carried out at such a rate that the picture appeared continuous due to persistence of vision. The means proposed to accomplish this point-by-point transmission was the scanning disc. At the time of its invention the necessary technique of handling and amplifying small currents had not yet been developed so that it was a number of years before this scanning principle could be put to practical use. However, the principle was sound and the scanning principle has been the basis of all televison systems since then.

While this development represents a great step forward, it was only attained at considerable expense of available picture signal. The loss is due to the fact that each element only contributes to the picture a small fraction of the total time, whereas with the first system suggested each element operated continuously. To make this clear, consider again the simple sixty-four element mosaic used by Rignoux and Fournier. Each photoelectric element was connected to the viewing screen by a separate conductor and the picture to be transmitted projected continuously on all the elements, so that a signal current passed through every light sensitive element all the time. To reduce the scanning system to a comparable case, assume that we have the same mosaic of sixty-four photosensitive elements, but that they are all connected to a common communication channel. The elements are covered with shutters (i.e., the scanning disc) which allow only the light from one element of the picture at a time to reach its corresponding photocell. These shutters are opened one at a time in rotation covering the entire picture twenty or thirty times a second. Thus each light sensitive element is only operating for a fraction of the total time equal to one over the number of picture elements, in this case one-sixty-fourth of the time.

In order to regain this lost signal and yet retain the principle of scanning, the development of the Iconoscope was undertaken. To illustrate the method of attack, consider again the sixty-four element array of photocells. Instead of scanning the elements

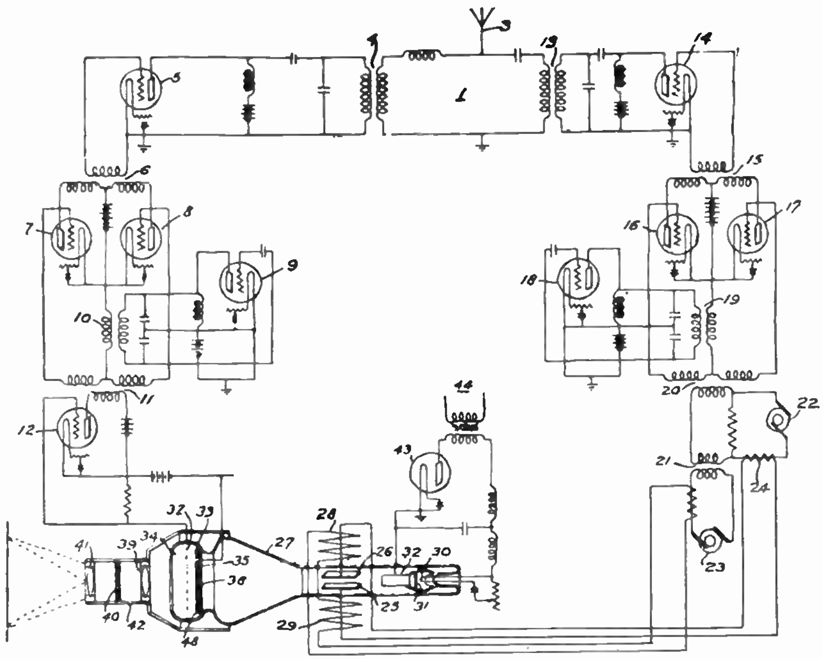

Figure 1

with shutters, assume that each element is connected to the contact points of a switch which connects them in rotation to the main communication channel. Thus the scanning is accomplished by means of a commutator switch.

So far, we have gained nothing over the previous method of scanning, but now if a condenser is placed across each of the photocells in such a way that it accumulates the entire charge released by the action of the light during the time the element is not connected to the communication channel, this charge can be used when the commutator switch again makes contact with this element. Therefore, photoelectric current is being released by every element continuously and this charge stored in the condenser belonging to that element until it is needed at the end of a scanning cycle.

The reduction of this principle to some practical form is obviously a difficult problem. The number of individual photocells and condensers required for a 360-line picture with a 4 to 3 aspect ratio will be of the order of 173,000 units, and it is quite apparent that a screen composed of that many conventional photocells and condensers is out of the question.

A solution devised by the author some years ago was to build up a mosaic screen which contained the equivalent of a vast

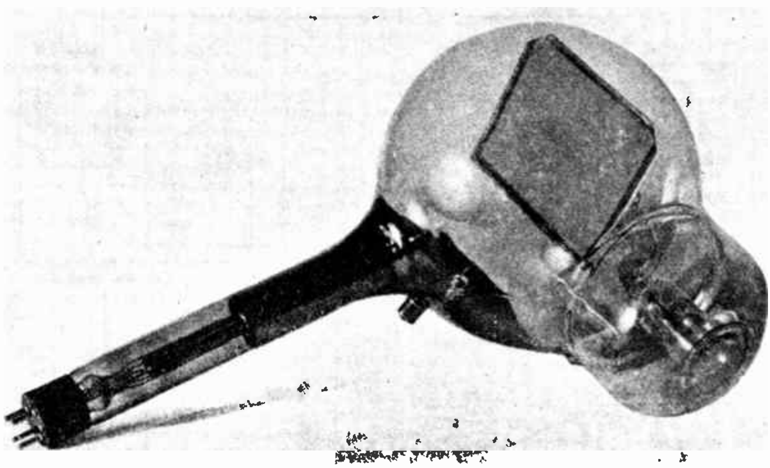

Figure 2

number of photosensitive elements and condensers. This mosaic was mounted in a cathode-ray tube in such a way that an electron beam could be used to commutate the elements. Fig. 1 shows one of these tubes together with its associated circuits taken from one of the author's early patents.[1] Aside from the advantage gained through the application of the principle of storing the charge on each element for the entire picture time. this tube had the additional advantage that it involved no mechanical moving parts such as a scanning disc, mirror screw or drum, the scanning being done electrically.

Although the first of this type of tube was built as far back as 1923, many years of research and development had to be undertaken before it was perfected sufficiently to meet the re-

[1] V. K. Zworykin, Patent No. 1,691,324 "Television System."

quirements of a satisfactory television system. The history of this development is interesting, but is somewhat outside the scope of this paper, which will be limited to a discussion of the tube as it is today.

The Iconoscope, as this type of tube has been named, is shown photographed in Fig. 2. It consists of an electron gun and photosensitive mosaic enclosed in a highly evacuated glass envelope. The arrangement of these elements is shown diagrammatically in Fig. 3.

The electron gun produces a narrow pencil of cathode rays which serves, as will be shown later, as a commutator to the tiny

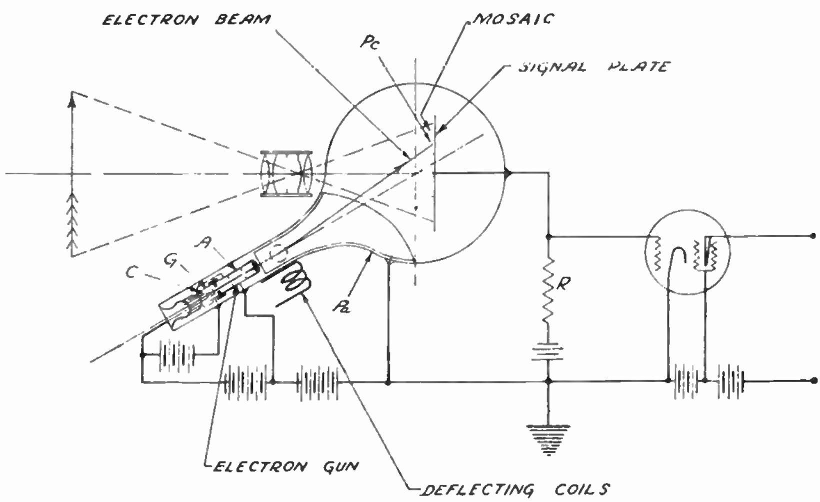

Figure 3

photocells on the mosaic. The gun is in reality a form of electron projector which concentrates the electrons from the cathode onto the mosaic in a very small spot. The electron optical system consists of two electron lenses which are formed by the cylindrically symmetrical electrostatic fields between the elements of the gun. Fig. 4 shows diagrammatically the arrangement of this gun, together with the equipotentials of the electrostatic fields making up the electron lenses. Below this diagram is the approximate optical analogue. Details of the gun construction are as follows: The cathode is indirectly heated with its emitting area at the tip of the cathode cylinder. It is mounted so that the emitting area is a few thousandths of an inch in front of an aperture in the control grid. A long cylinder with three defining apertures whose axis coincides with that of the cathode cylinder and control grid serves to give the electrons their initial acceleration and is known as the first anode. A second cylinder coaxial with the first anode and of somewhat greater diameter serves as second anode and gives the electrons their final velocity. The second anode is in general formed by metalizing the neck of the Iconoscope bulb, as shown in Fig. 3. The gun used in the Iconoscope is designed so that it will concentrate a beam current of from one-half to one microampere into a spot about five mils in diameter. Under ordinary operating conditions, a potential of

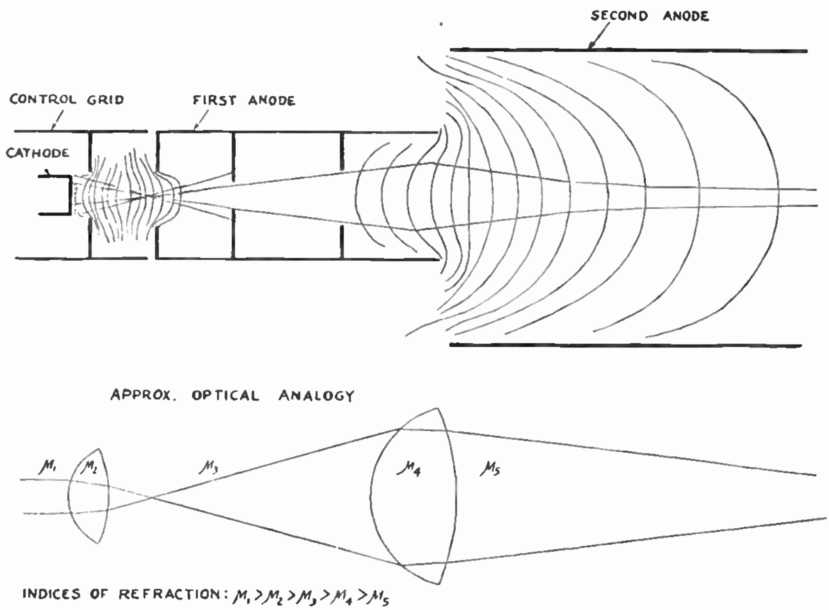

Figure 4

about a thousand volts is applied between the cathode and second anode and the voltage of the first anode adjusted until minimum spot size is obtained. The exact value of the beam current to be used will, of course, depend upon the type of picture to be transmitted and the exact conditions of operation.

The beam from the gun is made to scan the mosaic in a series of parallel horizontal lines repeated at thirty cycles per second. This is accomplished by two sets of magnetic deflecting coils arranged in a suitable yoke and slipped over the neck of the Iconoscope. These sets of coils are driven by two special vacuum tube generators supplying a saw-toothed current wave, one operating at picture frequency supplying the vertical deflecting coils, the other at horizontal line frequency driving the second set of coils.

The element which characterizes the Iconoscope is the mosaic. It consists of a vast number of photosensitive globules mounted on a thin mica sheet in such a way that they are insulated from one another. The back of this sheet is coated with a conducting metallic film which serves as a signal plate and is connected to the input of the picture amplifier. The appearance of the mosaic is shown in Fig. 5. Such a mosaic may be formed in a variety of ways. For the standard type of mosaic the silver globules are formed by reducing particles of silver oxide dusted over the mica. Under proper heat treatment the silver globules reduced from the oxide will not coalesce but will form individual droplets.

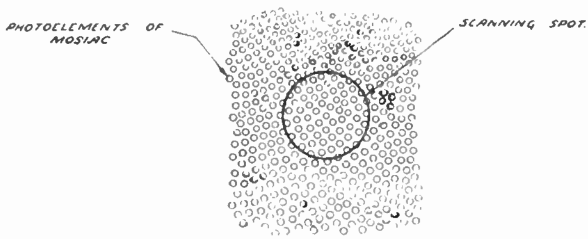

Figure 5

These droplets are sensitized after the mosaic has been mounted in the tube and the tube evacuated. The sensitization is similar to that used in the ordinary cesium photocell, that is, the silver is oxidized, exposed to cesium vapor, and then heat treated. The result is that the photoelectric response of these globules is about the same as that of a high vacuum cesium photocell, both in sensitivity and spectral response. The spectral characteristic is shown in Fig. 6. The cut-off in the violet part of the spectrum is due to absorption of the glass walls. It is evident that the Iconoscope with a quartz window is sensitive from well into the infrared, through the visible, and into the ultra-violet. Actual tests have produced images using radiation from 2000 Å down to more than 9000 Å.

The mica on which the silver droplets are mounted serves to insulate them from one another and further is made thin enough so that the capacity between each globule and the metallic signal plate will be reasonably large. The uniformity of cleavage sheets of mica, together with their excellent insulating properties, low dielectric hysteresis and low loss make them very suitable for this purpose. Other insulating materials, however, can be used;

for example, a thin film of vitreous enamel on a metal signal plate has proven very satisfactory.

The mosaic is mounted in the tube with the silver beads facing the beam. In order that the optical image may be focused on its front surface, it is placed in the tube in such a way that a normal to its face makes an angle of 30° to the axis of the electron gun.

In essence, the Iconoscope may be thought of as a plain mosaic made up of a great number of individual photocells, all connected by capacity to the common signal plate and commutated by the

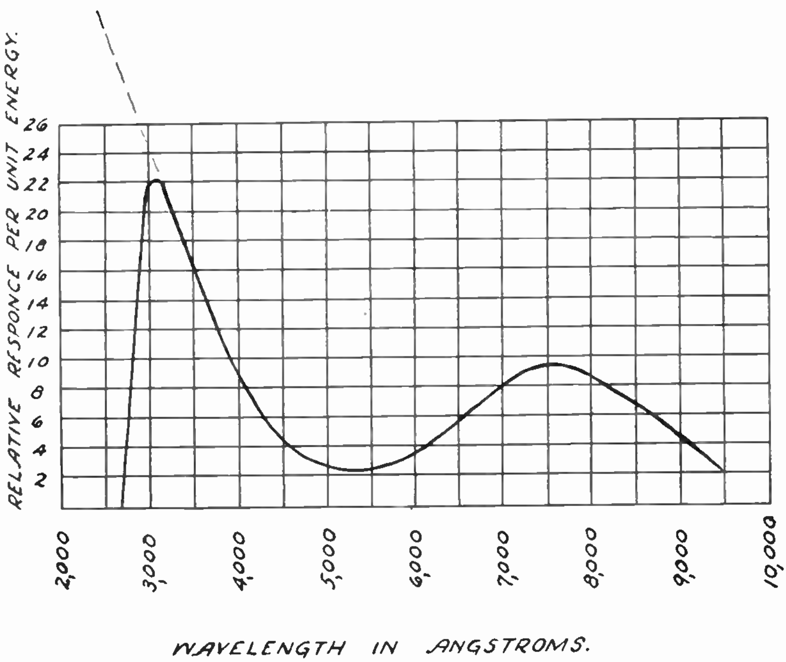

Figure 6

scanning beam. The fundamental cycle of operation is as follows:

Every silver globule making up the mosaic is photosensitized so that when a light image is projected on the latter the light causes electrons of a number proportional to the light brilliance to be emitted from each illuminated minute size photosensitive area. The resulting loss of electrons leaves each photosensitive area at a positive potential without respect to its initial condition which potential is then proportional to the number of electrons which have been released and conducted away so that the mosaic tends to go positive at a rate proportional to the light falling on it. As the electron beam scans the mosaic, it passes over each element in turn, releasing the charge it has acquired and driving it to equilibrium. Due to the fact that each element is coupled by capacity to the signal plate, the sudden change of charge of the elements will induce a change in charge on the signal plate and result in a current pulse in the signal lead connected to the amplifier. The magnitude of these pulses will be proportional to the intensity of the light falling on the scanned element. Thus the signal output from the Iconoscope will consist of a chain of current pulses corresponding to the light distribution over the mosaic. This chain can be resynthesized at the re-

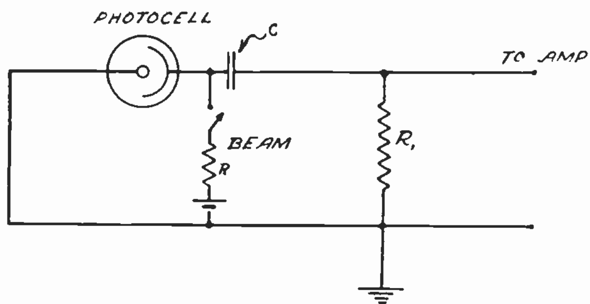

Figure 7

ceiver into a reproduction of the original image, as will be described later.

To clarify this cycle the equivalent circuit representation of a single element is shown in Fig 7. The beam is represented by the switch and series resistance R. This switch may be considered as being open except at such times as the beam is actually on the element. When the scanning beam moves off the element, the photo-emission from it starts to charge the condenser C, the rate of accumulating charge being proportional to the illumination on the element. In the next scanning cycle, the beam again sweeps over the element, closing the switch and discharging the element. During this discharging cycle the entire charge accumulated during the time the beam was not on the element must now flow through the input resistor R_1 generating an e.m.f. which is applied to the input of the picture amplifier.

In designing the mosaic, it is evident that the time constant of the circuit discharging the condenser C must be small enough to allow it to fully discharge during the time the beam is on the element. This condition requires that $C \times (R_1 + R)$ be less than the time the beam is on the element. In practice, this condition is not difficult to fulfill.

At this point it is interesting to compare the e.m.f. supplied to the amplifier by this storage system with the equivalent voltage from a non-storage system. The equivalent circuit for the non-storage case is shown in Fig. 8. The current through the input resistor R_2 will be:

$$I_s = \frac{F.s}{n}$$

where F is the light flux in the picture, s the sensitivity of photo-sensitive elements, and n the number of picture elements. The voltage to the input of the amplifier is:

$$V_2 = \frac{Fs\, R_2}{n}$$

In the storage case the charge accumulated by the element is:

$$Q = \frac{F.s}{n} t_p$$

where t_p is the picture time or $1/N$ for N pictures per second. When the beam strikes the element this charge leaves the condenser resulting in an average current of

$$i = \frac{Q}{t_e}$$

where t_e is the time the beam is on the element or $1/Nn$. This current is therefore:

$$i = \frac{F.s\, Nn}{n\ N}$$

and the voltage to the amplifier will be:

$$V_1 = FsR_1$$

Comparing the signal voltages generated in the two cases, we see that the ratio is:

$$\frac{V_1}{V_2} = \frac{FsR_1}{\frac{FsR_2}{n}} = n \qquad \text{where } R_1 = R_2$$

Where the number of picture elements is large as is the case in pictures with good definition, this gain in signal is extremely important. For example, the ratio in the case of a 360-line picture is 173,000 times the signal that could be obtained from the non-storage case.

In order to give a pictorial idea of the conditions on the surface of the mosaic, Fig. 9 is included. It represents the appearance of the charged image on the mosaic if it were visible to the eye. The region just behind the scanning beam is an equilibrium potential and therefore shows no visible image. As we examine the mosaic further away from the line just scanned, we find that

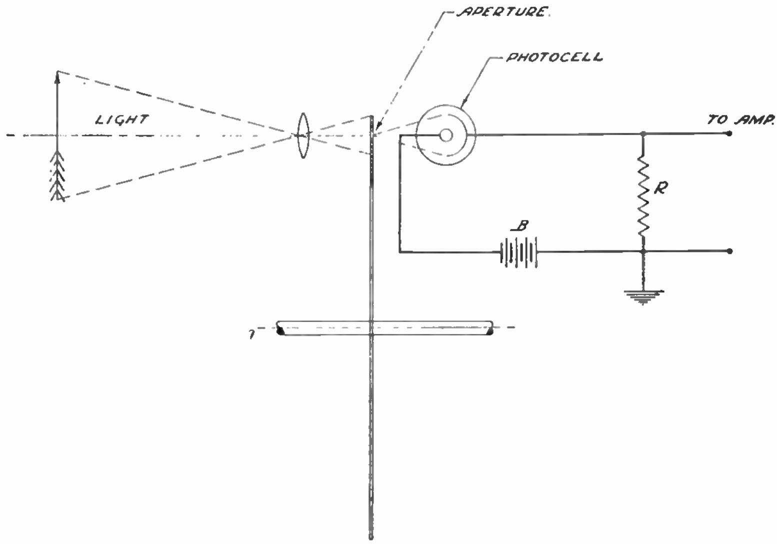

Figure 8

the charged image gets more and more intense since the elements have been charging for a greater length of time. Just ahead of the scanning beam the image reaches its maximum intensity.

The picture just drawn of the operation of the Iconoscope is very much simplified. A number of factors complicate this seemingly straightforward cycle. Among the most important of these complicating factors are the potential distribution over the mosaic and the redistribution of secondary electrons emitted from the elements under bombardment. If the average potential of the mosaic is measured in darkness while it is being scanned, it will be found to be between 0 and 1 volt negative with respect to the electrode which collects the electrons leaving the mosaic, that is, with respect to the second anode. However, the potential is not uniform over the surface of the mosaic. Elements directly under the beam are found to be in the neighborhood of 3 volts positive with respect to the second anode. As we investigate elements which have previously been bombarded, we find them less positive, until at a point one-quarter to one-third of the vertical distance along the mosaic from the point being bombarded, the potential has reached $-1\frac{1}{2}$ volts negative with respect to the collector. The rest of the mosaic is found to be at $-1\frac{1}{2}$ volts.

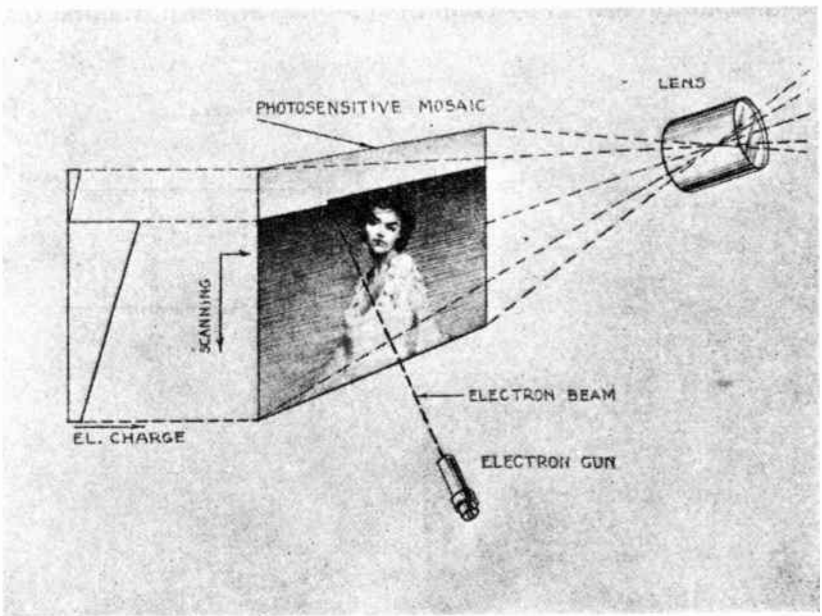

Figure 9

Cathode-ray oscillograph measurements of the potential distribution over the mosaic shows that it can be mapped somewhat as shown in Fig. 10.

In order to account for the potential distribution over the surface of the mosaic, it is necessary to consider what takes place among the secondary electrons emitted from the cesiated silver elements under bombardment. It is well known that when a cesiated silver surface is bombarded by an electron beam of the order of 1000 volts velocity, a secondary emission of 7 or more times the primary bombarding current can be collected. However, since the mosaic elements are insulated they must assume, when in equilibrium, a potential such that the secondary emission current equals the bombarding current. This potential is found to be about 3 volts positive with respect to the second anode. In the case of the mosaic in darkness it is obvious that the average secondary emission current leaving the mosaic must also be equal to the beam current since the mosaic is an insulator. Thus it must come to an equilibrium potential such that the average current escaping to the second anode equals the bombarding current.

Perhaps we should digress at this point and discuss more

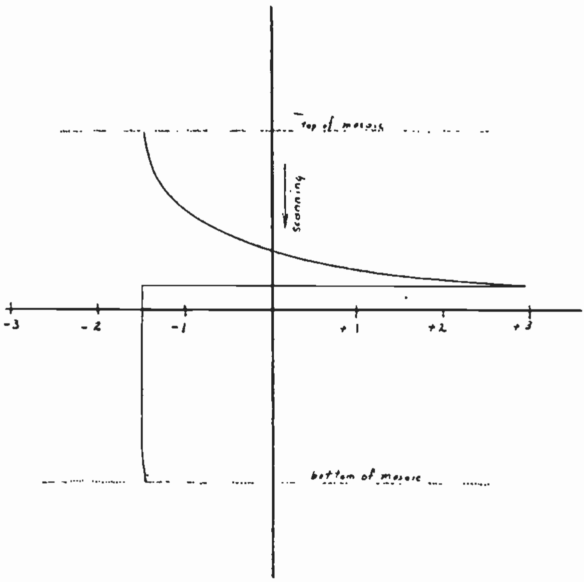

Figure 10

fully the mechanism by which an element acquires this positive equilibrium potential. Measurements of the velocity distribution of the secondary electrons from a bombarded surface show that they can be represented by a distribution curve such as shown in Fig. 11. In this figure the abscissa gives the velocity in electron volts of the emitted electrons, while the ordinates give the current per volt range in velocity composed of electrons having a given velocity. If the target is surrounded with an electrode to collect secondary electrons, the current it can collect will depend upon its potential relative to the target. When this collector is at zero potential the current reaching it will equal the total secondary emission as represented by the total area under the curve in Fig. 11. As the collector is made more negative, the current decreases since some of the electrons leaving will not have sufficient velocity to reach the electrode and will be driven back to the target. The current reaching the collector at some negative potential V_1 will be given by the area under the distribution curve from V_1 to the highest velocity; in other words

As the potential of the collector is decreased further eventually it will reach a point where the current collected just equals the current in the primary beam. At this point no current flows in the external lead to the target under bombardment. Experiment shows that for a cesiated surface such as is used to make up the globules on the Iconoscope mosaic, this potential is in the

$$i_c = \int_{V_1}^{\infty} f(V)\ dV$$

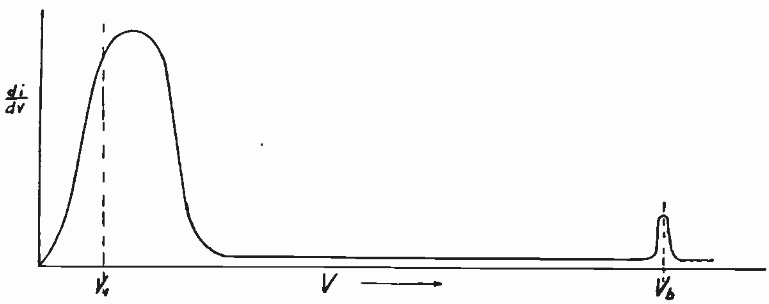

Figure 11

neighborhood of 3 volts. Hence, if an insulated target such as a mosaic element is bombarded more electrons will leave than arrive until the element reaches 3 volts positive, at which potential the element will be in equilibrium and the current arriving and leaving will be equal.

When the mosaic elements are scanned the secondary emission from them may be divided into three parts, one going to the second anode, another returning to the element itself, and a third being redistributed over the entire mosaic. This latter group which returns to the mosaic comes back as a more or less uniform rain of electrons having a maximum velocity of about 1½ volts.

This can be verified by removing a portion of the mosaic and substituting a metal sheet electrode in its place. If the mosaic, except for the substituted portion, is scanned and the current to the metal electrode measured, it will be found to decrease as the potential of the probe is decreased. At −1½ volts negative the current will be dropped to zero.

Let us now consider the operation of the Iconoscope in the light of the phenomena just discussed. In the first place, due to the potential of the mosaic, there is very little electrostatic field aiding the escape of photo-electrons from the illuminated elements. This means that the charging of the globules is dependent in a large measure upon the initial velocities of the electrons. Therefore, the photo-electric emission is not very efficient and becomes less so as the illumination is increased. It should be remembered, however, that the photo-emission occurs during the entire picture time, the charge being accumulated on the con-

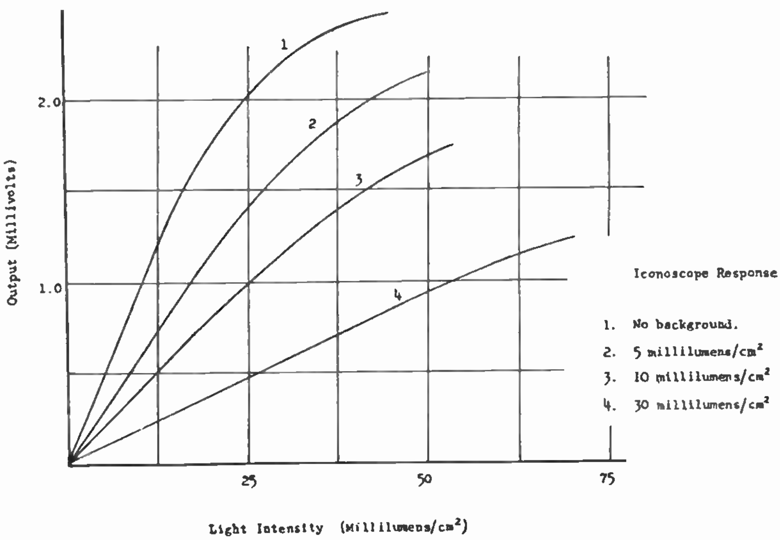

Figure 12

densers formed by the silver beads and the signal plate. In other systems, since photo-emission occurs only during the time a picture element is being scanned, there is a gain of the order of 10^5 to be had by using the storage system so that even if the abovementioned photoelectric inefficiency were insurmountable there is still a very great advantage in favor of the Iconoscope system.

As was pointed out in the discussion of the potential distribution, there is a line across the mosaic directly behind the scanning beam which is 3 volts positive with respect to the second anode, while just ahead of the beam the potential is in the neighborhood of $1\frac{1}{2}$ volts negative. There is, therefore, just ahead of the scanning beam, a row of elements which have a strong field aiding the leaving photo-electrons. This field very much increases the photo-sensitivity along this line and gives rise to a phenomenon known as line sensitivity. This phenomenon can be demonstrated very strikingly in the following way:

The image from a continuously run moving picture film (i.e., by removing the intermittent and shutter from a moving picture projector) is projected onto the mosaic of the Iconoscope. The film is run at such a rate that the frame speed is equal to the picture frequency of the Iconoscope and in a direction such that the image moves opposite to the vertical direction of scanning.

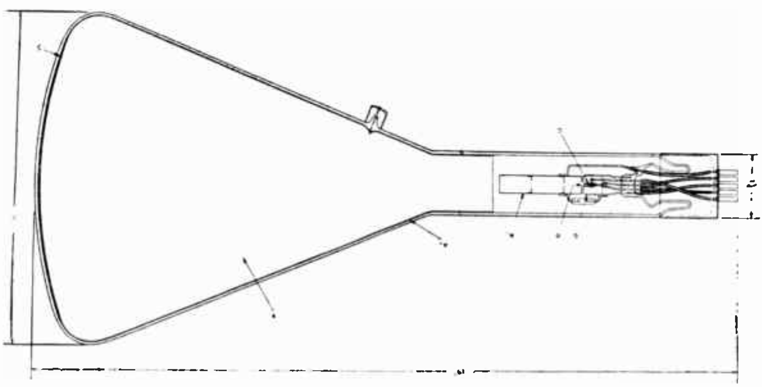

Figure 13

Under these conditions we find that the Iconoscope transmits a clear image of two frames of the moving picture film although, to the eye, there appears to be only a blur of light on the mosaic.

Thus we have two sources of signal, one the stored charge over the entire mosaic surface; the second from the sensitive line at the scanning beam. At low or normal light intensities by far the greater part of the signal comes from surface sensitivity, but under high illumination as much as 50% of the signal may come from line sensitivity.

As was pointed out above, due to the fact that the secondary emission is not saturated, some electrons from the point where the beam strikes the mosaic have not sufficient velocity to leave the mosaic entirely, but return to its surface as a shower of low-velocity electrons. The redistributed electrons act to some extent as a high resistance, short-circuiting the elements, since an element which is more positive than its neighbors tends to receive a greater share of these electrons. This resistance is, in effect, identical with that of the dynamic resistance of a triode tube and under normal operating conditions is high enough so that it does not produce a very serious loss in efficiency, but under high illumination where considerable difference in charge between nearby elements may be developed, this shunting resistance may become quite low with the result that there is a fairly large loss of signal.

This redistribution of electrons is furthermore responsible for the generation of a spurious signal. It appears as an irregular shading over the picture even when the mosaic is not illuminated. The cause of this signal is the variation in instantaneous secondary emission current escaping from the mosaic to the second anode. As has been pointed out, the average secondary emission from the mosaic must be unity, but when we consider

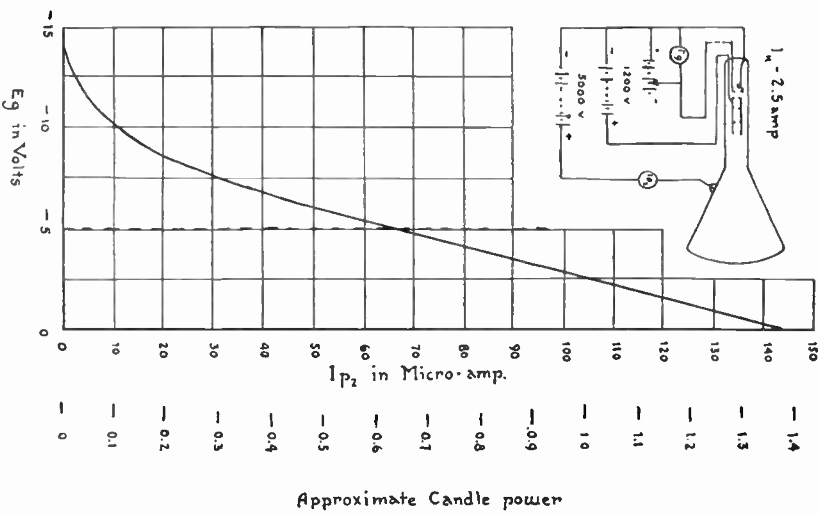

Figure 14

that a certain fraction of the secondary electrons from the point under bombardment returns to other parts of the mosaic, it is quite apparent that the instantaneous current leaving the mosaic may vary from point to point. This variation is produced by the lack of uniformity of potential and space charge over the mosaic.

It is interesting to note that if a clean sheet of metal is substituted for the mosaic, the spurious signal appears when it is scanned, provided the secondary emission is not saturated. The signal disappears, however, if the metal plate is made sufficiently positive or negative with respect to the second anode. Under these conditions, the secondary emission is either saturated or suppressed. In practice the effect of this signal can be eliminated by the introduction of a compensating signal. The spurious signal varies rapidly with beam current and under conditions of low beam intensity and moderately high illumination it is negligible compared with the picture signal.

So far, the scanning beam has been considered as some sort of commutating switch which sweeps over the mosaic. Actually, the beam does behave in just this way. The beam when falling on an element connects it through a resistance (dynamic, of course) to the second anode. This is obvious when we consider the action of the beam. As has been pointed out, the ratio of secondary electrons to primary electrons from a cesiated silver surface is about 7 when saturated. However, if the bombarded

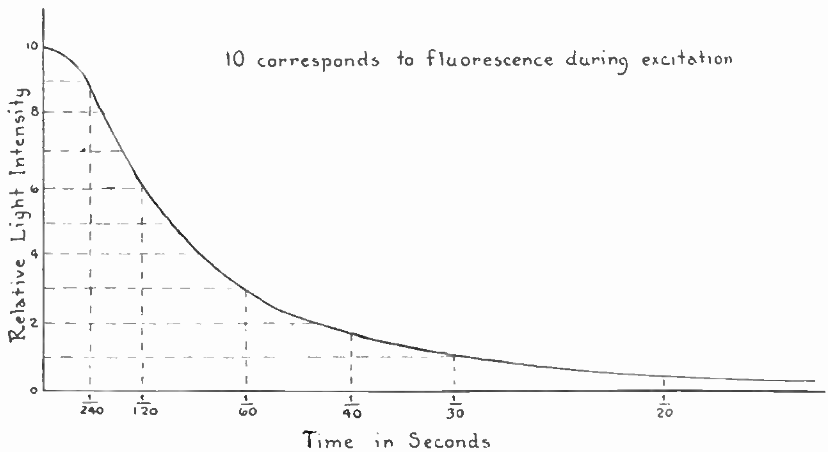

Figure 15

surface is made positive this ratio decreases, reaching unity at +3 volts and one-half at about 10 volts. From curves giving the secondary emission ratio of an element for various collector potentials, together with a knowledge of the beam current, the effective resistance connecting the bombarded element with the second anode can easily be estimated. This resistance turns out to be of the order of 10^6 ohms. If the beam current is too weak, it will not fully restore the illuminated element to equilibrium. Considering a stationary picture, and neglecting the effect of the redistribution of scattered electrons, this would not reduce the signal obtained from a given amount of light. However, it would cause a lag and consequent blurring of the image of a moving object. In the actual Iconoscope because of the rôle the beam plays in establishing the potential of the mosaic and because of the shunting effect of redistributed electrons, there is an optimum beam current at which the signal is a maximum for a given condition of light.

Taking into account the various factors tending to reduce the output of the Iconoscope, it is found that the net efficiency of conversion is in the neighborhood of 5 to 10%. In other words, the signal output is about 1/20 that which would be expected on the basis of the light flux reaching the mosaic, the saturated photo-emission of photo-electric elements, and the assumption

Figure 16

that the entire photo-current is stored by the mosaic. The efficiency of conversion is not constant, but as explained above depends on the amount of light used. The efficiency is a maximum at low light and decreases as the light is increased. This point will be considered again under a discussion of the actual performance of the Iconoscope.

Up to this point we have based our consideration of the relative merits of the storage and non-storage types of systems on a comparison of signal output alone. The recent development of the secondary emission multiplier makes it necessary to introduce other considerations into this comparison. The electron multiplier provides a means of amplifying a photoelectric current to almost any desired extent without introducing any additional "noise" into the signal. It might seem, therefore, that we could amplify the minute photo-current obtained by the conventional scanning system to such an extent that the sensitivity of the two systems were equal. This, however, cannot be done because even with a perfect amplifying system the statistical fluctuations in the original photo-current are amplified just as much as the signal is amplified. Because of this there is a definite limit to the sensitivity of this type of system imposed by the original shot noise in the photo-current. In the case of the Iconoscope there is a similar limit, but because the charge representing each picture element is so much greater than in the non-storage case

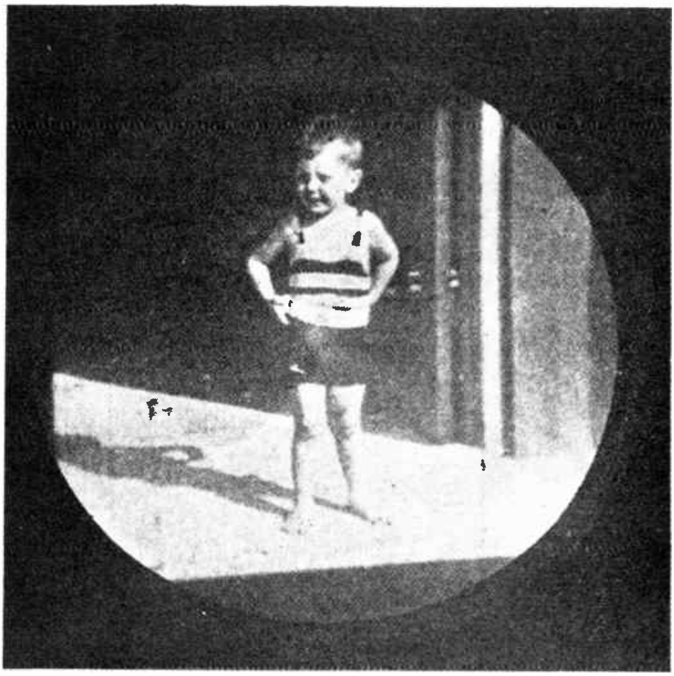

Figure 17

the ideal sensitivity is very much greater. Actually, in the type of Iconoscope described in this paper, the limit of the sensitivity is not set by the statistical fluctuations of the stored charge, but by the thermal noise in the coupling resistor to the amplifier. A quantitative comparison of the limiting sensitivity of the Iconoscope used at present, taking into account its inefficiency and imperfections, shows that it is able to operate at one-tenth the light required by a *perfect* non-storage system. This, of course, includes the use of an electron multiplier and applies to electrically scanned as well as mechanically scanned non-storage systems.

The resolution of an Iconoscope may be limited either by the size of the photoelectric elements or by the size of the scanning beam. The size of the silver globules in the Iconoscope described is many times smaller than a picture element so that many hundreds of them act together under the scanning spot. The resolution is limited, therefore, by the spot size. At present, the resolution adopted is about 360 lines, but when necessary the beam size can be reduced and the resolution made much higher.

Figure 18

The actual response of an Iconoscope under various conditions of illumination is shown in Fig. 12. The output is measured in millivolts across a 10,000-ohm coupling resistance and the light input measured in lumens per square centimeter on the mosaic. The curve showing the greatest response represents the signal output from a small illuminated area when the remainder of the mosaic is in darkness. The other curves of the family show the response from the same area when the mosaic is illuminated with a uniform background of light. The response is not linear but falls off as the illumination is increased until it reaches a saturation value. The saturated voltage output is nearly constant

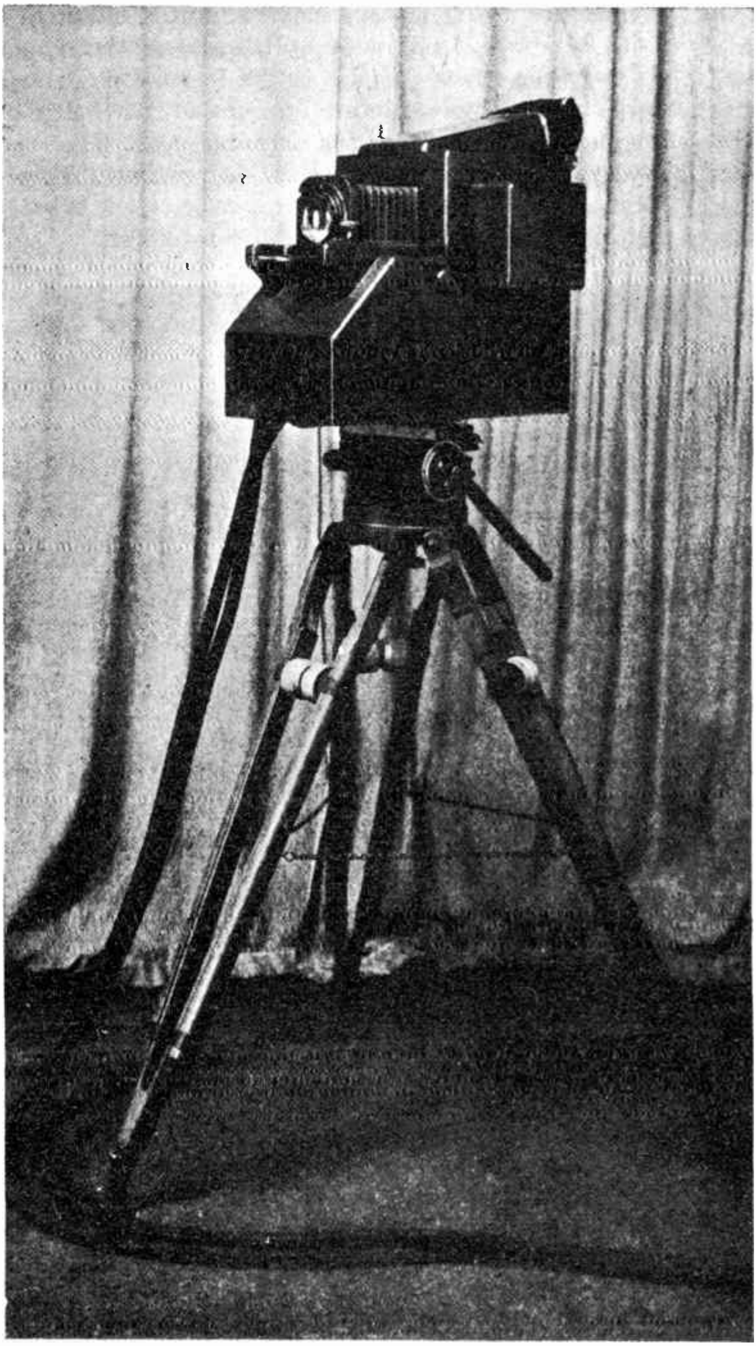

Figure 19

for tubes of a given design, but the slope of the response curve may vary from tube to tube depending upon treatment and is a measure of the sensitivity.

The decrease in sensitivity with illumination is not wholly disadvantageous in that it permits the transmission of a wider range of contrast over a given electrical system than would otherwise be possible. In a sense, this is similar to the compressor-expander systems used in sound recording.

In spite of the complicated manner of its operation and the factors mentioned reducing its efficiency, the Iconoscope is an extremely sensitive and stable device for obtaining television transmission. Excellent and consistent results are obtained under widely varying conditions of operation. The practical lower limit to light which can be used to transmit a picture is set by the "noise" in the picture amplifier. Measurements have been made to determine the illumination necessary for satisfactory operation. With an F/2.7 lens to focus the image on the mosaic, an average surface brilliancy of from 30 to 50 candles/sq. ft. on the object viewed gives completely satisfactory transmission. A recognizable image can be obtained from a good Iconoscope with 8 candles/sq. ft. using an F/16 lens, that is, with 1/150 the illumination mentioned above.

For comparison, the illumination of some scenes commonly met with is given in the following table:

Scene	*Location*	*Date*	*Time*	*Weather*	*Brightness*
Beach	Atlantic City	August	2:00 P.M.	Hazy	500
Boardwalk	Atlantic City	August	2:00 P.M.	Hazy	275
Street	Philadelphia	August	2:30 P.M.	Clear	200
Times Square	New York City	November	1:30 P.M.	Rain	40
Street Parade	East Orange	November	10:30 A.M.	Rain	40 to 60

It is evident that perfectly satisfactory outdoor pickup may be obtained under almost all average conditions of light.

The device used to reproduce the television picture is also electron operated. This tube, which has been named the "Kinescope," is similar to a cathode-ray oscilloscope in many respects. It consists of an electron gun for defining and controlling a cathode-ray beam and a fluorescent screen which becomes luminous under bombardment from the electron gun. A diagram of a typical Kinescope is shown in Fig. 13.

The cathode-ray beam is made to sweep across the fluorescent screen in synchronism with the scanning beam in the Iconoscope which is transmitting the picture. Furthermore, the current in the Kinescope cathode-ray beam is controlled by the signal im-

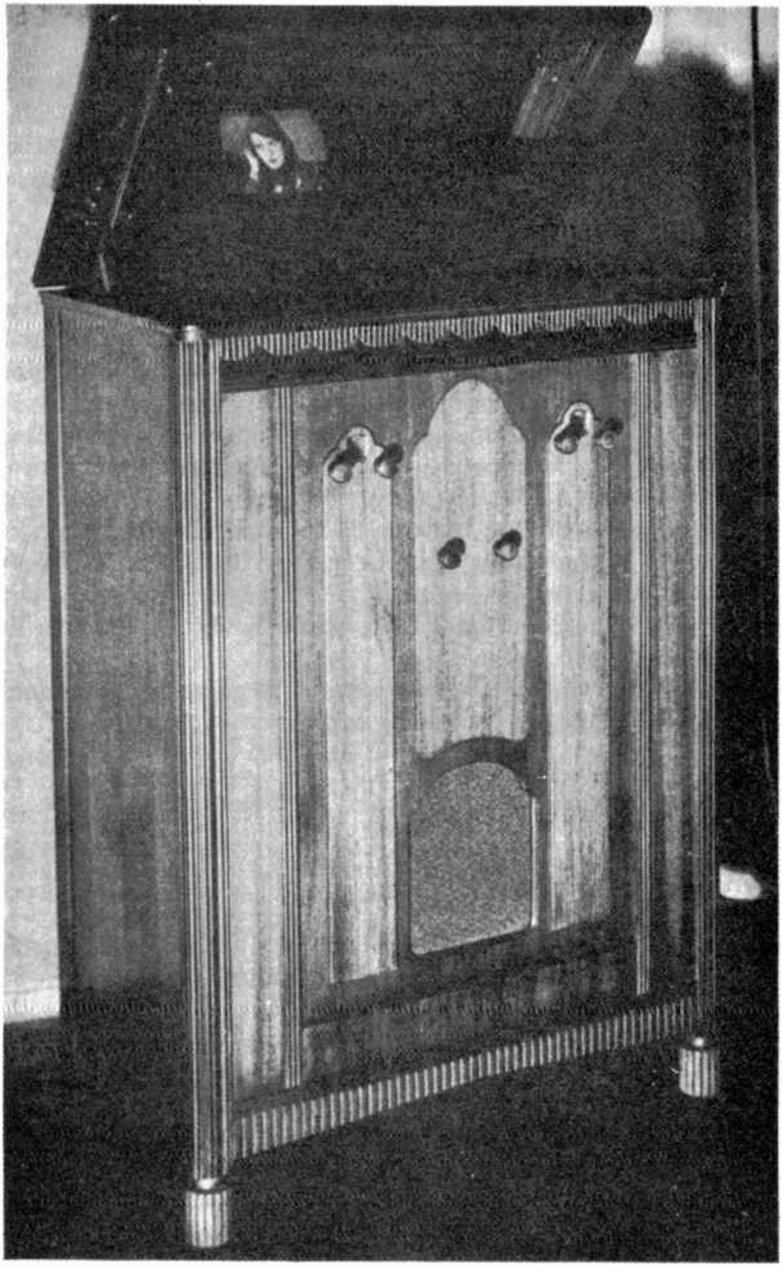

Figure 20

pulses generated at the Iconoscope. This control acts in such a way that the impulse corresponding to a bright area on the Iconoscope causes an increase in current, while a dark region causes a decrease. There will, therefore, be an exact correspondence both in position and intensity between the fluorescent illumination on the Kinescope screen and the light on the mosaic in the Iconoscope. A picture projected on the Iconoscope will therefore be reproduced by the Kinescope.

The electron gun in the receiving tube is similar in principle to that in the Iconoscope, but is made to handle larger currents and to operate at higher voltages. Furthermore, since the picture is reproduced by modulating the beam current, the control grid is a much more critical item. The control grid characteristic is determined by a number of factors such as the grid aperture, the spacing and geometry of the cathode, the first anode, etc. Fig. 14 shows a typical control characteristic for a Kinescope gun.

The fluorescent screen is made by coating the flat portion of the glass bulb with a synthetic zinc orthosilicate, very similar to natural Willemite. The synthetic material has high luminous efficiency, the light output at a given voltage being proportional to the current striking it. At 6000 volts the material gives nearly 3 candles per watt. The efficiency of light production varies somewhat with voltage used, but at higher beam velocities is nearly constant. This can be seen from the general relation between candlepower P, current intensity I, and applied voltage V, which is given by the equation:

$$P = AI\ (V\text{-}V_o)$$

A is a constant depending upon the phosphor and V_o the extrapolated minimum exciting voltage, which proves to be in the neighborhood of 1000 volts.

In addition to its high luminous efficiency, this material does not burn or disintegrate under bombardment with electrons. The phosphorescent properties of a fluorescent material are an important consideration. An ideal substance for television work should emit a constant amount of light for one entire picture frame and drop to zero at the end of this period. If the phosphorescent time is too long, the moving portions of a picture will leave a "trail." For example, the path of a moving ball will be marked with a comet-like tail. On the other hand, if the decay time is too short, flicker becomes noticeable. The phosphorescent decay curve for zinc ortho-silicate is shown in Fig. 15.

Fig. 16 shows a photograph of a "Kinescope" with a 9-inch viewing screen. This is only one of a number of sizes, both larger and smaller, and designs possible for the "Kinescope."

Between the transmitting Iconoscope and the reproducing Kinescope there is a chain of electrical equipment involving the picture amplifier, transmitter, radio receiver and synchronizing system. This field is much too large to cover in this paper and has been treated in detail elsewhere.[1]

In closing, it might be well to illustrate the performance of the systems with some photographs of televised pictures as they appear on the screen of the Kinescope. These are shown in Figs. 17 and 18. The appearance of a typical studio pickup camera using the Iconoscope is shown in Fig. 19, while that of a console type television receiver can be seen in Fig. 20.

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[1] Description of an Experimental Television Receiver, *Proc. I.R.E.*, Vol. 21, No. 12, December, 1933.
An Experimental Television System, Part II—Transmitter, *Proc. I.R.E.*, Vol. 22, No. 11, November, 1934.
An Experimental Television System, Part III—Receivers, *Proc. I.R.E.*, Vol. 22, No. 11, November, 1934.
An Experimental Television System, Part IV—Radio Relay Link for Television Signals, *Proc. I.R.E.*, Vol. 22, No. 11, November, 1934.

TELEVISION STUDIO TECHNIC*†

BY

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Summary—The studio operating technic as practiced in the NBC television studios today is discussed and comparisons are made, where possible, to motion picture technic. Preliminary investigations conducted to derive a television operating technic revealed that both the theater and the motion picture could contribute certain practices.

The problems of lighting, scenic design, background projection, and make-up are discussed, with special emphasis on the difficulties and differences that make television studio practice unique.

An explanation is given of the functioning of a special circuit used in television sound pick-up to aid in the creation of the illusion of close-up and long-shot sound perspective without impracticable amount of microphone movement. The paper concludes with a typical television production routine showing the coördination and timing of personnel and equipment required in producing a television program.

IF ONE were forced to name the first requirement of television operating technic and found himself limited to a single word, that word would undoubtedly be "timing." Accurate timing of devices and split-second movements of cameras are the essentials of television operation. Personnel must function with rigid coördination. Mistakes are costly—they must not happen—there are no second chances.

Why such speed and coördination? Television catches action at the instant of its occurrence. Television does not allow us to shoot one scene today and another tomorrow, to view rushes or resort to the cutting room for editing. Everything must be done as a unit, correct and exact at the time of the "takes"—otherwise, there is no television show.

Now, to discuss some preliminary investigations conducted before production was attempted, and to describe the equipment and technic used in meeting these production requirements. Technical details are deliberately omitted. Wherever possible, we shall compare phases of television operation with their counterparts in motion picture production.

For so new a medium as television it is, of course, an impossibility to present a complete and permanently valid exposition. Television technic and apparatus constantly advance. Some technic now current may be outmoded in a day or a month. We have only to recall the early days of motion picture production, when slow-speed film and

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inferior lenses were a constant limitation. So, with television, it is already possible to envision more sensitive pick-up tubes that will permit the use of smaller lenses of much shorter focal length, thus eliminating many of today's operating difficulties.

Production Technic Investigations

In May, 1935, the Radio Corporation of American released television from its research laboratories for actual field and studio tests. Long before the first program was produced in the middle of 1936, plans were laid, based on extensive research into the established entertainment fields, for the purpose of determining in advance what technics might be adaptable to the new medium of television. From the stage came the formula of continuity of action, an inherent basic requirement of television. This meant memorized lines and long rehearsals. Prompting could not be considered, for, as you know, the sensitive microphone which is as much present in television as it is in sound motion picture production, does not discriminate between dialog and prompting.

From the motion picture studio came many ideas and technics. If television is a combination of pictures with sound, and it is, no matter what viewpoint is taken, the result spells in part and for many types of programs, a motion picture technic at the production end. However, enough has already been said about the peculiarities of television presentation to justify saying that the movie technics do not supply the final answer. There remained the major problem of preserving program continuity without losing too much of motion picture production's flexibility. Our present technic allows no time for adjustments or retakes. Any mistake immediately becomes the property of the audience. The result of the entire investigation led to what we think is at least a partial answer to the problem. This technic, we hope, will assist considerably in bringing television out of the experimental laboratory and into the field of home education and entertainment.

General Layout of Facilities

In order to present a clearer view of our problems, we shall give a brief description of our operating plant. The present television installation at the National Broadcasting Company's headquarters in the RCA Building, New York, N. Y., consists of three studios, a technical laboratory, machine and carpenter shops, and a scenic paint shop. Our transmitter is located on the 85th floor of the Empire State Building. The antenna system for both sight and sound is about 1300 feet above the street level. Both the picture and sound signals are relayed from the Radio City Studios to the video and sound transmitters either by coaxial cable or over a special radio link transmitter.

One of the studios is devoted exclusively to televising motion picture film, another to programs involving live talent, and the third for special effects. It is the operation of the live-talent studio with which we are concerned in this paper.

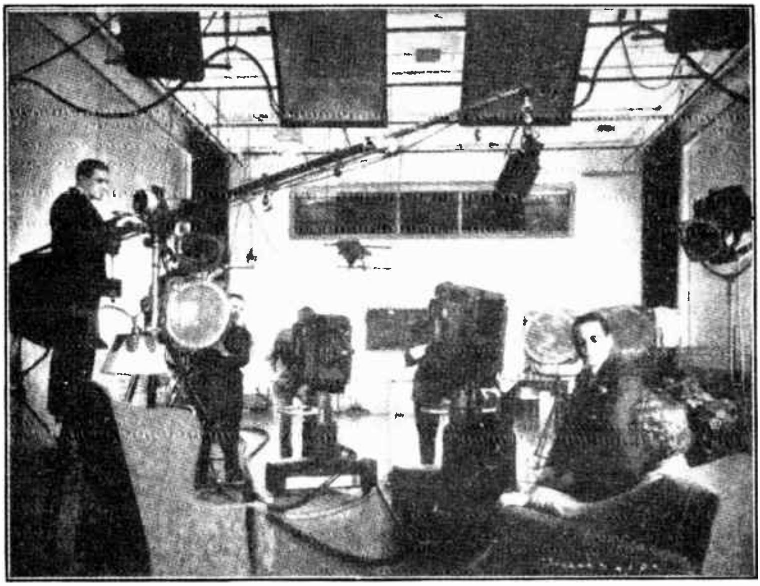

Fig. 1*(a)*—General layout of live-talent studio; control room at upper rear.

Description of Live-Talent Studio

Figure 1(*a*) shows the general layout of the live-talent studio. The studio is 30 feet wide, 50 feet long, and 18 feet high. Such a size should not be considered a recommendation as to the desired size and proportions of a television studio. The studio was formerly a regular radio broadcasting studio, not especially designed for television. To anyone familiar with the large sound stages on the motion picture lots, this size may seem small (Figure 1(*b*)). Yet, in spite of our limited space, some involved multi-set pick-ups have been successfully achieved by careful planning. Sets, or scenes, are usually placed at one end of the studio. Control facilities are located at the opposite end in an elevated booth, affording full view of the studio for the control room staff. Any small sets supplementing the main set are placed along the side walls as near the main set as possible, and in such position as to minimize camera movement. At all times, we reserve as much of the floor space as possible for camera operations and such floor lights as are absolutely essential. At the base of the walls and also on the ceiling are scattered numerous light-power outlets to minimize the length of lighting cables. At the rear of the studio is a permanent projection room for background projection.

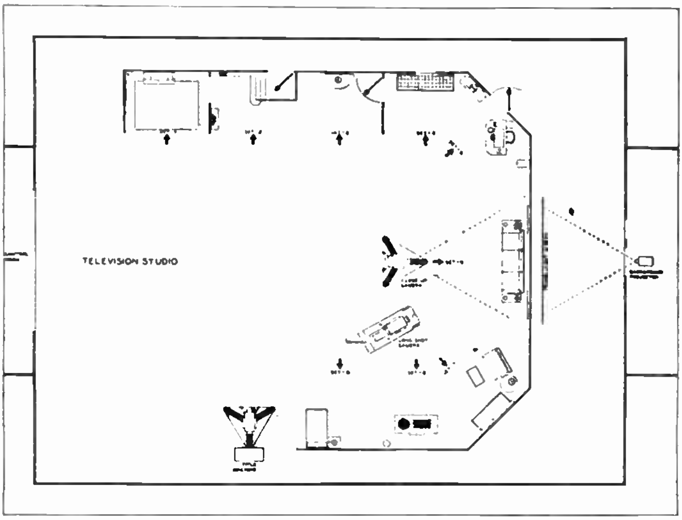

Fig. 1(*b*)—Television studio floor plan.

Camera Equipment

The studio is at present fitted for three cameras. To each camera is connected a cable. This cable is about two inches in diameter and fifty feet long; it contains 32 conductors including the well known coaxial cable over which the video signal is transmitted to the camera's associated equipment in the control room. The remainder of the conductors carry the necessary scanning voltages and current supplies for the camera amplifiers, interphone system, signal lights, etc. From this description, it is apparent that adding another camera in a television studio involves a much greater problem than that of moving an extra camera into a motion picture studio. In television, it is necessary to add an extra rack of equipment in the control room for each additional camera.

Movement of Cameras

One camera, usually the long-shot camera using a short-focal length lens, is mounted on a regular motion picture type dolly to insure stable movements. The handling of the dolly is done by a technician assisting the camera operator. It is impracticable to lay tracks for dolly shots as is often the motion picture practice, because usually each camera must be moved frequently in all directions during the televising of a studio show. Naturally, dolly tracks would limit such movement. The other television cameras utilize a specially de-

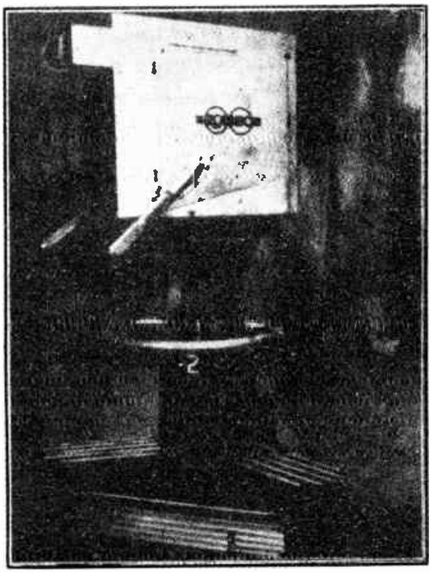

Fig. 2—Studio camera.

signed mobile pedestal (Figure 2). Cameras mounted on these pedestals are very flexible and may be moved in and out of position by the camera operators themselves. Built into the pedestals are motors which elevate or lower the camera; this action is controlled with push-buttons by the camera operators. A panning head, similar to those used for motion picture cameras, is also a part of the pedestal. It is perhaps needless to stress here that one of the strict requirements of a television camera is that it must be silent in operation. In the electronic camera proper there are no moving parts other than those used for focusing adjustments; hence, it is a negligible source of noise. When camera pedestals were first used they were the source of both mechanical noise and electrical disturbance when the camera-elevating motor was in use. Since then this problem has been overcome, and it can be stated that the entire camera unit is now free of objectionable mechanical noise or electrical surges.

Lens Complement

Each camera is equipped with an assembly of two identical lenses displaced 6 inches vertically. The upper lens focuses the image of the scene on a ground-glass which is viewed by the camera operator. The lower lens focuses the image on the "mosaic," the Iconoscope's light-sensitive plate. This plate has for its movie counterpart the film in a motion picture camera. The lens housings are demountable and interchangeable. Lenses with focal lengths from 6½ to 18 inches are used at present. Lenses of shorter focal length or wider angle of pick-up can not be used since the distance between the mosaic and the glass envelope of the Iconoscope is approximately 6 inches. Lens changes can not be effected as fast as on a motion picture camera, since a turret arrangement for the lenses is mechanically impracticable at present. However, it is probably safe to say that future advances in camera and Iconoscope design will incorporate some type of lens turret. Ordinarily, one camera utilizes a 6½-inch focal length lens with a 36-degree angle, for long shots, while the others use lenses of longer focal lengths for close-up shots. Due to its large aperture, the optical system used at present has considerably less depth of focus than those used in motion pictures, making it essential for camera operators to follow focus continuously and with the greatest care. This limitation will probably be of short duration, since more sensitive Iconoscopes will permit the use of optical systems of far greater depths of focus.

It is desirable here to point out a difference in focusing technic between motion picture cameras and television cameras. "Follow-focus" in motion pictures occurs practically only in making dolly shots. For all fixed shots, the lens focus is set, the depth of focus being sufficient to carry the action. Also, it is the duty of the assistant cameraman to do the focusing. This relieves the cameraman of that responsibility and allows him to concentrate on composition, action, and lighting. In television, the camera operator must do the focusing for fixed shots and dolly shots alike. This added operation, at times, is quite fatiguing.

Vertical parallax between the view finder lens and the Iconoscope lens is compensated for by a specially designed framing device at the ground-glass that works automatically in conjunction with the lens-focusing control. It may be of interest to note here that at first the television camera had no framing device. This meant that images, in addition to being inverted as they are in an ordinary view-finder, were also out of frame. The camera operator had to use his judgment in correcting the parallax. With this new framing device, the operator now knows exactly the composition of the picture being focused on the mosaic in his camera. The framing device can be quickly adjusted to accommodate any lens between 6½ and 18 inches focal length.

Because of the fact that several cameras are often trained on the same scene from various angles, and because all cameras are silent in operation, performers must be informed sometimes—such as when they are speaking directly to the television audience—which camera is active at the moment. Two large green bull's-eye signal-lamps

Fig. 3—Typical television set.

mounted below the lens assembly are lighted when the particular camera is switched "on the air."

Set Lighting

There are two outstanding differences between television lighting and motion picture lighting. A much greater amount of key light is required in television than in motion pictures. Also, a television set must be lighted in such a way that all the camera angles are anticipated and properly lighted at one time. Floor light is held to a minimum to conserve space in assuring maximum flexibility and speed of camera movements. Great care must also be taken to shield stray light from all camera lenses. This task is not always easy, since, during a half-hour performance, each camera may make as many as twenty different shots. Just as excessive leak-light striking the lens will ruin motion picture film, it has a definitely injurious effect upon the photosensitive mosaic and upon the electrical characteristics of the Iconoscope. A direct beam of high-intensity light may temporarily paralyze a tube, thus rendering it useless for the moment.

Sets

Television sets (Figure 3) are usually painted in shades of gray. Since television reproduction is in black and white, color in sets is

Fig. 4—Background projection window shot.

relatively unimportant. Chalky whites are generally avoided because it is not always possible to keep "hot lights" from these highly reflective surfaces which cause a "bloom" in the picture. This, in turn, limits the contrast range of the system. Due to the fact that the resolution of the all-electronic system is quite high, television sets must be rendered in considerable detail, much more, in fact, than for a corresponding stage production. As in motion picture production, general construction must be as real and genuine as possible; a marked difference, for instance, can be detected between a painted door and a real door. On the legitimate stage, a canvas door may be painted with fixed highlights; that is, a fixed perspective, because the lighting remains practically constant, and the viewing angle is approximately the same from any point in the audience. But, in television the perspective changes from one camera shot to another. Painted perspectives would therefore be out of harmony with a realistic appearance. This is also true in motion picture work. Sets must also be designed so that they can be struck quickly with a minimum effort and noise because it is often necessary to change scenes in one part of a studio while the show is going on in another part. At present, we find it desirable to construct television sets in portable and lightweight sections without sacrificing sturdiness.

Background Projection

The problems of background projection in television differ somewhat from those encountered in motion pictures. More light is necessary because of the proportionately greater incident light used on the sets proper (Figure 4).

Considering the center of a rear-screen projection as zero angle, we must make it possible to make television shots within angles of at least 20 degrees on either side of zero without appreciable loss of picture brightness. This requirement calls for the use of a special screen having a broader viewing angle than those used in making motion picture process shots. Also, in motion pictures, the size of the picture on the screen can be varied to the proper relation to the foreground for long shots or close-ups. For television, the background picture size can not be changed once the program starts. Our background subject matter must also be sharp in detail and high in contrast for good results.

At present, only glass slides are used. A self-circulating water-cell is used to absorb some of the radiant heat from the high-intensity arc. Also both sides of the slide are air-cooled. These precautions permit the use of slides for approximately 30-minute periods without damage.

Make-Up

This may be a suitable time to correct some erroneous impressions concerning the type of make-up used in television. It has never been necessary to use gruesome make-up for the modern all-electronic-RCA television system. At present, No. 26 panchromatic base, similar to that used for panchromatic film, and dark red lipstick is being used satisfactorily. From the very beginning, we have made tests to determine the proper color and shades of make-up, keeping in mind that a color closely approximating the pigmentation of the human skin is most desirable from the actor's psychological standpoint.

The Control Room

Now, a few words about the operations in the studio control room during a televised production (Figure 5). All camera operators in the studio wear head-phones through which they receive instructions from the control room. Directions are relayed over this circuit by the video engineer or the production director. Here the televised images are observed on special Kinescope monitors and necessary electrical adjustments are made. Alongside each of these monitoring

Fig. 5—The television control room. Note the two Kinescope monitors in the upper left corner.

Kinescopes is a cathode-ray oscilloscope which shows the electrical equivalent of the actual picture. Two monitors are provided in order that one may be reserved for the picture that is actually on the air, while the other shows the succeeding shot as picked up by a second or third camera. This enables the video engineer to make any necessary electrical adjustments before a picture goes on the air.

Seated immediately to the left of the video engineer is the production director whose responsibility corresponds to that of the director of a motion picture. He selects the shots and gives necessary cues to the video engineer for switching any of the cameras into the outgoing channel. The production director has, of course, previously rehearsed the performance and set camera routines in conjunction with the camera operators and the engineering staff. The camera operator has no control to switch his camera on the air. All camera switches, which are instantaneous, are made by electrical relays controlled by buttons in the control room. At present, the video engineer's counterpart in motion picture work is the editor and the film processing laboratory.

To the left of the production director sits the audio control engineer whose responsibility is entirely separate from that of the video engineer. He also is in a position to view the monitor, and may communicate by telephone with the engineer on the microphone boom. The audio engineer is responsible for sound effects, some of which are dubbed in from records. His job is somewhat similar to that of the head sound engineer on a motion picture production. Thus, we have the control room staff—three men who have final responsibility for the success of the completed show.

An assistant production man is also required on the studio floor. Wearing headphones on a long extension cord, he is able to move to any part of the studio while still maintaining contact with the production director in the control room during a performance on the air. Actors require starting cues, titles require proper timing, and properties and even an occasional piece of scenery must be moved. The assistant director supervises these operations and sees that the instructions of the production director are properly carried out.

Members of the studio personnel also to be mentioned include lighting technicians, the property man, and scene shifters, whose responsibilities parallel those of their motion picture counterparts. Specially trained men are also needed for operating title machines. In the future all titling will undoubtedly be done in a separate studio inasmuch as operating space in a television studio is at a premium. Today, however, title machines do operate in the studio and require the utmost care in handling. Types of titles used include dissolves and wipes similar to those used in moving pictures.

Sound Reproduction

As in motion picture work, a microphone boom is used in television production, and is operated in a similar way. Perspective in motion picture sound is accomplished by keeping the microphone, during a long shot, just out of the picture and moving it down closer to the action as the camera moves in for a close-up, thus simulating a natural change in perspective. In television this is not always possible because there are always three cameras to consider. This same condition prevailed in the early days of motion pictures when it was thought desirable to take a complete scene, shooting both long-shot and close-up cameras, at one time. In the television studio at least one camera is always set for a long shot while the others are in position for closer shots. If the microphone is placed in such a position as to afford a "natural" perspective for close-ups, the succeeding switch to a long shot would reveal the microphone in the shot. You in motion pictures can order a retake; in television broadcasting we can not rectify the mistake. It is quite obvious, therefore, that the man on the boom can not lower his microphone to the "natural" position for each camera shot. We therefore place the microphone in a position just out of range of the long shot. In order to accomplish some sense of perspective between long and close-up shots, a variable equalizer that drops the high and low ends of the spectrum is automatically cut into the audio circuits when the long-shot camera is on the air. In this operation, sufficient change in quality and level is introduced to aid the illusion of long-shot sound perspective. Of course, when a close-up camera is switched in, the audio returns to the close-up perspective quality once more. This may be called remote control sound perspective.

Special sound effects, music, etc., from the studio picked up from recordings are mixed in the control room. In motion pictures, some of the effects and most of the music are dubbed in after the actual shooting of the scene.

The general acoustical problems in a television studio are similar to those in a motion picture sound-stage. Walls and ceiling should be designed for maximum absorption to permit faithful exterior speech pick-up. A stage or studio must be designed to enable presentation of an exterior or an interior scene. With the studio designed for maximum absorption, illusions of exterior sound characteristics can be created. For interiors, the hard surfaces of the sets and props offer sufficiently reflective surfaces to create the indoor effect.

Typical Production Routine

After the foregoing discussion of the equipment and personnel, it may be interesting to follow an actual production from the beginning of rehearsal to its final presentation. For this example, assume that we are to produce a playlet (Figure 6). When the scenery has been erected, the first rehearsals begin without the use of cameras or lights. Besides familiarizing the actors with their lines, the rehearsals afford the production director and the head camera operator an opportunity to map out the action of the play. All action, including camera shots.

cues, and timing, is noted on a master script which thereafter becomes the "bible" of the production. Timing is very important because of the necessity of having a particular act time in with the other acts or film subject.

After several hours of rehearsing, the first equipment rehearsal is called. Cameras are checked electrically and mechanically. Focus controls and framing devices are lined up so that correct focus on the ground-glass is also correct focus on the mosaic plate. This completed, the cameras are ready for rehearsal. With the scene properly lighted, the camera operators begin working out movements to pick up the

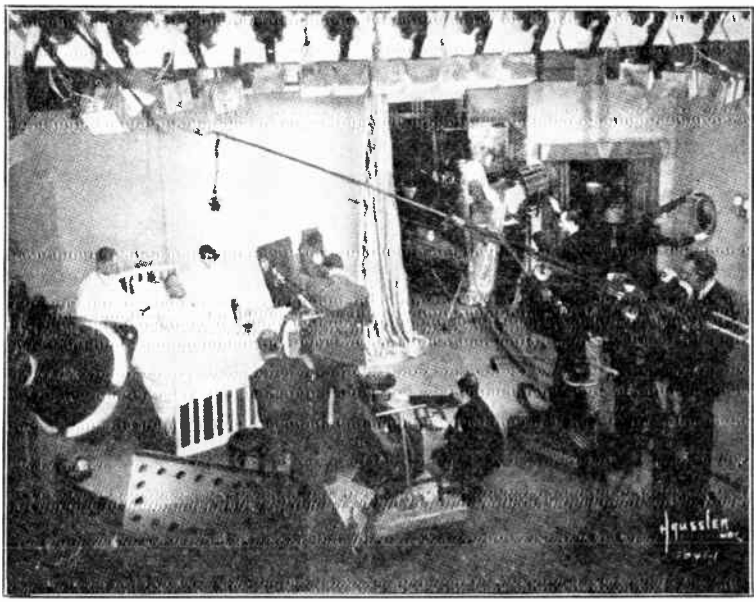

Fig. 6—*(Left)* Scene on the air. *(Right)* setting up for next scene.

desired shots in the proper sequence. The production director instructs the staff and personnel from the control room, speaking over a public-address system. Each shot is worked out and its camera location marked on the floor. At times, the actors may unconsciously depart slightly from the rehearsed routine during an actual show; the camera operator must be prepared and alert to make the best of the situation regardless of all previous floor markings. Continuity is so planned that while one camera is taking the action, another camera is moving to a new location and composing a new shot to be switched on at the proper time. This frees the first camera, which can now move to a third location, and so on. Sometimes during a twenty-minute performance each camera may take twenty different shots. Of course, besides different floor locations, the height and angle of the cameras must be varied to comply with good composition. During rehearsals, timing must frequently be revised to allow for the actual camera movements.

Finally, a dress rehearsal is scheduled. The complete program is televised, including any film subjects or slides that may be needed to complete the program. Frequently the program will begin with a short film leader, followed immediately by a newsreel or a short subject, the film portion of the program coming from the film-televising studio. While the film is running, the live-talent studio is continuously warned as to the time remaining before it must take over the program. Once the studio program goes on the air the production director is no longer able to use the public address system to communicate with the personnel in the studio. Instead, he uses a telephone circuit to his assistant in the studio, and, through the video engineer, communicates by phone with the camera operators.

Another standby warning is usually given when there is one minute to go. Then, as the cue to begin comes, the green light on the title camera is lighted. From this point, continuity must be rigidly preserved. As titles move from one to another, appropriate music is cued in and actors are sent to their opening positions.

With the completion of titles, the image is faded out electrically and cameras are switched to the opening shot. Performers begin their action on a silent cue from the assistant director, who is instructed from the control room. During this first scene, the camera previously picking up titles moves quickly into position to shoot a second view of the action. Again cameras are switched, permitting the first to move to a new position; and so the action proceeds. If the play has several scenes, the concluding shot of the first scene is taken by one camera while others line up on the new scene and wait for the switch. Frequently, there are outdoor scenes. These are filmed during the first stages of rehearsal for transmission from the film studio at the proper time during the performance. The switch to film is handled exactly as another camera switch, except that the switch is to the film studio instead of to one of the studio cameras. The projectionist must be warned in advance to have his projector up to speed and "on the air" at the proper instant to preserve the production continuity. This requires very critical timing, as you can well appreciate. When the film is completed the studio cameras again take over the next interior scene.

Upon completion of the studio portion of the program, one camera lines up on the final studio title, which usually returns the program to the film studio for a concluding film subject.

Since the first program on July 7, 1936, many television programs have been produced. Each has been a serious attempt at something new. Although much has been accomplished, there remain a vast number of unknowns to be answered before it can be said that television's potentialities have been even partially realized. Today, as this paper has indicated, television bears many points of similarity to motion pictures. As a matter of fact, it is likely that television would be somewhat handicapped if it were unable to borrow heavily from a motion picture production technic that has been built up by capable minds and at great expense over a period of many years. Infant television is indeed fortunate to have such a wealth of information at its disposal. Possibly continued experimentation will lead us toward a new technic distinctive of television. During its early years, however, television must borrow from all in creating for itself a book of rules. The first chapter of that book is scarcely written.

APPLICATION OF MOTION-PICTURE FILM TO TELEVISION*†

By

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Summary—*Motion-picture film will form an important source of programs for television broadcasting. Film projectors for this use are required to meet a number of conditions peculiar to television. Methods for projecting and utilizing motion-picture film are outlined. A specific film projector and associated television channel are described in some detail.*

In establishing a technique for producing films most suitable for television, equipment is needed to interpret the final results. Apparatus that will be used by broadcasting stations is described. A simpler system has been designed that may be useful for the specialized service of gaging the merit of films for television. This is described and its operation indicated.

Some very preliminary observations are included on the characteristics of films that have given good results in experimental work and in field tests.

THE production and utilization of motion-picture film for television programs introduces many new problems. It is the purpose of this paper to review these problems and to describe methods and apparatus for the use of film in television.

GENERAL DISCUSSION OF UTILIZATION METHODS

It is desirable first to review the general characteristics of two electronic television pickup systems, which are known to give practical results. In both systems the scene to be transmitted is projected upon a photo-emissive area or mosaic. The resulting "electrical image" is methodically explored by electronic means, one narrow strip or line at a time, in a process called scanning. The result of this scanning process is an electrical signal which varies in accordance with the scene brightness along the scanning lines. The information residing in this signal is used at the receiver to reconstruct the image—one element at a time—in a similar synchronized scanning process.

In one pickup system, exemplified by equipment using the Farnsworth dissector tube, only the light falling upon an element of the photo-emissive area at the instant that element is being scanned is effective in producing the signal. The other pickup system, exemplified by equipment using the Iconoscope, makes use of the principle of storage, whereby, when a particular photo-emissive element is scanned the light which has fallen upon that element since it was last scanned is effective in producing the signal.

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The characteristics of these pickup tubes determine the manner in which film can be used to provide television programs. In the system using the dissector tube which has no storage, for every instant that signal is transmitted, the film projector must supply a light image to the elemental area being scanned, though not necessarily from the entire frame. In the Iconoscope system utilizing storage, a charge image may be built up by a very brief projection of the image upon the photo-emissive mosaic, which is then scanned by an electron beam

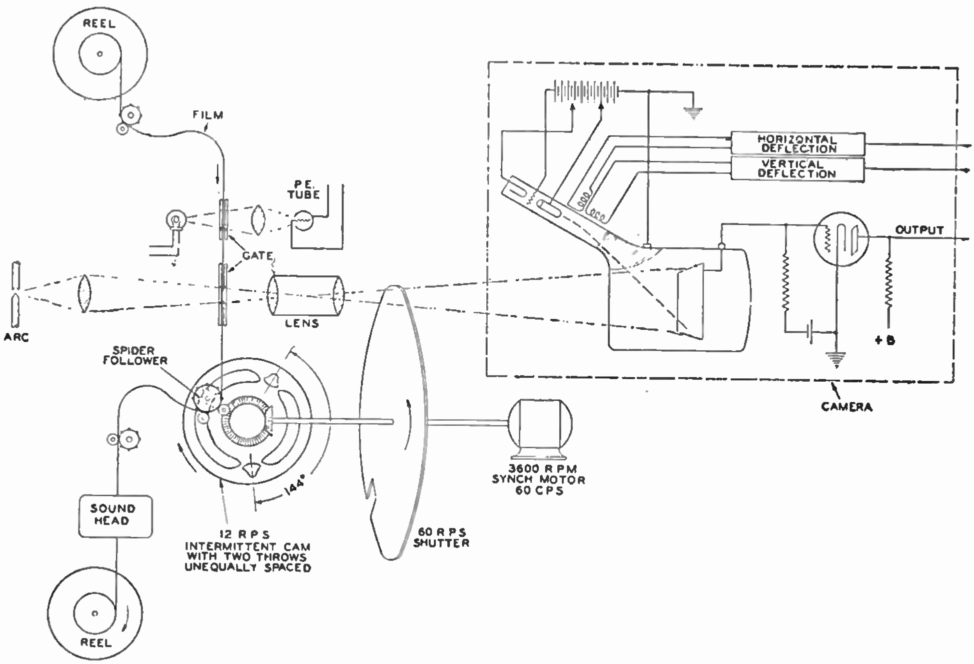

Fig. 1—Schematic of film projector for Iconoscope camera.

while the mosaic is dark to produce the signal. The film pull-down occurs during the relatively long interval while the mosaic is being scanned. The detailed discussion to follow will be based on the system utilizing the Iconoscope.

Discussion of Film Transmission System Utilizing an Iconoscope

Figure 1 shows schematically an Iconoscope camera and a special projector adapted to project standard 24-frame-per-second film upon the Iconoscope mosaic in such way as to generate television signals according to the Radio Manufacturers Association standards; namely, at 30 frames per second and 60 fields per second, interlaced. [1] The projector must flash a still picture upon the mosaic every 1/60 second with each flash lasting less than 1/600 second. Since the film must run at a mean speed of 24 frames per second for proper reproduction of sound and motion, it is evident that each frame must be projected more than once to provide the required sixty flashes per second. Since sixty divided by 24 is 2½, it would seem logical that each frame should be projected two and one-half times. This is impracticable, but a very satisfactory method is to project alternate frames of film two and three times each, respectively; for example, the even frames twice and the odd frames three times. Figure 2 shows the various steps of projection and scanning in proper relative time on a horizontal time scale. Since the light flashes are very brief, a relatively long (approximately 1/67 second) interval is available between flashes for

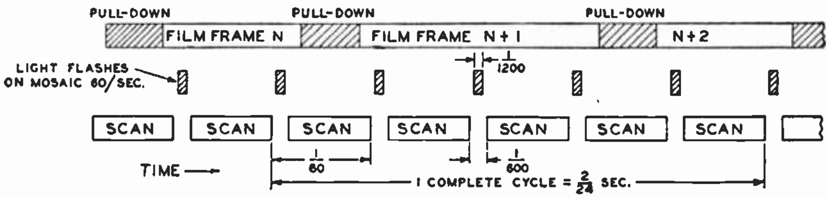

Fig. 2—Preferred sequence of events in film transmission by Iconoscope.

the film pull-down. However, if the full time available is used, the alternate pull-downs must occur at non-uniform intervals of 2/60 and 3/60 seconds, respectively. Note from this figure that the scanning or transmission times occur *between* adjacent light flashes so that the television picture signal is actually produced and transmitted during periods when no optical image is present on the mosaic. However, during these periods an electrical image is present in the form of bound electrostatic charges on the tiny photo-sensitized silver globules comprising the mosaic. It is the act of neutralizing or rather equalizing these charges by the electrons of the scanning beam which causes the useful signal current to flow from the conducting back coating of the mosaic plate.

Referring again to Figure 1, the film is drawn through an illuminated gate by an intermittent sprocket which is driven by an intermittent cam and spider-follower of the early Powers type. The

[1] G. L. Beers, E. W. Engstrom, and I. G. Maloff, "Some Television Problems from the Motion Picture Standpoint", *Jour. Soc. Mot. Pic. Eng.*, Vol. XXXII, pp. 121-136, February, 1939.

3600-r.p.m. special synchronous motor drives the cam at 12 revolutions per second through a suitable gear, thus pulling the film down 24 times per second, since the cam has two "throws" instead of the customary one "throw." In order to pull the film at unequal intervals as required, the "throws" are located 144 degrees and 216 degrees apart, respectively. The film picture in the gate is projected upon the small photoemissive mosaic of the Iconoscope by a standard projection lens. The light is chopped 60 times per second by a large rotating shutter, located near the lens. The shutter is accurately timed relative to the intermittent cam so that the film is always stationary when the light flashes occur.

The generator of synchronizing signals for the television deflecting system is synchronously controlled by the same 60-cycle power supply which drives the projector synchronous motor. The phase of this signal generator is adjustable so that the operator can make the short duration light flashes fall safely within the 1/600-second intervals between the vertical scanning periods with some tolerance on each side for slight phase displacements such as are caused by small changes in the mechanical load on the projector or by voltage variations. This adjustment is very important, as any abrupt change in the illumination of the mosaic during the picture signal transmission time produces a spurious light streak across the received picture.

An ordinary 3600-r.p.m. synchronous motor has two identical pole structures which can assume either polarity and hence such a motor can lock into synchronism in either of two phase positions, depending fortuitously upon starting conditions. Two such lock-in positions are one-half of a cycle of the power-supply frequency apart in time, which for a 60-cycle power system is 1/120 second. Inspection of the diagram of Figure 2 shows that displacing the light flashes 1/120 second with respect to the scanning periods would cause them to occur during instead of between the scanning periods. The abrupt change in mosaic lighting caused by a flash during the scanning period would produce a serious streak across the middle of the picture as mentioned above To prevent the frequent locking-in of the motor in the wrong position, a special synchronous motor is used which includes an additional d-c winding for fixing the polarity of the poles and thus determining the lock-in position with respect to the a-c power supply.

The sound head used is standard, since the mean speed of the film is 24 frames per second. It has been found that a suitable fly-wheel associated with the intermittent cam prevents any detectable deterioration of the reproduced sound due to the dissymmetry of the intermittent cam.

Other Projecting Sequences and Mechanisms

There is some evidence that the television picture transmitted by a system depending completely upon the storage principle might not be as satisfactory as one transmitted by a system in which the film image is projected upon the photo-emissive mosaic either continuously or during the entire scanning period. It is natural, therefore, that investigations of the latter type of system should have been made. So far, the results obtained have not been wholly satisfactory and certainly have not been as excellent as those produced by the storage method described in the previous section. However, refinement of certain projection methods may at some time in the future make other systems of greater interest. It is, therefore, of value to digress and review some of the various schemes that have been investigated.

For obtaining a continuous and constant light image on the Iconoscope photo-emissive mosaic, a commercial type of theater projector was used, in which the film passed the picture gate at constant speed

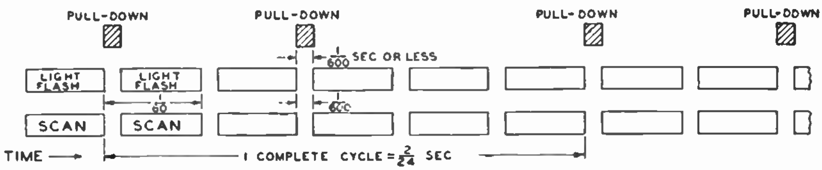

Fig. 3—Idealized sequence of events in film transmission by Iconoscope.

and a stationary projected image was obtained by means of an "optical intermittent." This projector employed several rocking mirrors on a rotating wheel. The lens system was properly proportioned for the projection of the small image required for the Iconoscope mosaic plate. In testing this system it was noted that the television performance was limited by various types of movement in the projected optical image and by low resolution. Motion of the optical image, in addition to causing objectionable motion in the received television picture, also contributed to loss of resolution in the picture. This is due to the storage action of the Iconoscope whereby the signal derived from each element of the mosaic in scanning is due to the summation of all the light which has fallen on that element since the preceding contact of the scanning beam. The effect is similar to that obtained when the optical image on a sensitized photographic plate moves during exposure.

Figure 3 shows a projection sequence by which an intermittent type projector might project film on an Iconoscope for the entire scanning time provided the pull-down occurred in the almost prohibitively short time of 1/600 second or less. This would permit projection throughout the entire scanning period. There is no apparatus now available for meeting the 1/600 second pull-down requirement. If suitable equipment could be developed it is doubtful if the film would withstand the stresses imposed by the rapid motion.

An experimental projector using a continuously moving film, and a rocking mirror for producing a stationary image, was built and tested. A diagrammatic view of it is shown in Figure 4. The cam-

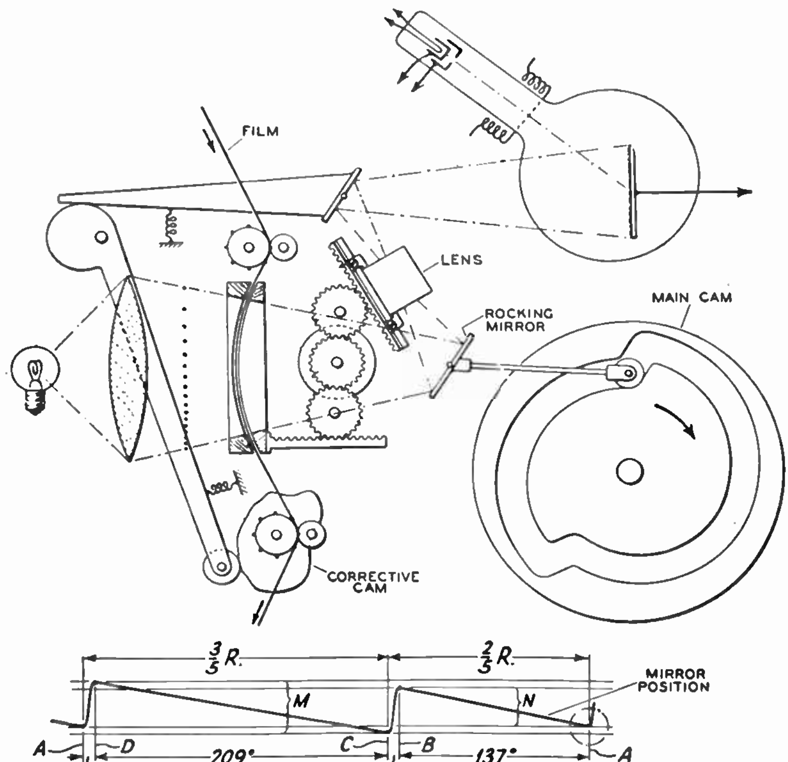

Fig. 4—Experimental rocking mirror projector.

driven mirror was arranged to neutralize accurately the film motion during the intervals marked "light flash" in Figure 3 and to return to receive light from the next consecutive film frame during the 1/600-second non-uniformly-spaced intervals marked "pull-down." Limitations were found due to slight non-uniform illumination of the approximately two and one-half frames of film always in the picture gate. This resulted in objectionable flicker in the television picture. Also, in spite of the very small amplitude of motion required for the rocking mirror, the cam and follower-roller created a very annoying noise and were subject to rapid wear.

Description of Film Projector

It is of interest to return now to the method for using film which is considered best at present, and review the apparatus in more detail.

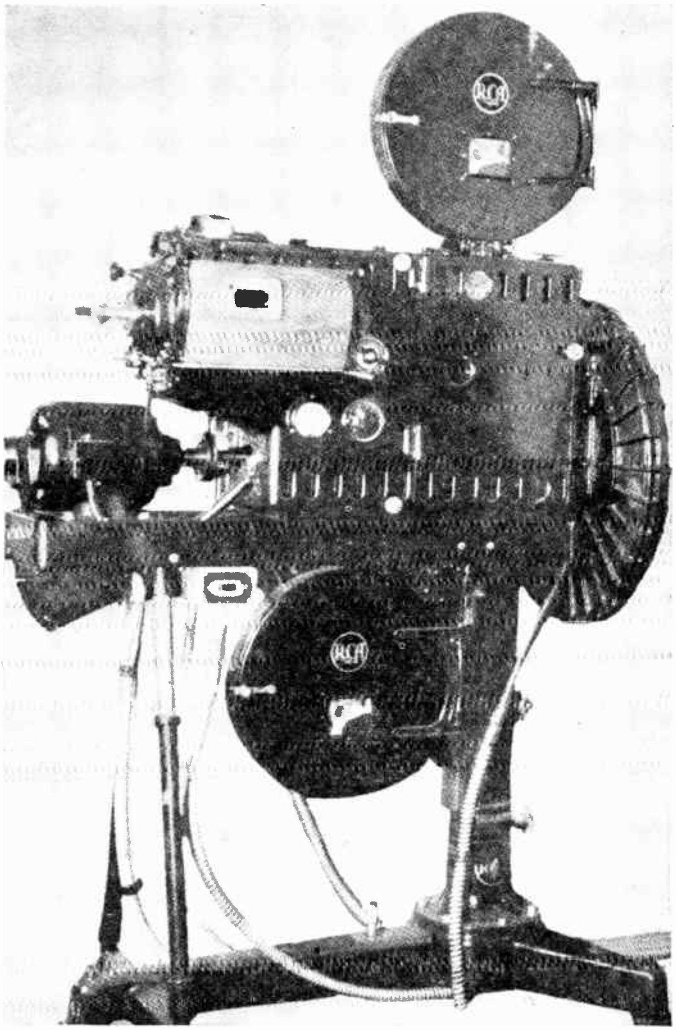

Fig. 5—RCA 35-mm sound motion-picture projector designed for 30-frame-per-second television with interlaced scanning.

Figure 5 is a general view of a 35-mm sound motion-picture projector* designed for 30-frame-per-second television with interlaced scanning.

* This projector was built to RCA specifications by International Projector Corp.

This projector differs from standard theater projectors in the following major respects:

1. A special shutter is used to provide efficient light pulses of very short time duration for projecting, 60 times per second, images of the film pictures onto the photo-emissive mosaic of the Iconoscope.

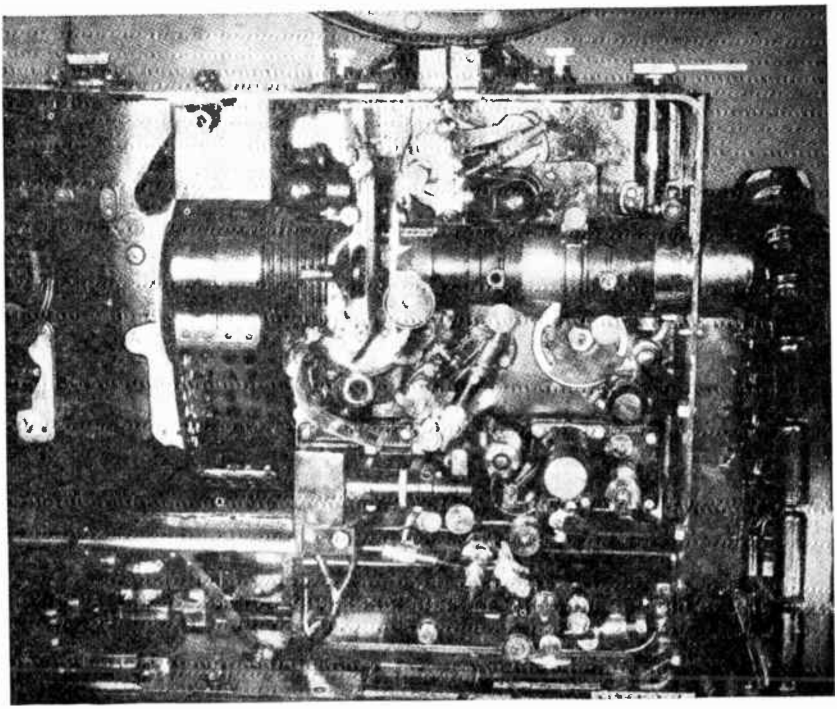

Fig. 6—Film projector for television with doors open.

2. The intermittent mechanism is designed for the three-to-two ratio of pull-down periods required in using 24-frame film for 30-frame television.
3. A special synchronous driving motor is used to assure that the projector mechanism always "locks-in" in proper time relation with the synchronizing pulses.
4. An additional film gate with light source and photo-electric cell is included near the picture gate for deriving a control potential which varies with the average density of the film.

In the projector shown in Figure 5, it was impracticable to locate the shutter between the light source and the film. The shutter was, therefore, mounted just beyond the projection lens. Sufficient clearance between the shutter and lens was provided to permit a limited movement of the lens for focusing. The time during which the image may be projected onto the photo-emissive mosaic of the Iconoscope is limited to the vertical return time of the scanning beam. With present television standards this is not more than 10 per cent of 1/60 second or 1/600 second.

In order to make efficient use of the projection lens, it is necessary for the aperture in the shutter to be at least as wide as the diameter of the lens. A large diameter shutter (23″) is necessary to meet this requirement. This shutter rotates at 3600 r.p.m. and has a peripheral speed of approximately 4¼ miles per minute. The shutter is enclosed in the circular housing which is shown at the extreme right-hand side of Figure 5. In the shutter housing opposite the projection lens is a window through which the picture is projected. The shutter disc is made of two overlapping sections of thin metal. These two sections can be rotated with respect to each other through a small angle in order to vary the width of the aperture. Figure 6 is a photograph showing the film side of the projector with the cover removed.

A second gate is located four frames of film above the picture gate. To the left of this gate, as shown in Figure 6, is a lamp housing. To the right of this gate is a photocell housing which also includes an optical system for forming an image of the lamp filament on the photocell. The output voltage from this photocell is rectified, and after being passed through a suitable filter is used to control the return-line blanking signals. The resultant variation in the blanking signals is used to control the average brightness of the reproduced picture. Figure 7 shows a view of the film side of the projector with a film threaded ready for projection.

Although the projector just described is equipped with a small 30-ampere arc, either an incandescent lamp or an arc may be used.

Equipment for Broadcasting Television Film Programs

In considering the production of motion-picture films for television, it is important to review the apparatus that will be used in the broadcasting station. The essential elements of a system for television transmission from motion-picture film are shown in Figure 8. These include: Film Projector; Iconoscope Film Camera; Camera Amplifier Equipment; Control Equipment; Monitor Equipment; Synchronizing Generator.

The Iconoscope camera used with the film projector includes deflecting circuits and a pre-amplifier for the video signals. This pre-amplifier provides a signal level suitable for transmission over a coaxial cable to the camera amplifier equipment. The camera is usually mounted on one side of a wall, with the film projector located on the other side. The picture is projected through a window in the wall into the camera onto the photo-emissive mosaic of the Iconoscope.

The camera amplifier equipment includes apparatus for amplifying further the video signals from the camera and a line amplifier to

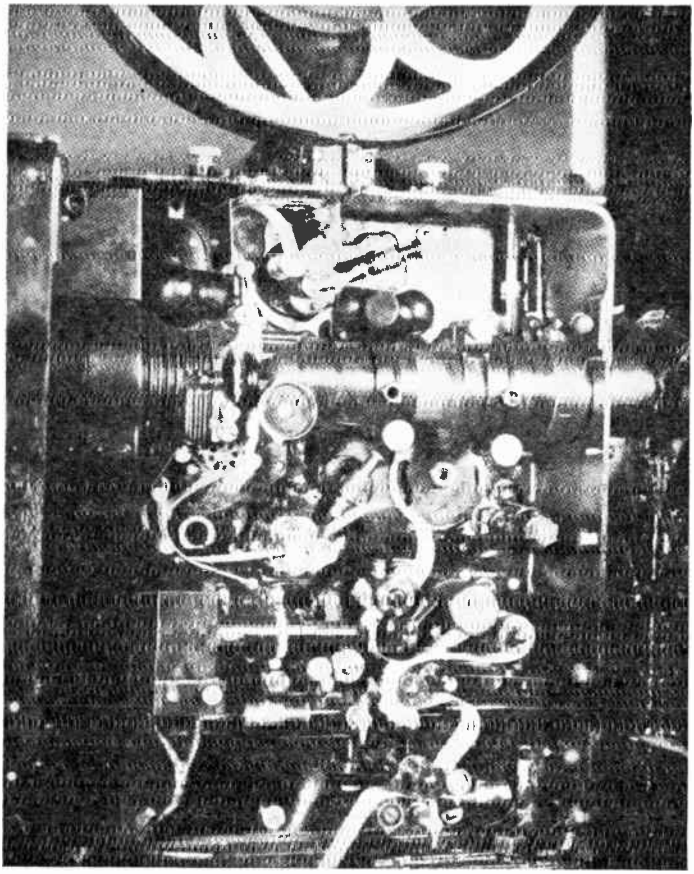

Fig. 7—View of film projector for television showing film path.

prepare these signals for transmission over coaxial cable to any desired location. Amplifiers providing suitable wave shapes for horizontal and vertical deflection of the Iconoscope beam are included as well as the power supplies for the several parts of the system. This equipment is usually rack-mounted in some convenient location.

The control equipment provides means for varying the video signal gain, the picture brightness, and the uniformity of the picture-background illumination (shading), and for starting and stopping the

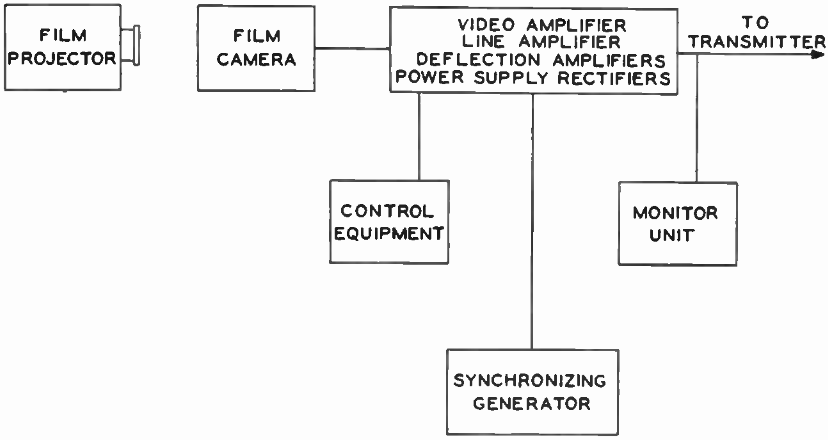

Fig. 8—System for television transmission from motion-picture film.

film projector. In an installation designed to provide a continuous program from motion picture film, where two or more film projectors and television channels are included, controls are also provided for switching from one channel to another.

The monitor equipment includes a 12″ Kinescope by means of which television images obtained from the film can be viewed. It also includes a cathode-ray oscilloscope for observing the wave shapes

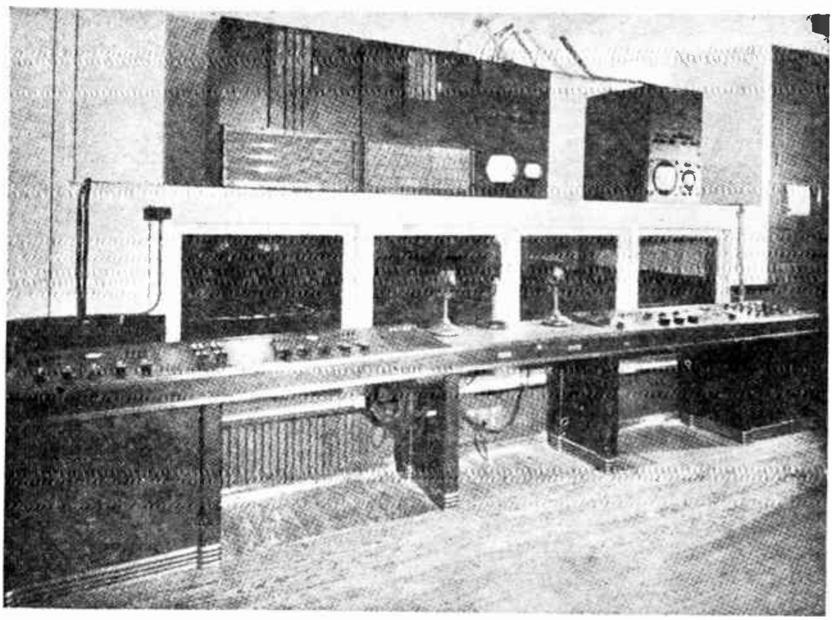

Fig. 9—Television control equipment for studio and film-type cameras.

and amplitudes of the television signals. This monitor equipment is usually located so that it may be observed conveniently by the operator manipulating the control apparatus.

The synchronizing generator supplies the several complex waveforms which are required to determine the timing of scanning processes in the transmitting equipment and to synchronize the reconstruction of the images at the receivers. The wave shapes of the synchronizing signals have been standardized by the Radio Manufacturers Association.

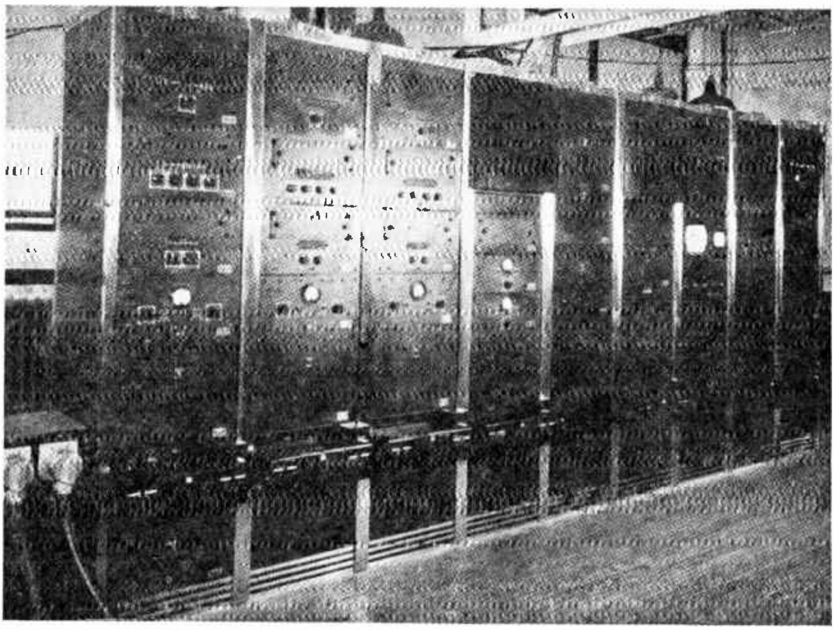

Fig. 10—Television terminal equipment suitable for television broadcasting stations.

Views of television equipment of a type suitable for television broadcasting stations are shown in Figures 9 and 10. Figure 9 shows an installation of control equipment for studio and film type cameras. This equipment is grouped on a common control console with the monitors mounted in a recess in the wall above the console. In this installation, the control engineer may look directly into the studio. Figure 10 shows a typical installation of racks of television-terminal equipment.

SIMPLIFIED TELEVISION APPARATUS

For specialized services, more simple and compact television equipment is desirable. Apparatus of this sort has been developed both for direct studio pickup and for film applications. The simplified equipment suitable for producing television signals and television images from motion-picture film includes all of the elements previously described, but in far more compact form. The equipment less the Iconoscope camera and the projector is included in one cabinet approximately 44 inches high, 34 inches wide, and 21 inches deep. This equipment produces a television signal which is suitable for transmission to remote viewing positions or for other uses.

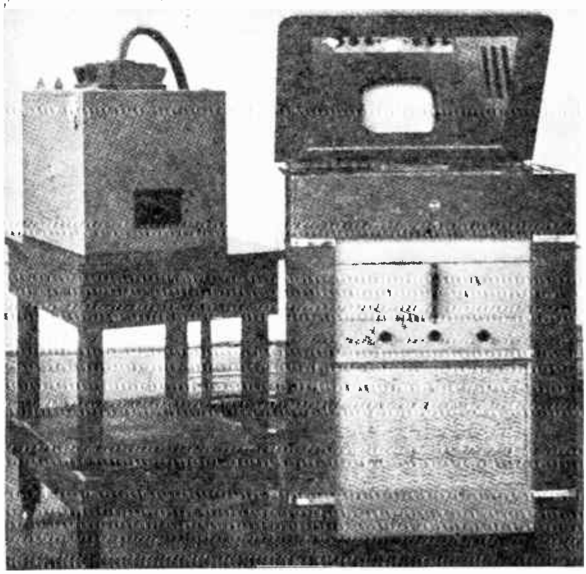

Fig. 11—Simplified television apparatus suitable for judging the merits of motion-picture film.

This simplified equipment is not as flexible in some respects as the broadcasting type of equipment, nor does it lend itself well to large complex systems. However, it does provide the facilities necessary for judging the merits of film for television use. In this simplification of apparatus and circuits, the synchronizing wave shapes do not conform entirely to the Radio Manufacturers Association standards. The synchronizing signals are, however, satisfactory for the self-contained monitor and for other receivers or reproducing devices, but the adjustments may be a little more critical than would be the case with standard synchronizing signals. Figure 11 shows a view of the equipment with the Iconoscope camera mounted on a simple wooden dolly.

Apparatus for Judging the Merits of Motion Picture Film for Television

An earlier paper* reviewed some of the limitations inherent in present-day television and compared these with similar limitations in motion-picture film and apparatus. Experience has indicated that the production of television pictures from a particular film is the only practical method for judging the merits of that film as television program material. It is, therefore, suggested that this method be used for checking and studying motion-picture films produced for television programs and for determining the usefulness of film available from other sources. Apparatus of the type used at the television-broadcasting station or apparatus of the simplified type just described will be satisfactory for this service.

Film Best Suited for Television

Laboratory work and field-test experience permits some preliminary generalizations on film that has given good results for television. Comment is here directed to the technical characteristics of film and not to the entertainment qualities. It appears that film having characteristics best suited for theater projection is also generally best for television. Studio sets having all dark backgrounds should be avoided. A goodly number of close-ups should be used, but these should be generously interspersed with long shots. Some experience may be necessary to take into account the resolution limits* of present-day television. Special processing of film does not seem to be necessary.

Film photographed in color directly from real life or nature appears satisfactory for television. Some cartoons in color have not given particularly satisfactory results. Thus, it appears that there may be no really serious technical problems in the production of motion-picture films suitable for television-program material.

THE ORTHICON, A TELEVISION PICK-UP TUBE*†

BY

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Summary—Extensive laboratory and field tests have shown that the Iconoscope is capable of transmitting clear, sharp television pictures even under conditions of unfavorable illumination. An analysis of the operation of the tube suggests that improved efficiency and freedom from spurious signals should result from operating the mosaic at the potential of the thermionic cathode, rather than near anode voltage. The beam electrons then approach the target with low velocity, and the number of electrons which land is dependent upon the illumination.

Several new designs were developed to make sure that the beam of low-velocity electrons was brought to the cathode-potential target in a well-focused condition, that the scanning pattern was undistorted, and that the focus of the beam was not materially altered by the scanning process. A magnetic field perpendicular to the target was found to be useful in focusing and guiding the beam. In some of the tubes, the scanning beam was released by a flying light spot moving over a photocathode. In other tubes it was found more convenient to develop the beam in an electron gun with a thermionic cathode. Special horizontal and vertical deflection systems capable of operating in the presence of a magnetic field were evolved.

The electron gun type of pick-up tube, which has been called an Orthicon, has a maximum signal current output over 300 times the noise in a typical amplifier. The signal is proportional to light intensity. The resolution is sufficient for the transmission of a 441-line picture. Spurious signals are neglible. Within the accuracy of measurement, all the photoemission is converted into video signals. In its present developmental form, the Orthicon gives promise of becoming a useful television pick-up tube.

INTRODUCTION

WITH the beginning of scheduled television broadcasting in New York, it is natural that a great deal of attention should be given to the commercial aspects of the art. Engineers realize, however, that research and development work must continue if future improvements are to be assured. Further investigations have, therefore, been carried on to provide ways to transmit clearer images, with less illumination. This paper will discuss an improved form of pick-up tube resulting from some of these investigations.[1]

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† Reprinted from *RCA REVIEW*, October, 1939.

‡ Now with the Research Department, RCA Laboratories Division, Princeton, N. J.

[1] Albert Rose and Harley Iams, "Television Pick-up Tubes Using Low-Velocity Electron-Beam Scanning," *Proc. I.R.E.*, Vol. 27, No. 9, pp. 547-555, September 1939.

At the present time, the RCA television system uses Iconoscopes to convert the optical image into a sequence of video signals for transmission to the receiver. Several previous publications have described these tubes and explained how they operate,[2,3,4] so that a brief review of their characteristics is sufficient to serve as a basis for a discussion of new pick-up tubes.

Extensive laboratory and field tests have shown that the Iconoscope is capable of transmitting clear, sharp television images, even under conditions of unfavorable illumination[5]. The spectral response is sufficiently like that of the human eye to give a natural appearance to the viewed scene. In the circuits associated with the tube, provision is made to "keystone" the deflection (so as to make the scanning beam move over the mosaic in a rectangular pattern) and to introduce shading signals into the amplifier (to compensate for the "dark spot" signal).

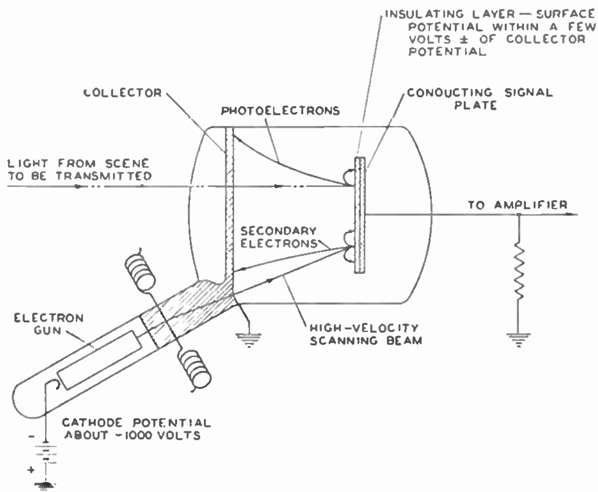

Fig. 1—Schematic diagram of an Iconoscope.

In the course of these tests, several significant discoveries were made. One of these was that the good operating sensitivity of the tube

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[2] V. K. Zworykin, "The Iconoscope—A Modern Version of the Electric Eye," *Proc. I.R.E.*, Vol. 22, No. 1, pp. 16-32, January (1934).

[3] V. K. Zworykin, "Iconoscopes and Kinescopes in Television," *RCA REVIEW*, Vol. 1, No. 1, pp. 60-84, July 1936.

[4] V. K. Zworykin, G. A. Morton and L. E. Flory, "Theory and Performance of the Iconoscope," *Proc. I.R.E.*, Vol. 25, No. 8, pp. 1071-1092, August (1937).

[5] Harley Iams, R. B. James, and W. H. Hickok, "The Brightness of Outdoor Scenes and Its Relation to Television Transmission," *Proc. I.R.E.*, Vol. 25, No. 8, pp. 1034-1047, August (1937).

is obtained in spite of an operating efficiency only 5 or 10 per cent of that which is theoretically attainable[4]. In other words, during typical operation only about one-third of the photoelectrons which the mosaic emits are drawn away, and only about one-quarter of the stored charge is effective in producing the video signal. This lowered efficiency is connected with the release of secondary electrons from the mosaic by the scanning beam.

The situation is illustrated in Figure 1, which shows the essential parts of an Iconoscope. Light in the optical image focused on the mosaic causes the emission of photoelectrons, leaving a pattern of charges corresponding in intensity to the light and shade of the scene to be transmitted. This pattern of charges is scanned by a beam of electrons, which strike the mosaic at high velocity. On the average, each beam electron releases several secondary electrons. Since the mosaic is an insulated surface, the electron current leaving it must (on the average) be equal to the electron current arriving. Thus, when the tube is in darkness, only as many secondary electrons can escape from the mosaic as there are beam electrons which arrive. The rest of the secondary electrons fall back on the surface near or far from the point of emission. Nonuniformities in the escape and rain of secondary electrons cause the "dark spot" signal.

Many of the numerous secondary electrons are emitted with appreciable velocity, so that the condition of one secondary electron leaving for each beam electron arriving means that the electric field near the mosaic is such as to hinder the escape of secondary electrons. This field also reduces the escape of the photoelectrons, which have lower average emission velocity. When the mosaic is lighted, those photoelectrons which escape contribute a positive charge to the lighted parts of the surface. These charges are partly dissipated by the rain of secondary electrons, but sufficient charge is stored during a frame period to produce a strong signal in an amplifier connected to the signal plate when the beam releases the stored charge.

Low-Velocity Electron Beam Scanning

If one could ignore the immediate practical problems and choose an ideal mode of operation for a television pick-up tube, he might want to provide a field strong enough to draw away all of the photoelectrons which are emitted, and he might prefer to do the scanning without involving secondary emission in the process. These conditions can be met by operating the mosaic, in known fashion, at the potential of the cathode in the electron gun.

Cathode-voltage operation is possible, for the potential of an insulated surface exposed to an electron beam is stable at this voltage[6]. High-vacuum cathode-ray tubes are usually operated so that the beam electrons strike the screen with high velocity, and liberate many secondary electrons. The screen potential then adjusts itself (usually near anode voltage) so that the number of secondary electrons which escape is equal to the number of beam electrons which arrive. However, the other stable condition occurs when the surface is at cathode potential. The beam electrons then approach the target, but are repelled and retire without striking. If the target becomes slightly positive (by photoemission, for example), the beam electrons land without producing appreciable secondary emission and restore the original voltage.

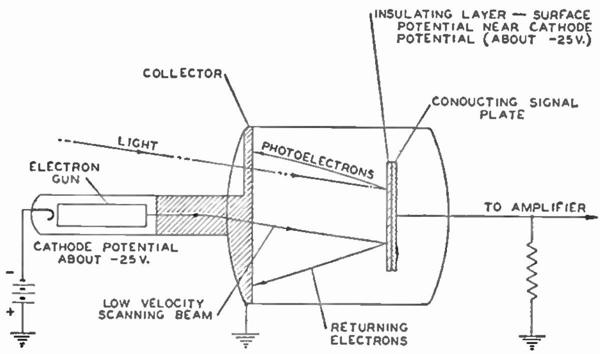


Fig. 2—Pick-up tube with cathode-potential target.

Figure 2 illustrates the operation of a television pick-up tube with its photosensitive target at cathode potential. In the absence of light the beam electrons approach the surface, are slowed to zero velocity, and then are drawn away. There is no signal in an amplifier connected to the signal plate. When light falls on the mosaic, all the photoelectrons are pulled away by the strong electrostatic field between the mosaic and the anode. The charges given the surface are not dissipated by a rain of electrons, but are stored until the scanning beam approaches. When the beam comes near a lighted area, and finds it a few volts positive, electrons land until their negative charge brings the surface to cathode potential again. The velocity with which the beam electrons reach the surface is so low that secondary emission is not involved to any appreciable extent. The signal is simply due to the impulses given the signal plate by the beam electrons, as they arrive at the lighted parts of the target.

[6] A. W. Hull, "The Dynatron," *Proc. I.R.E.*, Vol. 6, No. 1, p. 5, February (1918).

Design Problems

So far, much of the discussion has related to ideals. Some of the problems which must be solved before a pick-up tube can be made to operate in a satisfactory manner with its target at cathode potential are illustrated in Figure 3. When the electron beam is deflected, in usual fashion, at an angle to the axis of the tube, the electrons do not approach the target perpendicularly. The negative voltage needed to keep an electron from landing on the mosaic is only as great as the component of velocity perpendicular to the mosaic, and the electric field is not able to stop motion parallel to the surface. When conditions are as illustrated, the beam charges one part of the target to cathode potential, and other parts slightly more positively. Also, the point of contact of the beam may be elongated into a line. This situation may

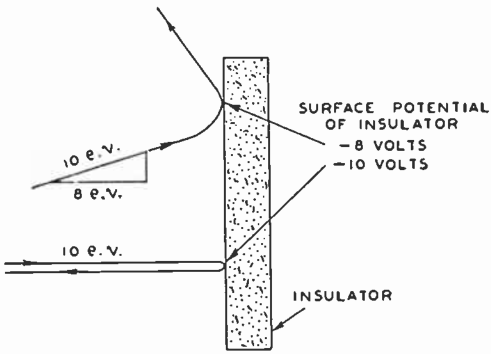


Fig. 3—Electron paths near target.

be expected to cause a loss of resolution at the edges of the picture, and unstable operation when the electrons move with considerable velocity tangent to the surface of the target. These considerations suggest that it would be preferable for the beam always to approach the mosaic nearly perpendicularly, or for the beam to be constrained.

Another important problem is that of providing sufficient beam current (about one microampere) in a beam which retains its small diameter when the electrons are slowed to almost zero velocity and are subject to strong local fields at the mosaic surface.

In the solution of these problems, a magnetic field perpendicular to the mosaic and extending to the source of the electron beam has been found useful. When the field is made sufficiently strong, the beam is focused and refocused many times between the cathode and the target. In a uniform magnetic field, the final size of the scanning spot is substantially the same as that of the source of the beam. The magnetic field can also be used to keep the beam electrons from proceeding very far across the surface of the mosaic at grazing incidence. Electrons which tend to move in this fashion cut across the lines of flux, and move in circular paths. The diameter of these circles can be made as small as is desired by increasing the strength of the magnetic field.

Several types of pick-up tubes based upon these principles have been designed. In some, the scanning beam is developed at a photocathode illuminated by a moving spot of light, while in others the beam originates at a thermionic cathode. The name Orthiconoscope (or Orthicon, for short) has been used to denote these tubes in which the target is operated at cathode potential. (The Greek prefix "orth",

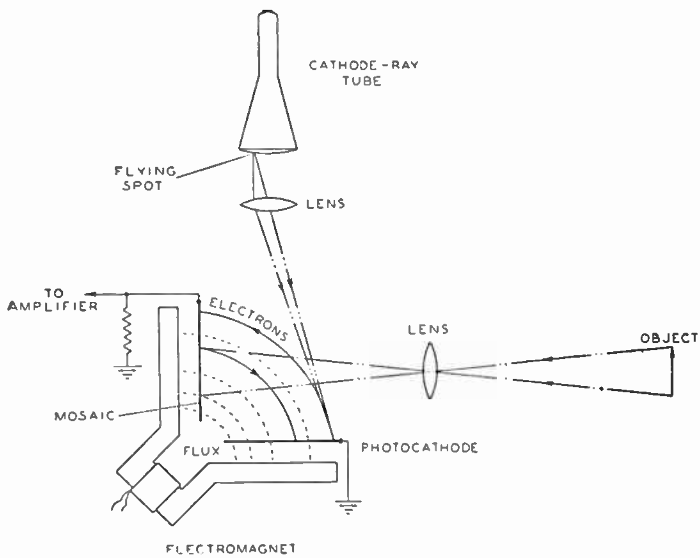


Fig. 4—Pick-up tube using photoelectrons for beam.

meaning straight, is added to the well known term "Iconoscope" to describe the linear relation between light and signal output, which has been observed.)

Tubes with Photoelectric Scanning Beam

One of the tubes which has been built and tested is illustrated schematically in Figure 4. A charge image of the scene to be transmitted is developed by focusing the optical image upon a conventional mosaic. The electron beam which scans the mosaic is produced by photoemission from a conducting photocathode, which is illuminated by a flying light spot focused from the face of a cathode-ray tube with a short-time-lag screen. At any instant, the light from a single spot on the cathode-ray-tube screen is focused on the photocathode. The emitted photoelectrons are guided by the curved lines of flux between the pole faces of an electromagnet, and the beam is also focused by this field. When beam electrons approach a lighted part of the mosaic, they are absorbed and produce signals in the amplifier. From the dark parts of the mosaic the beam electrons are reflected back to the photocathode. Photoemission from the mosaic is collected by the photocathode.

The picture transmitted by this tube is quite sharp, and is free from spurious shading. Streaking, which one might expect because of time lag in the luminescent material, is not apparent. Most of the discharging of charged areas takes place in the first fraction of a

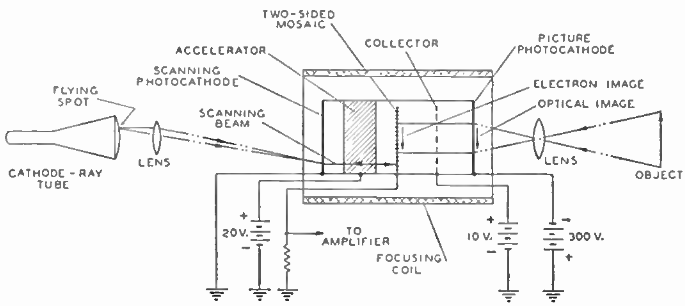


Fig. 5—Pick-up tube using secondary emission amplification.

microsecond, and after that the beam electrons fail to reach the target and their presence is not observed.

Tubes which incorporate secondary emission amplification of the image, in a fashion somewhat comparable with the method used in the Image Iconoscope[7], have also been built and tested. Such an arrangement is shown in Figure 5. In this case, the optical image is focused upon a translucent photocathode and the resulting photoemission is focused upon a two-sided mosaic by means of an axial magnetic field. Secondary electrons released from the mosaic by the high-velocity picture electrons are drawn away to a collecting electrode, giving image amplification. The electron beam which scans the other side of the mosaic is obtained from a flying light spot moving over another translucent photocathode, and is focused by the same axial magnetic field which focuses the electron image. As in the previous tube, the scanning beam restores the mosaic to the potential of the

[7] Harley Iams, G. A. Morton, and V. K. Zworykin, "The Image Iconoscope," *Proc. I.R.E.*, Vol. 27, No. 9, pp. 541-547, Sept. (1939).

scanning cathode. When the tube is operated, the gain in sensitivity due to secondary emission amplification is readily observed.

While a number of tubes which use photoelectric scanning beams have been tested, these two examples are sufficient to indicate the methods that have been used. These methods make possible the scanning of a large mosaic with a well-focused beam of low-velocity electrons, resulting in the transmission of video signals free from spurious signals. Further, an increase in sensitivity through secondary emission amplification may be obtained. However, because the auxiliary apparatus to generate the flying light spot represented a complication, an investigation was made of tubes in which the beam is derived from a thermionic cathode, and is deflected in the presence of a magnetic field.

Tube with a Thermionic Scanning Beam

The most important problem in the design of a pick-up tube, operating with its target at cathode potential, and using an electron gun to

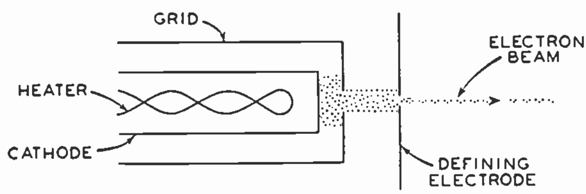


Fig. 6—Electron gun.

generate the scanning beam, is that of deflecting the beam without defocusing it. In the case of the tubes described above, each point on the target had a corresponding point on the photocathode such that the same magnetic line intersected both of them. Photoelectrons generated at one end of the line at the photocathode were focused at the target with substantially one-to-one magnification. A slight enlargement of the scanning spot occurred due to the emission velocities of the photoelectrons transverse to the magnetic field. If, in some way, larger transverse velocities were introduced into the motion of the beam electrons, they would describe helices of larger amplitude around the magnetic lines and result in a larger spot at the target. In the case of a tube using an electron gun, only the central point on the target is normally connected with the cathode by a magnetic line If electrons from the gun are to reach other points on the target, either they must cross the magnetic lines of the axial field or they must be guided to other points by warping the axial magnetic field. Both of these devices are used in the form of Orthicon to be described. To insure, however, that the deflected spot is not larger than the undeflected spot it is necessary that no significant amount of velocity transverse to the magnetic field, imparted to the beam electrons by the deflection system, be retained by the electrons as they approach the target.

The simple electron gun shown in Figure 6 generates a stream of electrons moving parallel to the axis with a velocity of about a hundred volts. The cross section of the beam is limited to the size of a picture element by the defining aperture in the last electrode. In this state, the beam enters the deflection system and in this state it should emerge except for a displacement from the axis.

The high-speed horizontal deflection is accomplished by a pair of electrostatic deflection plates in combination with the axial magnetic

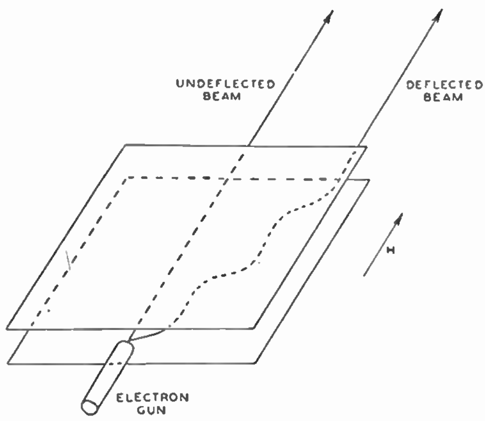


Fig. 7—Path of electron beam in electric and magnetic fields.

field. A somewhat schematic representation of the path of the beam through the plates is shown in Figure 7. The beam is seen to be deflected in a plane parallel with the plates and to diverge from the axis only while it is between the plates. After leaving the plates, the beam continues parallel to the axis. The amplitude of deflection is proportional to the electric field and the transit time of the electrons through the plates, and inversely proportional to the strength of the axial magnetic field. Since the maximum amplitude of deflection is limited to the width of plates, these must be as wide as the target to be scanned. The wiggles in the beam in Figure 7, when viewed from the end of the plates, that is along the axis, appear as a series of cycloids. This is the two dimensional path described by electrons moving in crossed electric and magnetic fields. If the electric field from the deflection plates could be sharply cut off at the entrance and exit edges, the transit time of the beam could be adjusted so that only an integral number of cycloids would be performed by the beam in passing through the plates. In this way, none of the transverse velocity represented by the cycloidal motion would be retained by the beam after it left the plates. While this is a possible arrangement, it has been found that a less critical way of insuring that the emergent beam retains substantially none of the transverse velocity acquired in the plates is to suppress the amplitude of cycloidal motion within the plates. The amplitude of cycloidal motion may be considerably reduced by admitting the beam to the plates through a gradually increasing electric field and similarly letting it leave through a gradually decreasing field. For this reason, the deflection plates are flared out at the entrance and exit ends.

Two significant distinctions are to be noted, in the above account, between electrostatic deflection in the presence of a magnetic field, as

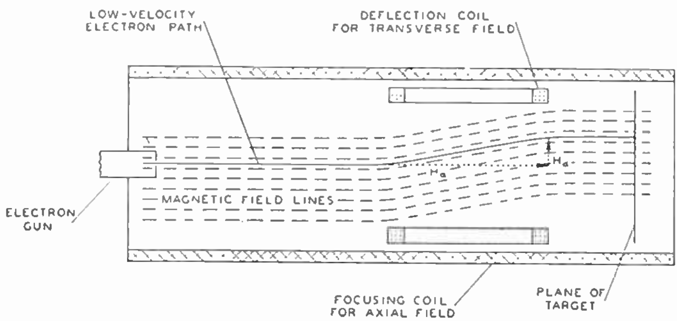


Fig. 8—Path of electron beam in warped magnetic field.

in an Orthicon, and electrostatic deflection in a magnetic field free space, as in the usual cathode-ray tube. First, the plane of deflection, which is perpendicular to the plates in the usual electrostatic deflection, has been rotated through ninety degrees into a plane parallel with the plates, in an Orthicon. Second, while the usual plates impart a transverse velocity to the beam which causes the beam to continue to diverge from the axis after leaving the plates, the plates in an Orthicon cause the beam to diverge from the axis only while the beam is between the plates. The axial magnetic field constrains the beam to motion parallel with the axis after it leaves the plates.

The low-speed vertical deflection is accomplished by a pair of magnetic coils. Here, again, while magnetic coils are used in the usual cathode-ray tube without an axial magnetic field, their action in an Orthicon is essentially different by virtue of this field. Briefly, the axial magnetic field rotates the plane of deflection through ninety degrees and causes the electrons in the beam to move parallel to the axis after leaving the deflection coils. The average path of the beam through a pair of deflection coils immersed in an axial magnetic field is shown in Figure 8. The amplitude of deflection is proportional to the magnitude and axial length of the deflection field and inversely proportional to the magnitude of the axial magnetic field. From Figure 8, it is evident that the separation of the coils must be as large as the height of the target to be scanned. While Figure 8 shows the average path of the beam to follow the magnetic lines closely, the actual path contains a helical motion imparted to the electrons due to their passing through a curved magnetic field. For small deflections, the amplitude of this helical motion is of the order of, or less than, the helical motion of the electrons due to their emission velocities. The magnetic coils may therefore be considered as a means of displacing the beam from

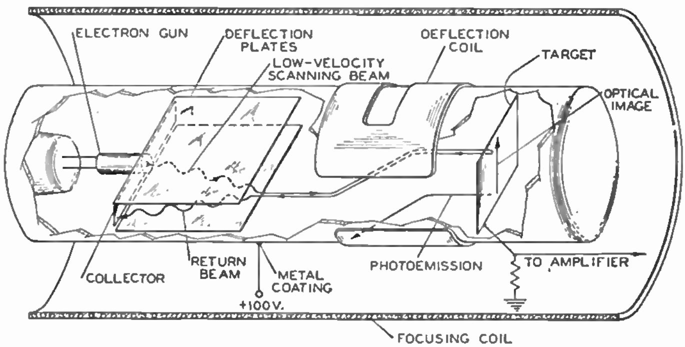


Fig. 9—Schematic diagram of an Orthicon.

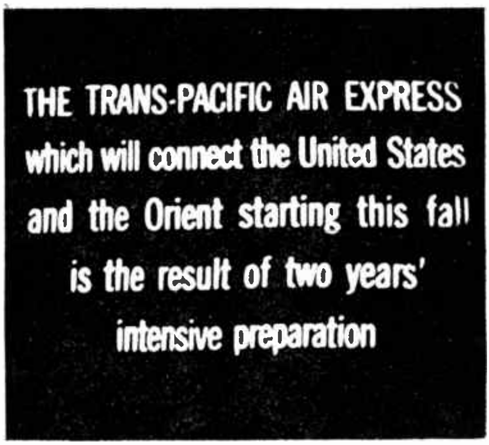
the axis, without contributing any significant transverse velocity to the beam electrons.

The beam, as it leaves the deflection system and approaches the target, is in substantially the same condition as when it left the gun, except that it may be displaced from the axis. Since the target surface is at cathode potential, the beam passes through an electric retarding field sufficient to slow it to near zero velocity at the target. The target is a thin mica sheet, the side facing the beam being covered with a mosaic of photosensitive elements and the side away from the beam being coated with a translucent conducting film of metal known as the signal plate. The optical picture is focused on the photosensitive side through the translucent signal plate. The photoelectrons are drawn away from the zero-potential surface to various positive electrodes near the target (see Figure 9). As described earlier, if the element approached by the beam has been unlighted, the beam electrons will not land on the target, but will be brought to rest near it and be accelerated back away from the target. If the element has been lighted, enough of the beam electrons will land to replace the photoelectrons that have been drawn away during the previous frame time. In this way, the beam maintains the target at cathode potential and generates the video signal. The electrons which do not land at the target retrace substantially their original path as they return toward the electrostatic plates. In passing back through the plates, the beam diverges from its going path in the direction of the original deflection, as shown in Figure 9. The beam eventually strikes an elongated collector electrode, also shown in Figure 9.

Operating Characteristics

The description of the operation of an Orthicon would lead one to expect a set of simple operating characteristics. This expectation has been borne out by numerous observations on the tubes. Briefly, substantially no signal is transmitted with no light on the target. With an optical picture focused on the target, a signal proportional to the light intensity at each point is transmitted. The maximum signal is limited by the amount of beam current. In some tubes a modulated beam current of one microampere has been observed. This corresponds to a signal current about three hundred times the noise level of a typical television amplifier.

Not only is the transmitted signal proportional to the light on the target, but also the conversion of possible photoemission into signal takes place at substantially 100 per cent efficiency. This requires that the photoemission from the target be saturated throughout the frame time, that the charge be fully stored for that time, and that all of the stored charge be useful in producing a video signal when the scanning beam passes over it. Tests made on Orthicons have shown that the photoemission from the target is saturated when the collector electrodes surrounding the target are more than twenty volts positive with respect to the target. Since these electrodes are usually about plus one hundred volts, a saturated photocurrent is assured under static conditions. During a frame time, the photoemission from the target is swept over the collecting electrodes by the field from the vertical deflecting coils. To test whether the photocurrent was saturated equally throughout the frame time and equally stored, a spot of light was projected on the target once a frame time for about one-tenth of the frame time. The signals, both of the photocurrent and of the discharge process, were observed on an oscilloscope. The storage time was varied from zero to a full frame time by varying the time at which the spot was projected on the target. No variation in either of these signals was observed throughout the full range of storage time, indicating that the photocurrent was equally saturated and the charge equally stored throughout the frame time. A final overall test, which compared the transmitted signal with a known amount of light on a target of known photosensitivity, showed that (within ability to measure) all of the stored charge (or by the above tests, all of the photoemission at the target) was utilized in producing video signal. As a result of this high operating efficiency, Orthicons with a target photosensitivity of one microampere per lumen exhibited approxi-



(Courtesy R. D. Knell)

Fig. 10—Television picture transmitted by an Orthicon.

mately the same operating sensitivity as Iconoscopes with a target photosensitivity of ten microamperes per lumen.

Observations on the resolution of the transmitted picture showed more than four-hundred-line resolution over the entire picture and as high as six-hundred-line resolution in the center. For the particular size of target used, two and one-half inches wide, this means that the scanning beam near zero velocity can resolve elements on the target less than one two-hundredth of an inch apart.

Two representative pictures transmitted by an Orthicon are shown in Figures 10 and 11. No shading-compensation signals were introduced into the television system from which these pictures were taken.

CONCLUSIONS

As a result of the tests which have been described, it may be concluded that the operation of a television pick-up tube with its target at cathode potential makes possible (1) the efficient conversion of photoemission into video signals, (2) a large signal output, and (3)

(Courtesy R. D. Knell)

Fig. 11—Television picture transmitted by an Orthicon.

the elimination of spurious signals. While developmental Orthicons incorporating these features have been built and operated, additional work to determine optimum designs is in progress.

SOME FACTORS AFFECTING THE CHOICE OF LENSES FOR TELEVISION CAMERAS*†

By

H. B. DeVore and Harley Iams‡

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Summary—*The design of a television camera for a particular use involves the selection of a lens which, under the poorest conditions of illumination which are expected, will form a sufficiently bright image to meet the requirements of the pickup tube used. When the size and sensitivity characteristics of the pickup tube and video-frequency amplifier, the required signal-to-noise ratio, and the angle of view of the camera are known, the specifications of the lens can be computed.*

If a tube having a sensitive surface of width W inches and an operating sensitivity of s microamperes per lumen is to be used in picking up a scene having a surface brightness B candles per square foot, with a horizontal angle of view α, the following equations can be used to determine the lens to be used:

$$\text{focal length: } F = \frac{W}{2}\cot\frac{\alpha}{2} \text{ inches}$$

$$\text{numerical aperture: } f = \frac{0.064\, W\sqrt{TBs}}{\sqrt{10^6 \bar{I}_n N}}$$

in which T is the light-transmission factor of the lens, $\bar{I}_n$ is the equivalent root-mean-square noise current at the input of the amplifier used with the tube, and N is the required signal-to-noise ratio.

The paper discusses the derivation of these equations and includes charts to facilitate computations.

Comparison of predicted results with the observed performance of the apparatus have shown good correlation.

The public would like to see television pictures of many events, sometimes of subjects that are difficult to transmit. The engineer often wishes to know beforehand whether the camera available will produce the desired results, or how to design one which will. The purpose of this discussion is to outline the factors involved in such a design, particularly matters related to the choice of a lens suitable for the pickup tube which is to be used.

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‡ Now with the Research Department, RCA Laboratories Division, Princeton, N. J.

The specifications of the lens are determined by the size and sensitivity characteristics of the pickup tube to be used, the brightness of the scene, the noise characteristics of the pickup tube and the video-frequency amplifier, the required signal-to-noise ratio, and the angle of view of the camera. These factors determine the focal length and diameter of the lens. The lens having thus been specified, it is of interest to determine the depth of field for the system. However, as increasing depth of field and brightness of image impose contradictory requirements on the aperture of a lens, the depth of field can only be increased with a given pickup tube at the expense of signal-to-noise ratio in the transmitted picture.[1]

The operating sensitivity of a pickup tube may be affected by a number of factors and a direct quantitative comparison of different tubes is, therefore, quite difficult to make. This operating sensitivity, depending on the manner in which the tube operates, may differ greatly from the photosensitivity of the sensitive surface. By the choice of somewhat simplified conditions, measurements may be made which permit a useful estimate of the operating sensitivity. The method of measurement has been described in a recent publication.[2]

A spot of light is made by forming on the sensitive surface of the pickup tube an image of a slit illuminated by a lamp operated at a color temperature of 2870 degrees Kelvin. This color temperature had become standard for phototube data, and so has also been adopted for pickup-tube measurements. The signal from the output of the video-frequency amplifier is observed on an oscillograph which has its horizontal sweep synchronized with the horizontal scanning of the pickup tube. The signal from the slit then appears as a narrow peak superimposed on the base line which represents the signal from the black area of the picture. The amplifier-oscillograph system is calibrated by applying a known alternating current through a small standard resistor in series with the amplifier input. By varying the intensity of illumination in the slit and observing the corresponding signal currents delivered to the amplifier, it is possible to obtain a curve showing the performance of the tube. If the spot of light is moved to any part of a uniformly sensitive surface, the signal output remains constant. Hence, the amplitude of the short pulse of current due to a small spot of light can be taken as a measure of the signal current which would flow continuously if the whole target were lighted.

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[1] Harley Iams, G. A. Morton, and V. K. Zworykin, "The image iconoscope," *Proc. I.R.E.*, vol. 27, pp. 541-547; September, 1939.

[2] R. B. Janes and W. H. Hickok, "Recent improvements in the design and characteristics of the iconoscope," *Proc. I.R.E.*, vol. 27, pp. 535–540; September, 1939.

It has been found convenient to express the operating sensitivity of a pickup tube under given conditions as the quotient of the signal current produced at the amplifier input by scanning the slit of light, divided by the light flux required to illuminate the entire picture at the intensity of the slit image. The sensitivity (within a substantially linear part of the characteristic curve) may be expressed in units of microamperes per lumen. This form of expression for tube sensitivity has two advantages. The curves for different tubes may be compared directly; size or type of tube is immaterial. Second, computation of the signal output is simplified. Since the output impedances of nearly all pickup tubes in general use are high compared with normal amplifier input impedances, the signal, expressed in microamperes of modulated

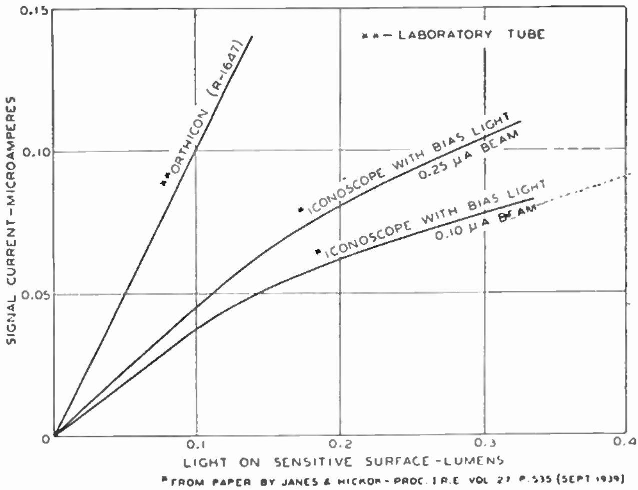

Fig. 1

current, is independent of the amplifier input resistance which is used. (The noise level in the amplifier, however, depends upon the capacitance of the pickup tube and input circuit.) The curves of Figure 1 represent the performances of some typical pickup tubes.

The tubes for which the characteristics are given in this figure are laboratory models, and the curves are not necessarily to be taken as being representative of these classes of tubes. They are, rather, presented to illustrate a convenient form for representing the sensitivity of a pickup tube.

These curves will not necessarily be accurate for subjects other than a light spot on a black background, but it is believed that the relative forms and positions of the curves will be maintained for other typical scenes.

The signal output which is required of a pickup tube is determined by the background "noise" originating in the pickup tube and amplifying system used with the tube. Since signal and noise will be amplified together, an initially unsatisfactory signal-to-noise ratio can never be improved. The noise generated in the pickup tubes which do not use electron multipliers is small compared with that originating in the amplifier. The simplest input for the amplifier is that in which the signal current of the pickup tube flowing through an input resistor produces a potential drop which is applied to the grid of the first tube, usually a high-transconductance pentode. The noise current originating in this type of amplifier consists of two significant components: thermal-agitation noise developed in the input impedance, and current fluctuations in the plate circuit of the tube. The mean-square, thermal-agitation-noise current is given by

$$d\,\overline{i_t^2} = \frac{4kT}{R}\,df \tag{1}$$

for a frequency band df, in which R is the input resistance, T the temperature in degrees Kelvin, and k the Boltzmann constant. The tube noise may be conveniently expressed in terms of the equivalent resistance which, connected between grid and cathode, would cause mean-square current fluctuations of the same magnitude in the plate circuit by reason of thermal agitation, the resistor being assumed to be at 300 degrees Kelvin. Data giving values of this equivalent resistance for a number of tubes are available in the literature. If this resistance is denoted by R_t, the equivalent mean-square fluctuation current in the input circuit resulting from the tube is

$$d\,\overline{i_s^2} = 4kTR_t\left[\frac{1}{R^2} + (\omega C)^2\right]df \tag{2}$$

in which C is the total input capacitance (pickup tube, amplifying tube, and circuit).

The total equivalent input noise current caused by the tube and circuit is obtained by adding (1) and (2)

$$d\,\overline{i_n^2} = d\,\overline{i_t^2} + d\,\overline{i_s^2} = 4kT\left\{\frac{1}{R} + \frac{R_t}{R^2} + R_t(\omega C)^2\right\}df. \tag{3}$$

The second term R_t/R^2 is, for any probable values of R_t and R, entirely negligible. The total current is obtained by integrating (3) over the pass band of the amplifier and is

$$\overline{I_n^2} = \int_0^{f_m} \overline{i_n^2}\, df = \frac{4kT}{R} f_m \left\{1 + \frac{R_t R (\omega_m C)^2}{3}\right\}. \tag{4}$$

The root-mean-square noise current will then be

$$\overline{I_n} = 2\sqrt{\frac{kT}{R} f_m \left\{1 + \frac{R_t R (\omega_m C)^2}{3}\right\}}. \tag{5}$$

Analyses of more complicated input circuits have indicated that these do not give any appreciable reduction in noise level as compared with this simple one.

When part of the amplification is obtained by means of an electron multiplier, as in some dissector tubes, a different expression should be used to determine the effective noise current. In this case, the noise originates predominantly at the photocathode. The expression for equivalent noise current at the cathode as given by Larson and Gardner[3] has the form

$$i\Delta_t = \sqrt{\frac{i_0 e}{\Delta t}}$$

where $i\Delta_t$ = the current responsible for the noise,
i_0 = the current for one picture element,
e = the electronic charge, and
Δt = the time for the transmission of one picture element.

This expression may be converted to the following formula for root-mean-square noise current at the input of the multiplier

$$\overline{I_n} = \sqrt{2ei_0 f_m}.$$

The signal-to-noise ratio (N) which is required for a television picture has been discussed by Zworykin, Morton, and Flory.[4] Their

[3] C. C. Larson and B. C. Gardner, "The image dissector," *Electronics*, vol. 12, pp. 24–27 and 50; October, 1939.

[4] V. K. Zworykin, G. A. Morton, and L. E. Flory, "Theory and performance of the iconoscope," *Proc. I.R.E.*, vol. 25, pp. 1071–1092; August, 1937.

conclusions are that for an average picture, the ratio of peak picture signal to root-mean-square noise of 3 to 1 is highly objectionable; 10 to 1 gives an acceptable picture, while 30 to 1 is excellent. A ratio of 10 to 1 may be assumed to be the minimum useful value. For this ratio, the signal current which the pickup tube must deliver can be computed, and from the sensitivity of the tube, the total light which must be projected on its sensitive surface can be calculated. For the case of the conventional amplifier.

$$L = \frac{10\,\overline{I_n}}{s \times 10^{-6}} = 10^7\,\frac{\overline{I_n}}{s} \tag{6}$$

L being the light flux in lumens and s the operating sensitivity in microamperes per lumen.

The first consideration in the selection of a lens which will deliver this amount of light to a given television pickup tube is the field that must be covered. This may be expressed as the angle of view of the transmitted picture measured in a horizontal plane. The specification of the angle of view determines the focal length of the lens, which may be computed from the formula

$$F = \frac{W}{2} \cot \frac{\alpha}{2} \tag{7}$$

in which W is the width of the sensitive area of the tube and α is the horizontal angle of view.

The choice of the angle of view determines the effective magnification of the picture. It is generally accepted that the most favorable viewing distance for a television picture is about five times the height of the picture. At this point the distance between two successive scanning lines in a 441-line picture subtends an angle of about 1.7 minutes, which is approximately the resolution limit of the eye. The width of the picture subtends an angle of about 15 degrees. If the camera uses a lens which covers this angle of view, the reproduced image will have the same appearance as the original scene viewed from the camera. With a lens which gives an angle of view of 30 degrees, the reproduced picture will have the same appearance as the original scene viewed by an observer at a distance from the scene about twice that of the camera. The perspective will, of course, be somewhat faulty, but considerable distortion of the perspective is tolerable, as photographic experience has shown.

The angle of view being given, the diameter which a lens must have to enable a scene of specified brightness to be picked up satisfactorily is determined by the amount of light which must fall on the sensitive area of the tube.

The intensity of illumination in the image formed on a plane surface by a lens has been worked out by a number of people. A useful expression has been given by Goodwin.[5] For a uniform source having a brightness of B candles per square foot, the intensity in the image is

$$E = \frac{\pi B T}{4f^2} \cos^4 e \tag{8}$$

lumens per square foot, where T is the transmission factor, f is the f number of the lens, and θ is the angle to the axis of the system made by a light ray striking the area under consideration. For large angles of view, the intensity at the edges of the image falls off rapidly compared with that at the center. While observing that such a decrease in intensity across the field is present, it is convenient for the purposes of calculation to assume that the field is uniformly illuminated at the intensity that exists at the center. This is, to some extent, justified by the fact that the center of interest in a picture is in general in the center of the field.

In a tube in which the sensitive surface has a width W and an aspect ratio 4:3, the area of this surface is 3/4 W^2, and the light falling on it is

$$L = \frac{3\pi}{16} \frac{BTW^2}{f^2}. \tag{9}$$

Solving (9) for the numerical aperture, we obtain

$$f = \frac{0.77W\sqrt{T}}{\sqrt{L/B}}. \tag{10}$$

As B is given in candles per square foot and L is in lumens, then W must be expressed in feet. A more convenient expression is

$$f = \frac{0.064W\sqrt{T}}{\sqrt{L/B}} \text{ with } W \text{ given in inches.} \tag{11}$$

[5] W. N. Goodwin, Jr., "The Phototronic photographic exposure meter," *Jour. Soc. Mot. Pic. Eng.*, vol. 20, pp. 95–118; February, 1933.

The transmission coefficient of the lens T, which represents chiefly losses by reflection at the air-glass surfaces in the lens, depends on its structure. Values given in the literature for typical lenses range from 40 to 70 per cent transmission, being generally lower as the lens aperture increases. For purposes of calculation a value of 60 per cent may be taken as a reasonable mean. For this value,

$$f = \frac{0.050W}{\sqrt{L/B}}. \tag{12}$$

This relationship is plotted in the curves of Figure 2.

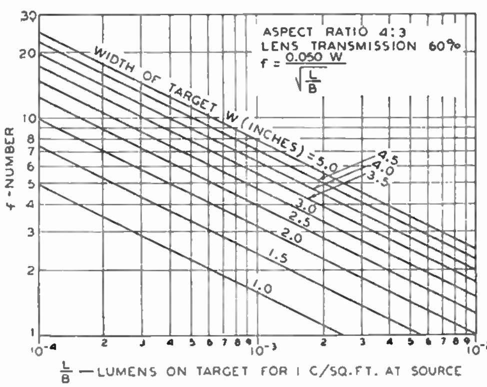

Fig. 2

The lens diameter may be found from the expression

$$A = \frac{F}{f}. \tag{13}$$

The angle of view and image brightness required by the tube determine completely the lens which must be used for a given brightness of scene. It is also desirable to be able to estimate the depth of field of the system.

If a lens of diameter A and focal length F forms on a plane surface at F_1 an accurately focused image of an object at a distance P_1, then for any other object distance the image of a point source of light will be a circle having a diameter depending on the object distance. If the maximum tolerable diameter of this "circle of confusion" is δ, then the limiting distances from the lens between which the image will be considered satisfactorily focused will be given by

$$P_2 = \frac{\dfrac{FA}{\delta}}{\dfrac{FA}{\delta P_1} + 1} \quad \text{and} \quad P_3 = \frac{\dfrac{FA}{\delta}}{\dfrac{FA}{\delta P_1} - 1}\,. \tag{14}$$

In photographic technique, it has long been accepted that the limit of the tolerable circle of confusion is 0.010 inch on a picture to be viewed at 10 inches distance. That is, δ may subtend an angle of 0.001 radian or about 3.5 minutes. In a television picture the problem of choosing a suitable value for δ is complicated by the scanning process As the angle subtended by the width of a scanning line at normal viewing distance (1.7 minutes) is about half of the size of the photographically tolerable circle of confusion, it might be expected that the effect on δ of the scanning process should be relatively small.

In an attempt to determine a suitable magnitude for the permissible circle of confusion in the case of television pickup, the following experiment was made. A photograph was made of a number of men located over a range of measured distances from a camera which was accurately focused for infinity. In the picture so obtained, the size of the circle of confusion corresponding to each man's position was calculated from the constants of the camera lens. Reference lines were drawn to take care of subsequent magnification and a lantern slide made from the photograph. This was projected on a number of pickup tubes at several magnifications. In each case judgment was rendered by a number of observers as to the man in the transmitted picture who appeared to represent a dividing point between those in acceptable focus and those definitely out of focus. From these observations the acceptable circle of confusion was calculated. Values were obtained ranging from about 1/120 to 1/180 of the height of the scanning pattern. No differences were noticed among the different types of tubes tested, all of which were capable of better than 450-line resolution, as observed from transmission of a standard test pattern.

It seems evident that the departure from perfect focus which is tolerable depends to a considerable extent on the type of subject being transmitted. Much more latitude is certainly permissible in pictures of people than in subjects having high detail contrast, such as a resolution pattern, to take an extreme case. More work along this line is needed before accurate statements can be made concerning depth of field. For the present, it is believed that a useful estimate can be made by assuming the tolerable circle of confusion to be 1/200 of the picture height, i.e., about two scanning lines. With the standard aspect ratio 4:3,

$$\delta = \frac{3W}{4 \times 200} = \frac{W}{267}. \tag{15}$$

Substituting this in (14) for depth of field, we obtain

$$P_2 = \frac{\dfrac{267\,FA}{W}}{\dfrac{267\,FA}{W\,P_1} + 1} \qquad P_3 = \frac{\dfrac{267\,FA}{W}}{\dfrac{267\,FA}{W\,P_1} - 1} \tag{16}$$

or, since from (7),

$$\frac{F}{W} = \frac{\cot \dfrac{\alpha}{2}}{2}$$

$$P_2 = \frac{133\,A \cot \dfrac{\alpha}{2}}{\dfrac{133\,A \cot \dfrac{\alpha}{2}}{P_1} + 1} \qquad P_3 = \frac{133\,A \cot \dfrac{\alpha}{2}}{\dfrac{133\,A \cot \dfrac{\alpha}{2}}{P_1} - 1} \tag{17}$$

A convenient concept in depth-of-field calculations is that of hyperfocal distance (HD). This is the distance beyond which all objects are in good focus when the lens is focused accurately for infinity. Placing $P_1 = \infty$ in the expression for P_2 in (14), we obtain

$$HD = \frac{FA}{\delta} = 133\,A \cot \frac{\alpha}{2}. \tag{18}$$

The expressions for the near and far limits of focus can be rewritten in terms of the hyperfocal distance as follows:

$$P_2 = \frac{HD}{\dfrac{HD}{P_1} + 1} \qquad P_3 = \frac{HD}{\dfrac{HD}{P_1} - 1}. \tag{19}$$

It is seen that when a lens is focused accurately for the distance HD, all objects from $HD/2$ to infinity are in good focus. Further, when the lens is focused for a distance $P_1 = HD/k$ in which k is any constant, the region of good focus extends from $HD/(k+1)$ to $HD/(k-1)$.

For a given lens aperture and angle of view, the hyperfocal distance may be read on the chart of Figure 3.

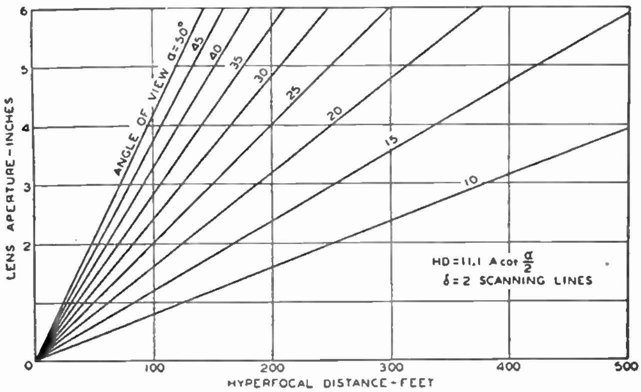

Fig. 3

By way of summary, it might be well to run through a representative series of calculations. Assume that an iconoscope which has a mosaic 4.75 inches wide and a sensitivity of 0.4 microampere per lumen is to be used. The camera is to cover a 30-degree angle of view and is to be used at a baseball park where the brightness of the players is expected to be at least 50 candles per square foot.[6]

From the angle of view and the width of the mosaic, it is found that the required lens must have a focal length of

$$F = \frac{W}{2} \cot \frac{\alpha}{2} = 2.38 \times 3.73 = 8.9 \text{ inches.} \tag{7}$$

[6] Data relative to the brightness of some typical scenes are given by Harley Iams, R. B. Janes, and W. H. Hickok, "The brightness of outdoor scenes and its relation to television transmission," *Proc. I.R.E.*, vol. 25, pp. 1034–1047; August, 1937.

The preamplifier to be used with the iconoscope uses a type 6AC7/1852 input tube (equivalent grid resistance for noise, 500 ohms) with a 200,000-ohm input resistor. The total tube, circuit, and iconoscope capacitance may be 26 micromicrofarads.

The amplifier has a 4-megacycle pass band. Then the root-mean-square noise current will be

$$\bar{I}_n = 2\sqrt{\frac{kT}{R} f_m \left\{1 + \frac{R_t R(\omega C)^2}{3}\right\}}.$$

$$= 2\sqrt{\frac{1.37 \times 10^{-23} \times 3 \times 10^2 \times 4 \times 10^6}{2 \times 10^5} \left\{1 + \frac{5 \times 10^2 \times 2 \times 10^5\ (6.28 \times 4 \times 10^6 \times 2.6 \times 10^{-11})^2}{3}\right\}} \quad (5)$$

$$= 2.2 \times 10^{-9} \text{ ampere.}$$

If the entire mosaic were uniformly illuminated at the brightness of the high lights, the luminous flux required to produce a satisfactory signal-to-noise ratio of 10 to 1 would be

$$\mathrm{L} = 10^7 \frac{\bar{I}_n}{s} = \frac{10^7 \times 2.2 \times 10^{-9}}{0.4} = 0.056 \text{ lumen} \quad (6)$$

and for a scene having a brightness of 50 candles per square foot, the ratio $L/B = 0.056/50 = 1.12 \times 10^{-3}$ lumen per candle per square foot.

From the curves of Figure 2 the f number of this lens will be

$$f = 7.1$$

and the lens will have a diameter

$$A = \frac{F}{f} = \frac{8.9}{7.1} = 1.25 \text{ inches.} \quad (13)$$

We find then that the use of an $f/7$ lens of 8.9 inches focal length will permit a picture to be transmitted under the above conditions which will have a just passable signal-to-noise ratio.

From Figure 3 we find the hyperfocal distance of this lens to be

$$HD = 52 \text{ feet.}$$

When this lens is focused at a distance of 52 feet the depth of field will include all objects from 26 feet to infinity. When the lens is focused on an object 20 feet away, the field will include objects between 14.5 feet and 32.5 feet from the camera.

Improvement in the quality of the picture with respect to noise may be obtained by increasing the aperture of the lens. For example, to obtain a signal-to-noise ratio of 30 to 1 would require an $f/4$ lens of the same focal length. This improvement, however, will be accompanied by a corresponding decrease in depth of field ($HD = 88$ feet).

Checks on the usefulness of computations such as those given above have been made in a number of cases, involving several different pickup tubes. The tests were made by observing television pictures transmitted under a variety of conditions, and comparing the picture quality with that which the formulas would lead to expect. From these tests it was found that the computations led to conclusions which were reasonably close to the observer's judgment of picture quality.

THE RCA PORTABLE TELEVISION PICKUP EQUIPMENT*†

By

G. L. Beers,[a] O. H. Schade,[b] and R. E. Shelby[c]

Summary—Spot news, athletic events, parades, etc., form an important source of television program material. In the spring of 1938, field experiments were started in New York City with mobile television pickup equipment. Two telemobile units were used each of which was about the size and shape of a 25-passenger bus and weighed 10 tons. The limitations of these telemobile units are discussed. Lightweight television pickup equipment has recently been developed. The new equipment includes a small iconoscope camera, camera auxiliary, camera-control, and synchronizing-generator units, and an ultra-high-frequency relay transmitter and receiver. Most of the units are about the size of a large suitcase and weigh between 40 and 70 pounds. Each of the units is described and some of the practical applications of the equipment are indicated.

Introduction

In the spring of 1938 field experiments were started in New York City with mobile television pickup equipment. Two telemobile units were used, one of which contained standard rack-mounted equipment for two cameras and the other housed a 159-megacycle 300-watt transmitter. Each unit was about the size and shape of a 25-passenger bus and weighed 10 tons. The total power required to operate both units was approximately 20 kilowatts. Field tests with the mobile units have definitely proved their usefulness in providing entertaining television programs. The size, weight, and power requirements of these units, however, have imposed definite restrictions on their use. In order to minimize these restrictions light-weight portable television pickup equipment has recently been developed. It is the purpose of this paper to describe the several units of this equipment and indicate some of its possible applications.

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General Characteristics of the Equipment

Past experience with all types of television pickup equipment has shown that it is desirable to locate all the control equipment at some central point if effective program supervision with a minimum of personnel is to be obtained. It is therefore essential that provision be made in portable television pickup equipment so that long lengths of camera cable can be used between the control equipment and the cameras. This requirement was responsible, to a considerable extent, for the division of the equipment into the several units shown in the block diagram in Figure 1. In this diagram the units for a complete system with a single camera are outlined by the solid lines. The

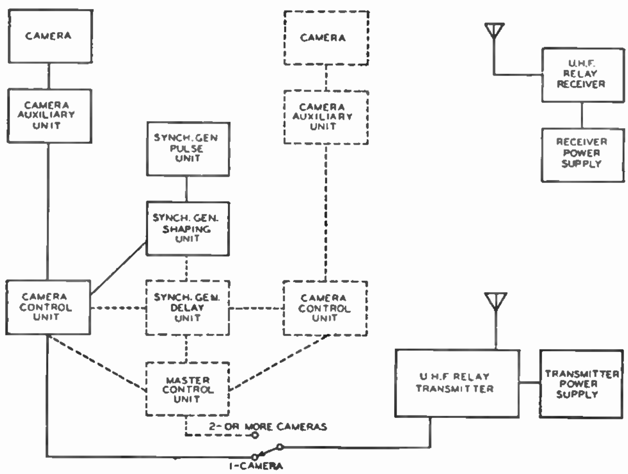

Fig. 1—Block diagram of complete portable television pickup system.

additional units required for a second camera are shown by the dotted lines. The only units which must be duplicated to add a third camera are the camera auxiliary and camera-control units. The receiver shown in the diagram is normally located at or near the main television transmitter and is therefore not a part of the equipment which must be transported to the remote pickup point.

The equipment is designed to produce synchronizing signals in accordance with the Radio Manufacturers Association standards. All the video-frequency amplifiers are adjusted to pass a frequency band from 30 cycles to 5 megacycles. Lengths of camera cable up to 500 feet can be used between the camera and camera-control equipment so that any two cameras can be separated by distances up to 1000 feet.

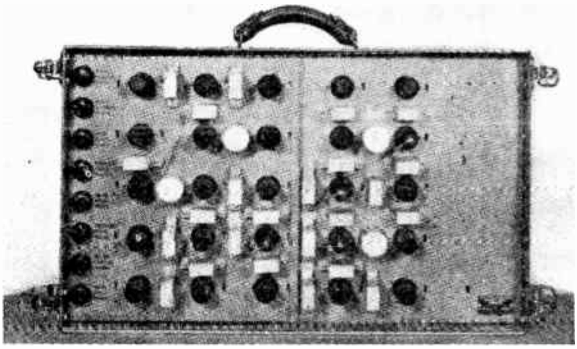

Fig. 2--Synchronizing-generator shaping unit—tube side, cover removed.

The equipment operates from any suitable 110-volt, 60-cycle, single-phase power-supply system. The power consumption for the portable equipment with one, two, and three cameras is 1400, 2000, and 2500 watts, respectively.

All the units are designed to make the tubes and circuit components as accessible as possible. The suit-case type of construction which is used for the camera auxiliary, camera-control, master-control, and three synchronizing-generator units is illustrated by Figures 2 and 3. These photographs show both sides of the synchronizing-generator shaping unit. The accessibility of the tubes on one side of the unit and the circuit components on the other is clearly illustrated. The central chassis portion of the unit is welded to the outside case to form a rigid unit. A view of the complete unit with the side covers in place

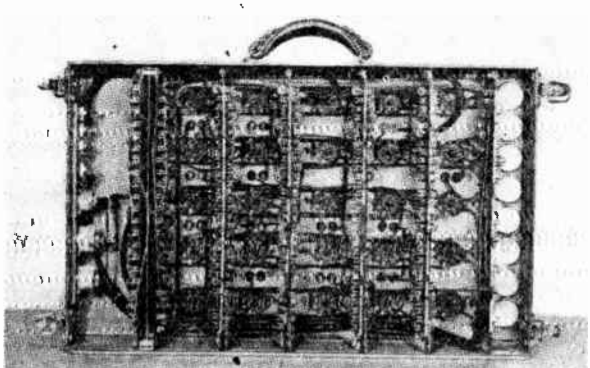

Fig. 3—Synchronizing-generator shaping unit—circuit component side, cover removed.

is shown in Figure 4. The over-all dimensions of the suit-case type units are $8 \times 15 \times 25$ inches and their weights vary between 45 and 72 pounds. The camera weighs 28 pounds and its tripod 30. The weights of the transmitter and its power-supply unit are 60 and 190 pounds. The total weights of the portable pickup equipment less interconnecting cables for one, two, and three cameras are 550, 850, and 1050 pounds, respectively. These weights can each be reduced by 250 pounds when the equipment is used at locations from which the television signals can be sent by coaxial cable to the main transmitter. The camera cable used with the equipment weighs approximately 0.6 pounds per foot. If 500-foot cables are used with each of three cameras the total weight of these cables is approximately the same as the total

Fig. 4—Synchronizing-generator shaping unit with covers in place.

weight of the equipment units.

The functions of the individual units are discussed in the descriptions which follow.

Camera

In order to keep the camera dimensions as small as possible the camera was designed to use the new 4½-inch iconoscope.

Development work on small iconoscopes has been in progress for several years. As the dimensions of an iconoscope are made smaller with a corresponding reduction in the mosaic area a loss in resolution and sensitivity is normally expected. A new gun structure which has recently been developed has made it possible to obtain adequate resolution from the 4½-inch iconoscope. Tests on this tube have also shown that its operating sensitivity when using a lens of a given aperture is substantially the same as that of the standard iconoscope. This unexpected sensitivity is attributed to the smaller spacing between the several tube elements which results in a more-efficient collection of the secondary electrons. This increase in the electron-collecting efficiency enables the tube to be operated at a higher average beam current for a given ratio of signal to dark-spot voltage.

Figure 5 is a photograph of the camera mounted on a standard

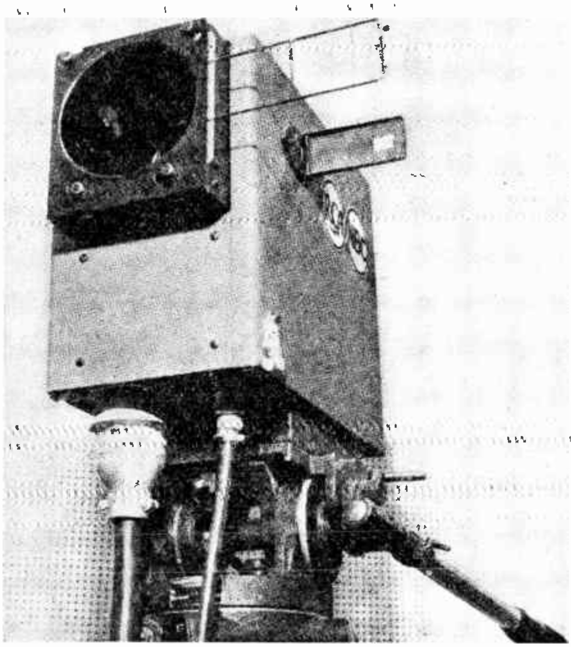

Fig. 5—Portable television camera on tripod.

motion-picture tripod and shows the wire-frame view finder which is used by the cameraman to keep the scene to be televised within the field of the camera. Focusing is done remotely by observing the picture on the kinescope in the camera-control unit. A Selsyn motor at the camera-control unit is used to operate a similar motor in the camera which in turn drives the lens carriage.

The internal construction of the camera is shown by the photographs in Figures 6 and 7. The focusing motor is housed within the

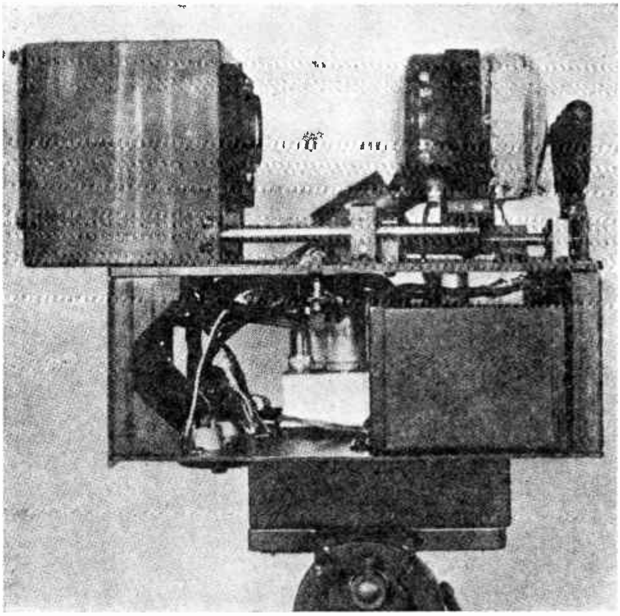

Fig. 6—Right side of portable television camera, cover removed.

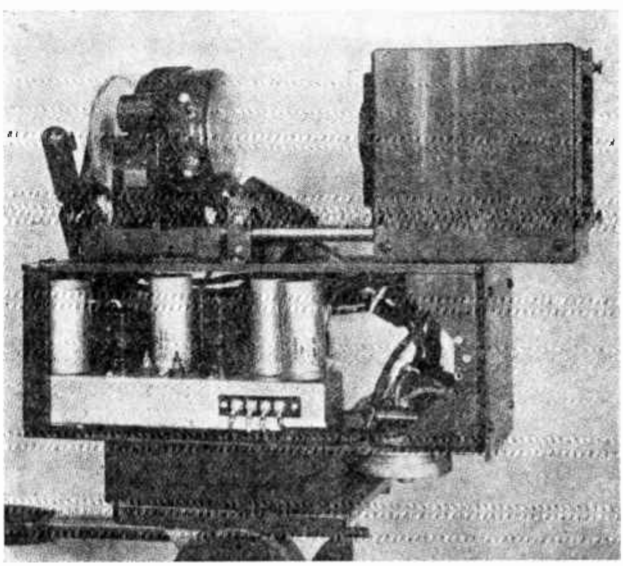

Fig. 7—Left side of portable television camera, cover removed.

rectangular shield which is visible in the lower right-hand corner of Figure 6. The two-stage pre-amplifier which can be seen in Figure 7 is used to raise the picture signals derived from the iconoscope to a satisfactory level to transmit over a short length of coaxial cable to the camera auxiliary unit. The iconoscope with its deflection yoke, the lens carriage, and the two shielded bias lights, which are mounted in back of the iconoscope, are all clearly shown in this photograph.

Figure 8 shows the lens-mounting arrangement. Lenses are interchanged by loosening the four thumbscrews shown in the photograph

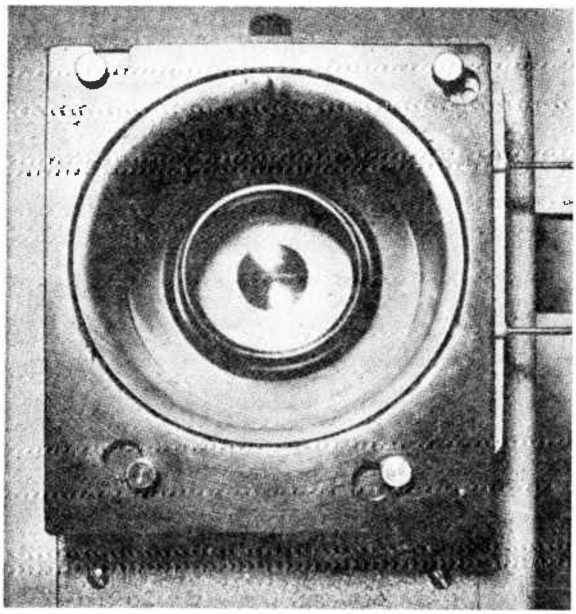

Fig. 8—Portable television camera—lens mounting.

and then rotating the lens mounting slightly in a counterclockwise direction. The complete lens-mounting assembly can then be removed by pulling it forward. Another lens is attached to the camera by reversing the procedure.

Camera Auxiliary Unit

The problem of obtaining satisfactory deflection of the iconoscope beam when a long length of camera cable is used is greatly simplified when the horizontal deflection power is developed in or near the camera. The use of a camera auxiliary unit makes it possible to meet this requirement and at the same time keep the dimensions and weight of the camera as small as possible. In addition to the horizontal-deflection circuits the camera auxiliary unit contains a four-stage video-frequency amplifier, iconoscope blanking and protection circuits, and a power-supply rectifier. This unit is connected to the camera through an eight-foot length of camera cable and is usually located between the legs of the camera tripod.

The video-frequency amplifier which contains both a high-frequency peaking and low-frequency losing circuit is used to raise the video-frequency signal from the pre-amplifier in the camera to a sufficient level so that a satisfactory signal-to-noise ratio is obtained at the receiving end of a 500-foot length of camera cable. This amplifier is assembled as a complete unit on a small chassis which is flexibly

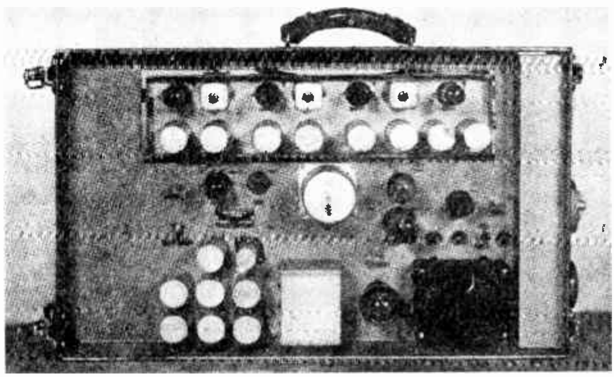

Fig. 9—Camera auxiliary unit—tube side, cover removed.

mounted in an opening in the main chassis of the camera auxiliary unit. The construction of the amplifier and the method of mounting are illustrated in Figures 9 and 10. It will be noted that this construction maintains the general arrangement of having all the tubes accessible from one side of the unit and the circuit components and wiring accessible from the other.

The voltage wave developed across the iconoscope deflection yoke is used to produce a vertical iconoscope blanking pulse. A protective circuit is provided by which the grid of the iconoscope receives a high negative bias if for any reason the deflection of the iconoscope beam is interrupted thereby preventing damage to the iconoscope mosaic. Horizontal saw-tooth waves produced in the camera-control unit are transmitted to the camera auxiliary unit over a flexible coaxial line included in the main camera cable. These waves are amplified by a

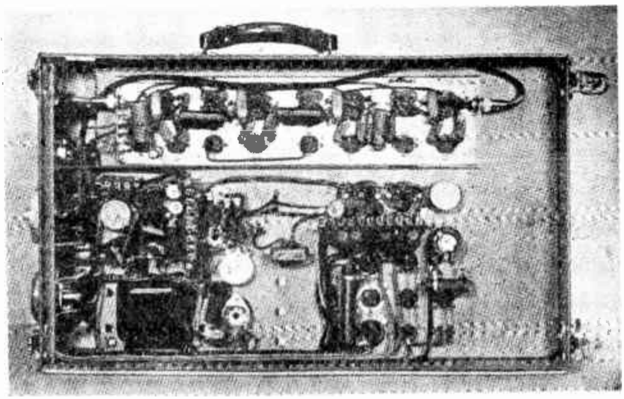

Fig. 10—Camera auxiliary unit—circuit-component side, cover removed.

two-stage amplifier in this unit and fed to the iconoscope deflecting yoke through a step-down transformer. The power-supply rectifier in the camera auxiliary unit supplies anode potentials to all the tubes in both this unit and the camera. A 15-conductor rubber-covered camera cable is used to provide the electrical connections between the camera auxiliary and camera-control units. The outside diameter of this cable is slightly under one inch. As previously stated, lengths of camera cable up to 500 feet can be used between the camera auxiliary and camera-control units.

Camera Control Unit

The camera-control unit is normally the central control point at which all the operating adjustments are made while the equipment is in use. The several functions of this unit are indicated by the block

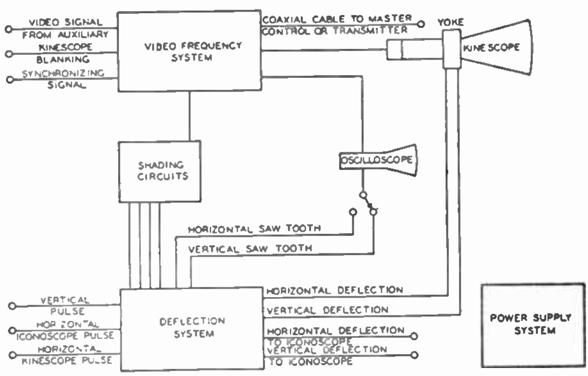

Fig. 11—Block diagram of camera-control unit.

diagram shown in Figure 11. The video-frequency system shown in this diagram amplifies the video-frequency signals received over the camera cable from the camera auxiliary unit. Blanking and shading signals are inserted in this portion of the system. The shading signals which are used are saw-tooth and parabola waves at both line and field frequencies. Controls are provided for varying both the amplitude and phase of these signals. In case only a single camera is used synchronizing signals can be inserted in the video-frequency system of the camera-control unit. Suitable signal potentials are supplied to the seven-inch kinescope which is used to monitor the picture and the two-inch oscilloscope which is used to observe the wave shapes of the picture signals. The video-frequency system is also designed to feed a two-volt peak-to-peak signal to a 75-ohm coaxial cable. Controls are provided for varying the video-frequency gain and the amplitude of the kinescope blanking signals.

In the deflection system line and field-frequency impulses received from the synchronizing generator are used to produce saw-tooth waves which are supplied to the iconoscope, kinescope, and oscilloscope. Provision is made in the synchronizing-generator delay unit for delaying the kinescope horizontal-deflection impulses with respect to the iconoscope impulses. Facilities are included in the camera-control unit for keystoning the horizontal deflection of the iconoscope. A switch is provided so that horizontal deflection of the oscilloscope at either line or field frequency can be obtained. Kinescope and iconoscope width, height, and centering controls are included. The line and field-frequency saw-tooth waves produced in the deflection system are also used as shading signals in the video-frequency system.

The power-supply system includes a high-voltage rectifier for supplying anode potentials to the iconoscope and kinescope. A low-voltage rectifier is used to supply anode potentials to all the other tubes in the camera-control unit. Focus and bias controls for the kinescope and iconoscope are included in this portion of the system.

Figures 12 and 13 show both sides of the camera control unit with the side covers removed. The front of this unit showing the kinescope, oscilloscope, and the several control knobs is illustrated by the photograph in Figure 14. A metal cover is supplied which protects these tubes and knobs when the equipment is not in use.

Master-Control Unit

When more than one camera is used to televise a desired scene some means must be provided for switching from one camera to another and

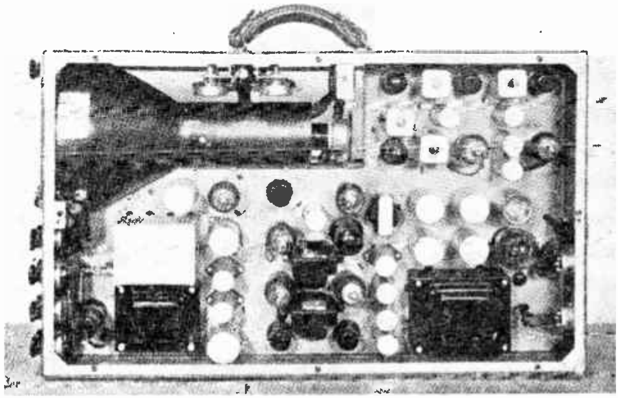

Fig. 12—Camera-control unit—tube side, cover removed.

for monitoring the "on-the-air" picture. In the portable pickup equipment these requirements are filled by the master-control unit. The block diagram in Figure 15 shows the several functions of this unit. The video-frequency system amplifies the signals received from the camera-control unit and supplies them to the kinescope and oscilloscope. Synchronizing signals are normally inserted in the video-frequency system of the master-control unit. The line amplifier in this unit is designed to provide a four-volt peak-to-peak signal across a 75-ohm line. A separate 75-ohm output circuit is included which can be used to feed an additional monitor unit. The video-frequency system is provided with an interlocked switching arrangement by which any one of four

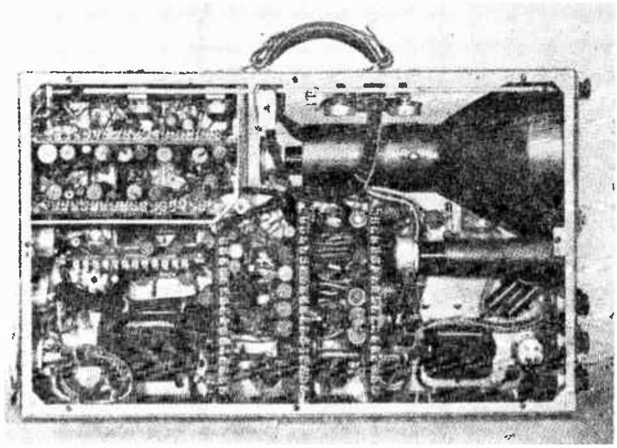

Fig. 13—Camera-control unit—circuit-component side, cover removed.

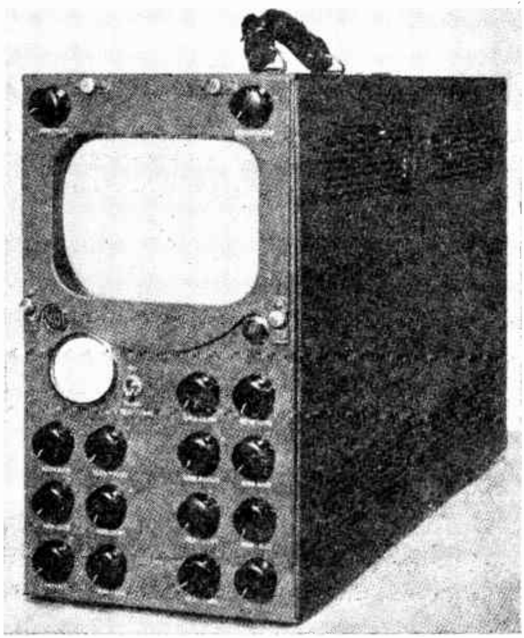

Fig. 14—Camera-control unit—front view.

input signals can be selected, amplified, monitored, and fed to the outgoing line. Indicator lights on both the master-control unit, each camera-control unit, and camera show which camera is "on the air."

The deflection system for the master-control unit employs synchronizing and deflection circuits which are essentially the same as those used in television receivers. The picture observed on the kinescope in

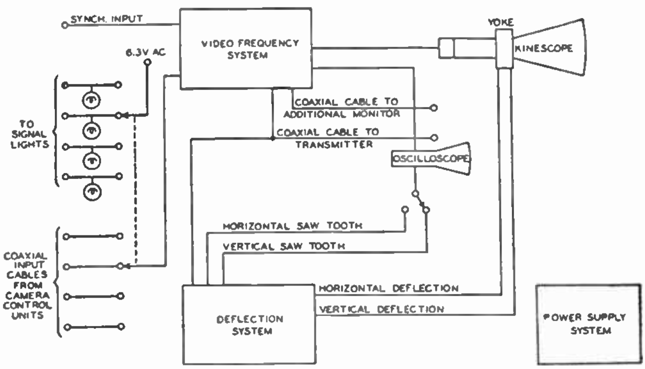

Fig. 15—Block diagram of master-control unit.

the master-control unit is therefore an indication of the performance to be expected at the receiving locations.

The low-voltage and high-voltage rectifiers used in the master-control unit are similar to those included in the camera-control unit. Figure 16 is a front view of the master-control unit and shows the switching and indicator-light arrangement.

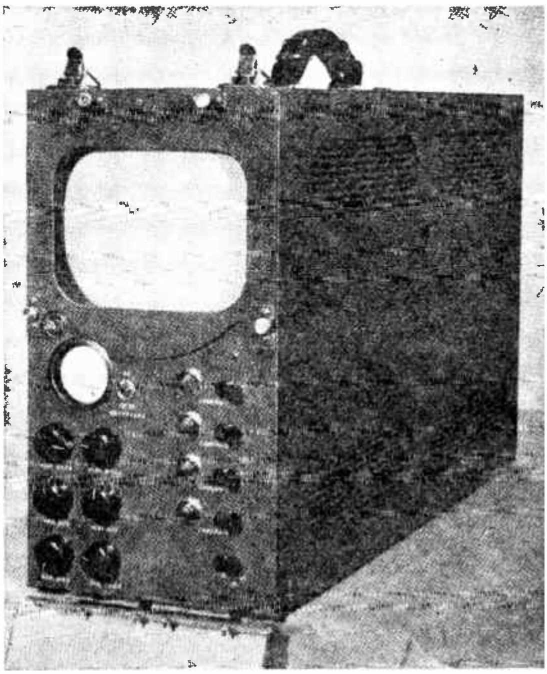

Fig. 16—Master-control unit—front view.

Synchronizing Generator

The portable synchronizing generator is designed to produce pulses in accordance with the standards of the Radio Manufacturers Association. The synchronizing generator is divided into three units, the pulse unit, the shaping unit, and the delay unit.

Pulse Unit

The pulse unit contains an electromechanical pulse generator for producing 26,460- and 60-cycle pulses. This type of pulse generator was used because it gave the desired electrical characteristics with a minimum of equipment. This generator consists of a brass disk having 441 peripheral teeth and rotated by a 3600-revolution-per-minute synchronous motor. This disk revolves inside a stationary brass ring having 441 teeth on its inner circumference. The clearance between the teeth on the rotor and stator is approximately 0.012 inch. A single radial fin is used on the rotating disk in conjunction with a similar stationary fin to produce the 60-cycle pulses. Direct polarizing voltage is applied between the stators and rotating disk through resistors. The current variations through the resistors in accordance with the changes in capacitance produce the desired voltage pulses. The use of a large number of teeth on both the rotor and stator minimizes the effect of inaccuracies in the width of the teeth and the spacing between them since each pulse is produced by the average change in capacitance caused by each of the 441 teeth on the rotor and stator. In addition

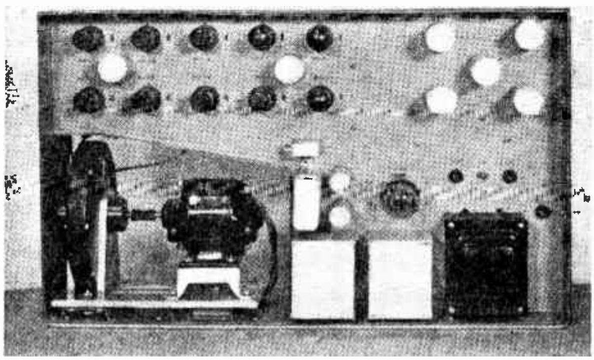

Fig. 17—Synchronizing-generator pulse unit—tube side, cover removed.

to the pulse generator the pulse unit contains tubes and associated circuits for shaping the pulses and for obtaining pulses at line frequency by selecting every other one of the 26,460 pulses. A power-supply rectifier to provide anode potentials for the complete synchronizing generator is also included in the pulse unit. Figure 17 shows the tube side of the pulse unit. The pulse generator is shown in the lower left-hand corner of the figure. The rotor and stator are completely surrounded by a bakelite housing.

Shaping Unit

The shaping unit receives 60-, 13,230-, and 26,460-cycle pulses from the pulse unit. Four sets of pulses are provided by the shaping unit as follows:

Iconoscope horizontal driving pulses
Iconoscope vertical driving pulses
Blanking pulses
Synchronizing pulses

Although the synchronizing pulses are formed by combinations of several pulses the leading edge of each pulse is the leading edge of a 26,460-cycle pulse. Controls are provided for varying the width of the several pulses. Photographs of the shaping unit with the side covers removed have been previously shown in Figures 2 and 3.

Delay Unit

When two or more cameras are used and they are connected to the

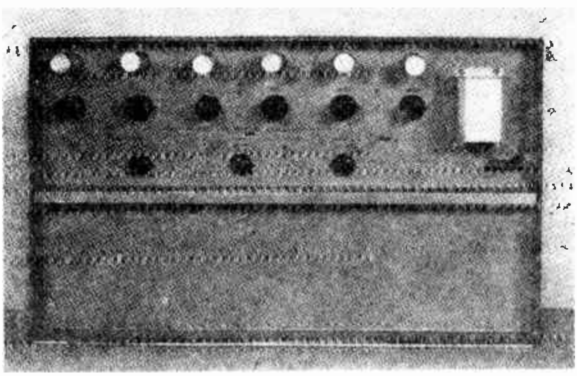

Fig. 18—Synchronizing-generator delay unit—tube side, cover removed.

control equipment through cables which differ greatly in length it is necessary to delay the driving pulses to the camera connected to the shortest cable so that the pulses returning to the control equipment from this camera correspond in time with those returning from the camera connected to the longest cable. The synchronizing-generator delay unit contains an artificial line which is used to delay the driving pulses to any one of three cameras by an amount corresponding to any normal length of camera cable up to 500 feet. Buffer tubes are used between the switches and the artificial line so that the characteristics of the line are not affected by the various lengths of camera cable. Swich positions are provided for 50-, 100-, 200-, 300-, 400-, and 500-foot lengths of cable. Figure 18 is a photograph of the tube side of the delay unit and shows the switch knobs for varying the delay for each camera. The photograph shown in Figure 19 illustrates the construction of the artificial line and the circuit components used with the buffer tubes. The space in the bottom of both sides of this unit is used to carry spare tubes.

Relay Transmitter

One of the requirements of any portable pickup system is that some provision must be made for conveying the signal from the remote point to the location where it can be utilized. The wide frequency band used in television makes this problem especially difficult. One obvious solution is of course a portable transmitter for relaying the signal from the remote pickup point to a suitable receiving location near the main transmitter. An ideal transmitter for this type of work must be reasonably rugged, light in weight, and deliver sufficient power to

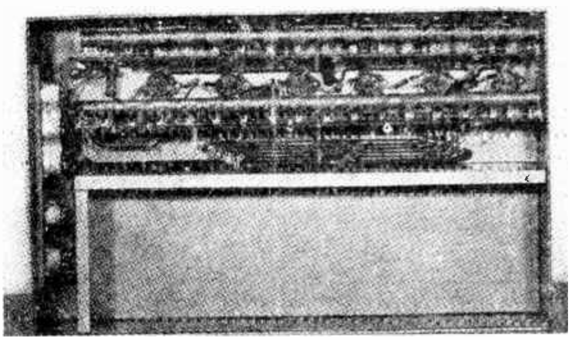

Fig. 19—Synchronizing-generator delay unit—circuit component side, cover removed.

provide a satisfactory service range. The ultra-high-frequency television relay transmitter was designed to meet these requirements. This transmitter is crystal-controlled and will deliver a peak power of 25 watts at any specified frequency between 280 and 340 megacycles. The radio-frequency portion of the transmitter consists of four stages; crystal oscillator, two multiplier stages, and the power-amplifier stage. Two neutralized 1628 triodes are used in the power amplifier. All the circuits which are resonant at carrier frequency are "transmission-line circuits." All the other radio-frequency circuits are conventional *L-C* circuits.

The video-frequency portion of the transmitter consists of three
and 5 megacycles. An input of 2 volts, peak to peak, is sufficient to
stages which are adjusted to pass the frequency band between 30 cycles

grid-modulate the power-amplifier stage completely. The direct-current component of the video-frequency signal is restored in the grid circuit of the modulator stage, and direct-current coupling is employed between the modulator plate and the power-amplifier grids.

The monitoring system in the transmitter consists of a diode rectifier, a video-frequency amplifier, and a two-inch oscilloscope. Provision is made so that the output of the video-frequency amplifier can

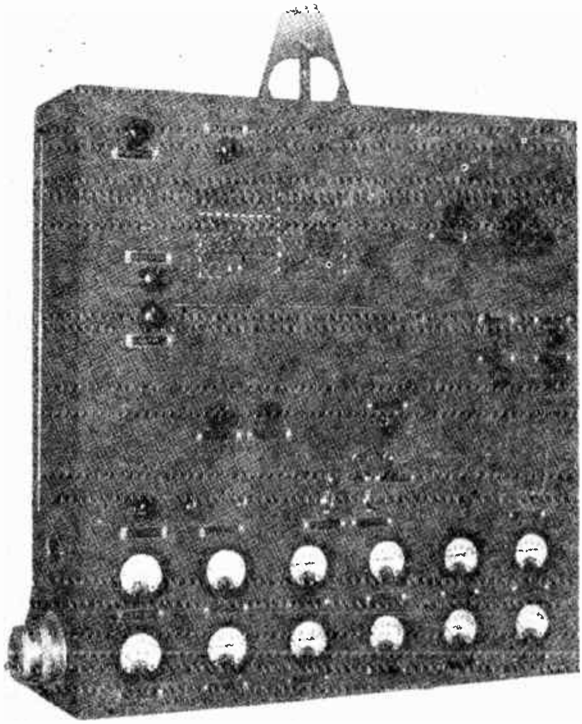

Fig. 20—Ultra-high-frequency relay transmitter—front view.

be fed to a master-control unit so that the complete picture can be monitored.

The transmitter output system is so arranged that either a coaxial line or balanced feeder system may be used. Small antennas having high directivity are readily obtainable at the transmitter frequency. A unidirectional array using eight half-wave elements has given a measured power gain of 12 in field tests.

Figure 20 is a front view of the transmitter. The overall dimensions of this unit are $6 \times 24 \times 26$ inches. The antenna transmission-line clamping unit is shown extending from the top of the transmitter unit. The monitoring oscilloscope is viewed through the circular opening in the upper right-hand corner of the picture. The meters at the bottom of the unit indicate the currents in the various tubes.

The rear view of the transmitter with the doors open is shown in Figure 21. The location of the various circuit components, tubes, and transmission lines can be seen in this photograph.

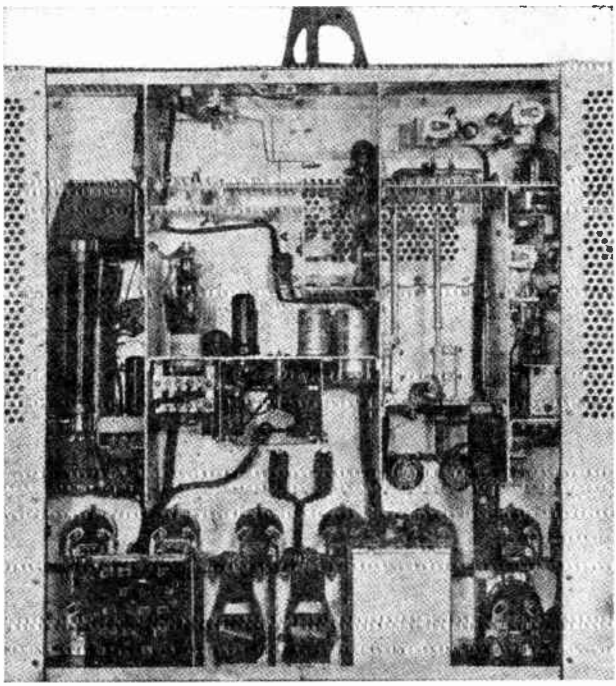

Fig. 21—Ultra-high-frequency relay transmitter—rear view, doors open.

Transmitter Power-Supply Unit

This unit contains two rectifier systems which furnish all the direct voltages necessary for the operation of the transmitter. Figure 22 shows a view of this unit.

Relay Receiver

The relay receiver is a superheterodyne designed to operate from a 150-ohm antenna transmission line. Coupled fixed-tuned "transmission-line" circuits are used in the input system of the receiver. These

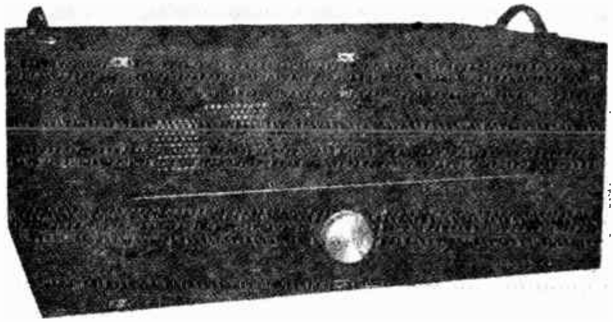

Fig. 22—Ultra-high-frequency relay transmitter power-supply unit.

circuits are adjusted to pass a frequency band of 12 megacycles in the range between 280 and 340 megacycles. The oscillator circuit is also of the "transmission-line" type. The intermediate-frequency amplifier consists of seven transformer-coupled stages. The second-detector circuit is direct-current coupled to the automatic-gain-control rectifier, so that the automatic-gain-control voltage is proportional to the peak value of the incoming video-frequency signal, i.e., synchronizing peaks. Provision is made for disconnecting the automatic gain control and using manual gain control if desired. The video-frequency amplifier supplies an output of about 2 volts, peak to peak, across a 75-ohm line. Figures 23 and 24 show the front and rear views of the receiver. The front view of the power-supply unit is given in Figure 25.

Practical Applications of the Equipment

Television programs have been broadcast as a public service during

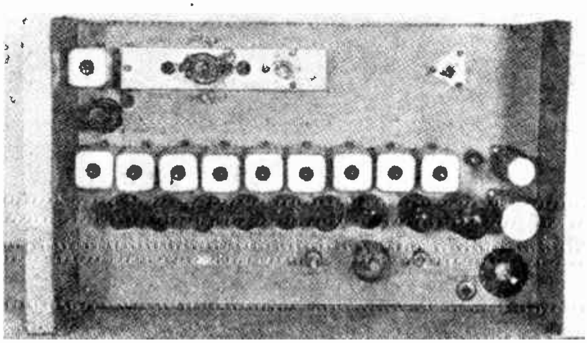

Fig. 23—Ultra-high-frequency relay receiver—front view.

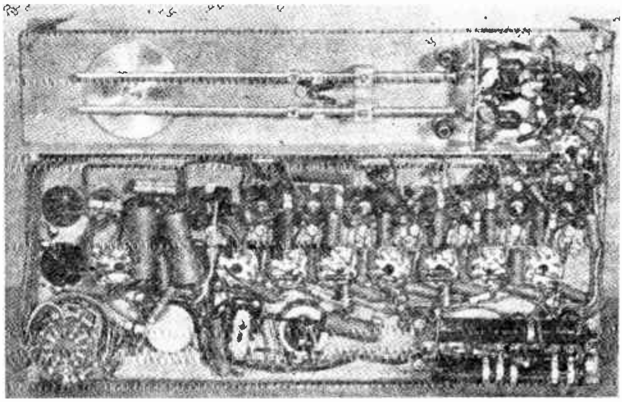

Fig. 24—Ultra-high-frequency relay receiver—rear view.

the past year. The telemobile units previously mentioned have been used in many "on-the-scene" pickups and these have almost all been very popular. The pickup of many potentially interesting programs has been impracticable because of the size, weight, and power requirements of the mobile units. Although the cameras associated with these units can be operated at distances up to 500 feet from the unit housing the control equipment, this in many instances is not sufficient because the control unit cannot be placed in an advantageous location. The alternating-current input power, especially the three-phase for the transmitter unit, frequently has been very difficult to obtain.

The new "suitcase" type of portable pickup equipment, therefore, greatly increases the program potentialities outside the studio. The size, weight, power requirements, and flexibility of the equipment are

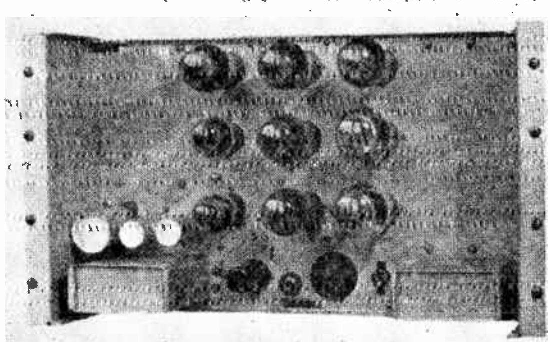

Fig. 25—Ultra-high-frequency relay receiver power-supply unit.

such that for the first time program pickups aboard airplanes, boats, and automobiles, while in motion, are possible. The program possibilities which the equipment creates are thus evident for it is obvious that the extension of the television eye to points outside the studio is an even greater boon to television than was the corresponding extension of the microphone in sound broadcasting. The new equipment has recently been used aboard an airplane to pick up scenes of New York and transmit them to Radio City for a program broadcast from the transmitter in the Empire State Building.

It is quite possible to carry the complete pickup apparatus into any building, amusement park, theater, etc., in order to televise events which are inaccessible to pickup equipment mounted permanently in a truck. The few kilowatts of single-phase alternating-current power which are required for the entire equipment can be obtained in most locations.

Another important application of the new equipment is the televising of regular sound broadcast programs in the studios in which they are normally presented. Although such use does not provide all the flexibility of a studio permanently equipped for television, it does permit a very useful extension of pickup facilities for certain types of programs.

The portability of the transmitter is a great advantage in remote pickup work. In many locations it is necessary to erect an antenna on the roof of a building or some other high structure in order to obtain line-of-sight transmission to the receiving point. In the case of a transmitter mounted permanently in a truck a rather long radio-frequency transmission line is required. The problem of adjusting such a line to carry television signals without serious reflections is a difficult one, even in a permanent installation and for portable or mobile work the difficulties are still greater. The new equipment makes it possible to locate the transmitter on the roof or upper floor of a building so that a short radio-frequency transmission line can be used to the antenna. In this case the transmitter is connected to the pickup equipment by means of a flexible coaxial cable or other video-frequency line. The transmission of the video-frequency signals over a suitable line presents a considerably less serious problem than the transmission of radio-frequency power over a line of equivalent length.

Acknowledgments

The authors wish to acknowledge the individual and co-operative efforts of the many engineers who participated in the development of this equipment.

ANALYSIS AND DESIGN OF VIDEO AMPLIFIERS*†

By

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Part II

Introduction

THIS paper is an extension of a previous one[1] which contained an elementary discussion of the factors influencing the performance of wide-band video amplifiers. The principal items of interest were an analysis of the effects of variable time delay and amplitude response on the reproduced picture, and a discussion of one method of high-frequency compensation. Notes on certain measurement techniques, and a discussion of low-frequency requirements were also included.

It is sufficient, in the present case, to repeat, as explained in the previous paper[1], that the ideal video amplifier should have flat frequency response and constant time delay over the band of frequencies required for adequate reproduction of the transmitted picture.

The importance of maintaining the characteristics of individual video stages as close to the ideal values as possible is accentuated in cases where numerous stages are connected in cascade. This is true because the overall gain is equal to the product of the individual stage gains, while the net time delay is equal to the sum of the time delays of the individual stages. It is interesting to note that this applies particularly to video-amplifier chains, where thirty or more stages may be used in a television transmitter.

It may be said, in general, that it is quite difficult to maintain both time delay and gain constant over a wide band. Generally a compromise is made, with neither the gain or delay exactly constant, but with both satisfactorily close to optimum values.

Furthermore, the correction expedients which are applicable at one end of the video band have no effect at the other end. Thus, the use of peaking coils for maintenance of constant gain at high frequencies does nothing whatsoever to the low-frequency performance.

Because of this segregation of the video band into two distinct regions, it is desirable to treat the high- and low-frequency characteristics as separate problems. This procedure will be followed in this

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article, with the first section containing a discussion of the high-frequency aspects of the problem, and the last section dealing with the low-frequency performance of the amplifier.

SECTION I

High-Frequency Considerations

All of the video-amplifier circuits which have appeared to date consist essentially of resistance-coupled stages, each provided with some form of high-frequency gain and phase correction. The decrease in amplification and introduction of phase distortion at the higher frequencies in an uncompensated amplifier is a direct result of the existence of unavoidable shunt capacitances, which are found in any circuit containing vacuum tubes and associated components (resistors, wiring, sockets, etc.). The reactance of these shunt capacitances appears as part of the plate-circuit load, and causes its impedance to decrease as the frequency is increased, resulting in a decrease in gain at the higher frequencies. The loss in gain is accompanied by a phase delay, which, with no shunt reactance in the plate load, would normally be zero. The manner of variation with frequency of this phase delay is important in determining the constancy of time delay over the video band.

There are several ways to reduce the effect of the load-circuit capacitance. One method involves the use of a very small load resistor, whose resistance is so low compared to the reactance of the shunt capacitance at the highest video frequency that the reactance has no effect on the gain or phase characteristics. This arrangement would possess no practical advantages, because of the great loss in gain per stage entailed by the use of a small plate resistor. (The gain in a video-amplifier stage may be taken generally as the product of the tube's mutual conductance and the plate-load impedance, since pentodes are used almost universally.)

A more practical way to obtain adequate high-frequency performance is to employ a circuit containing inductance to offset the loss in gain due to shunt capacitance. In this way essentially constant gain may be obtained, without resorting to abnormal reductions in the value of plate-load resistor.

The expedients employed to extend the frequency band in which constant gain obtains are described variously as correction circuits or peaking circuits. These may take any of several forms, depending upon whether the wide-band characteristics are obtained by inserting a peaking coil in the load circuit, to maintain the load impedance at a constant value, or whether the desired effect is obtained by the use of a coupling circuit, such as a low-pass filter, between successive stages of the amplifier.

Four types of high-frequency video load circuits will be discussed here: (1) Uncompensated load circuit. (2) Compensated circuit containing a peaking coil in series with the load resistor—known as shunt peaking. (3) Compensated load circuit in which a π-type low-pass filter is employed as the coupling element—known as series peaking. (4) Combination of shunt and series peaking.

The analysis of these various types of load circuits and the evaluation of their relative merits is somewhat simplified and made more readily adaptable to direct comparison by the use of the following list of symbols and definitions.

T_1 and T_2 = Two successive tubes of a video amplifier circuit.

R_L = Load resistor in plate circuit of T_1.

C_T = Total capacitance shunting the load circuit. This includes tube and wiring capacitances.

C_1 = Total output capacitance of T_1.

C_2 = Total input capacitance of T_2.

$C_2/C_1 = m$.

L_1 = Inductance of peaking coil in series with plate-load resistor of T_1 (shunt peaking).

L_2 = Inductance of peaking coil connected between plate of T_1 and grid of T_2 (series peaking).

f_o = Top frequency in the video band.

f = Any frequency in video band above 1 kc.

Φ = Phase delay in radians (caused by reactance in plate-load circuit).

$T = \dfrac{\Phi}{\omega}$ = Time delay in seconds (due to reactance in plate-load circuit).

Δ_T = Departure from constant time delay (seconds)

It should be noted at this point that, in general, maintenance of a flat frequency-response characteristic (at high frequencies) in a video amplifier stage usually will result in sufficiently uniform high-frequency time delay so that correction expedients which might be applied to produce an entirely uniform delay (and which might alter the response somewhat) are not usually necessary or desirable. This, of course, depends largely on the total number of stages in cascade to be employed for a given purpose, for, as pointed out previously, the overall gain characteristic is the product of the individual stage gains, whereas the total time delay is the sum of delay characteristics of each stage.

In this connection it is important to observe that the high-frequency performance of the amplifier determines the quality of the picture along any horizontal line, i.e., the horizontal detail and resolution. If both gain and delay characteristics are flat the picture is reproduced exactly. If the gain is constant in the video band and the time delay varies with frequency, all the high-frequency components are reproduced precisely in their proper relative amplitudes, but the location of the various picture elements is not correct, because of the different amounts of time taken for passage of the different frequencies. This results in inferior reproduction of horizontal detail.

It is difficult to determine precisely the maximum permissible variation in time delay in a complete television system. Some authors[2] suggest limiting the total variation in time delay to 0.1 μsec. Data calculated for a typical case shows that, for a 441-line picture (10-inch horizontal dimension on a 12-inch tube) a variation in time delay of 0.1 μsec. up to a top frequency of 2.5 Mc would cause the 2.5 Mc component to be displaced laterally by 0.015-inch with respect to the low-frequency components. This would amount to about one picture element (at 2.5 Mc) in the horizontal direction. It should be obvious that, for a given total tolerable time-delay variation, the permissible departure from constant delay-time per stage decreases as the number of stages increases, since the total delay variation is the sum of the individual-stage delay variations. Therefore, in receivers, where at the most, only three or four video stages would be employed, the permissible variation in time delay per stage is greater than in cases where a large number of stages are used in an amplifier chain. The line amplifiers in television transmitters fall in the latter category. Furthermore, in receivers, the delay variations in the i-f circuits are generally much greater than the delay variations in the video amplifiers, consequently attention is generally directed toward minimizing departure from constant delay in the i-f circuits. The procedure in designing video amplifiers for receivers consists principally, therefore, in obtaining a flat gain characteristic over the video band, while the time delay is permitted to depart from a constant value, within reasonable limits.

The magnitude of the time-delay variations, Δ_T (departure from the desired constant value) in an amplifier stage may be written in a number of ways. One method of expressing Δ_T, which will be used in this article, evaluates the departure from constant time of transmission as a fractional part of a period at the top video frequency, i.e., $\Delta_T = K/f_o = KT_o$.

Circuit 1—Resistance-Coupled Video Amplifier Utilizing No High-Frequency Peaking Expedients

The plate load Z_L comprises the load resistor R_L in parallel with the total shunt capacitance, C_T. The gain, which is equal to $Gm\ Z_L$, falls off as the frequency is increased according to

$$\text{gain} = Gm\ Z_L = \frac{Gm\ R_L}{\sqrt{1 + (2\pi f C_T R_L)^2}}$$

If we let R_L equal the reactance of C_T at the top frequency, f_o, we have

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

and

$$\text{gain} = \frac{Gm\ R_L}{\sqrt{1 + (f/f_o)^2}}$$

At this frequency where $R_L = \dfrac{1}{2\pi f_o C_T}$, the gain is 70.7 per cent of the gain at low frequencies ($f = 10$ kc, for instance). The departure from constant time delay at f_o is $0.034/f_o$, i.e., 3.4 per cent of the period T_o at the top frequency, f_o. With $f_o = 3$ Mc, Δ_T is 0.011 microseconds. This is the difference in time delay caused by the presence of shunt capacitance in the plate-load circuit.

It should be evident that the gain of this type of load circuit is not sufficiently constant to permit its use in a video amplifier, unless the load resistor is made small compared to the total shunt-load reactance at the top video frequency.

While this analysis is included primarily to demonstrate the behavior of an uncompensated circuit, and as a basis of comparison for other compensated circuits to follow, it can be put to use as a means for measuring the total load-circuit capacitance of a video stage. The method, described in detail in Part I of this article, makes use of the fact that the gain of an uncompensated stage falls to 70.7 per cent of its low (10 kc) frequency value at a frequency for which the reactance of the capacitance in question is equal to the plate-load resistance. The measurement of the point of 0.707 response may be determined by noting the frequency at which the input to the stage under test must be increased to $\sqrt{2}$ times its low-frequency value, to maintain constant stage output.

The indicating device may include the following tube in the chain, which should have a low (100-ohm) resistor connected in its plate circuit to provide a voltage drop which can be read on a vacuum-tube voltmeter. The bias of this second tube should be maintained at its operating value, to preclude any error due to input capacitance variation with bias. A variation of this connection applies the vacuum-tube voltmeter across the load resistor R_L (with the following tube in circuit) and measures the output across R_L. The capacitance contributed by the vacuum-tube voltmeter must be known and taken into account in this measurement.

The total output-circuit capacitance can also be measured by a substitution method, in which a "Q" meter may be employed. A coil is selected to resonate with 100 $\mu\mu f$ or so on the "Q" meter at some frequency between 500 and 2000 kc. The circuit is resonated and then the capacitance terminals of the "Q" meter are connected across the output circuit of the stage under test. The plate-load resistor is removed and the plate-supply voltage of T_1 is turned off. The second tube operates normally, with its bias fixed at the operating point. The amount by which the "Q"-meter calibrated-capacitance must be changed, in order to re-establish resonance in the "Q"-meter circuit, is equal to C_T, the total shunt capacitance in the video stage.

Note that the resonant voltage which appears across the "Q"-meter tank circuit must be limited in amplitude to prevent rectification in the second tube's grid circuit, which would result in a change in bias and in second-tube input capacitance.

CIRCUIT 2—VIDEO STAGE COMPENSATED BY A COIL IN SERIES WITH THE LOAD RESISTOR (SHUNT PEAKING)

This type of video stage may be compensated (the plate load-circuit impedance made essentially constant over the required frequency band) by inserting a properly proportioned inductance in series with the load resistor. The peaking-coil inductance is determined by the values of R_L, C_T, and the top video frequency, f_o.

R_L is chosen to equal the reactance of C_T at the top frequency, f_o, (C_T is measured with L_1 not in circuit, by either of the methods previously described). Therefore,

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

The value of L_1 is determined from $2\pi f_o L_1 = \frac{R_L}{2}$ at the top frequency, f_o. Hence,

$$L_1 = \frac{R_L}{4\pi f_o}$$

The resonant frequency of L_1 and C_T is seen to be $\sqrt{2}$ times the top video frequency, f_o.

The gain is essentially constant, up to the frequency f_o, and is equal to GmR_L.

The time delay, in terms of f and f_o is

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

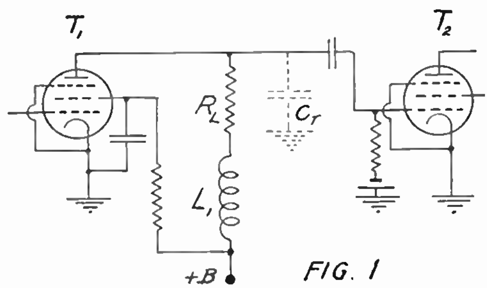


FIG. 1

The difference in time delay over the video band (from 1 k_c to f_o) is $0.0231/f_o$ seconds. With $f_o = 3$ Mc, this time-delay departure corresponds to 0.0077 microseconds. Note that the gain over the video band, GmR_L, is equal to the gain which would be experienced in an uncompensated stage having zero shunt-load capacitance, and the same value of plate-load resistor as is used here. This improvement in gain characteristic is achieved with no increase in the departure from constant time delay; in fact an improvement of 30 per cent in the approach to constant time delay has been obtained.

The values selected for compensating the circuit

$$R_L = \frac{1}{2\pi f_o C_T} \text{ and } L_1 = \frac{R_L}{4\pi f_o}$$

are not necessarily productive of the best phase and amplitude response. Other authors[2] have shown that more nearly constant time delay and amplitude response may be obtained by using slightly different values of R_L and L_1.

If we designate the ratio of load resistance, R_L to capacitive reactance X_c at the top frequency by p, and the ratio of inductive to capacitive reactance at f_o by s, we have

$$p = \frac{R}{X_c} = 2\pi f_o C_T R_L$$

$$s = \frac{X_L}{X_c} = (2\pi f_o)^2 L_1 C_T$$

The values chosen in the preceding case are $p = 1.0$ and $s = 0.5$. If, instead, we use $p = 0.85$ and $s = 0.3$, the time-delay curve is almost precisely flat, and the gain variation over the frequency band is slightly less than in the case previously described. However, this latter arrangement entails the use of a lower value of load resistor, so that the gain is decreased 15 per cent at all frequencies.

As a typical case, consider a video amplifier employing Type 1851 tubes. The total load-circuit capacitance ($C_{in} + C_{out}$ plus wiring and strays) is about 25 $\mu\mu f$. Let the top video frequency be 3 Mc, in which case $X_c = 2120$ ohms. If $p = 1$ the load resistor would also be 2120 ohms, and the coil inductance (for $s = 0.5$) would be

$$\frac{2120}{2 \times 2\pi f} = 56 \ \mu h$$

The results would be satisfactory on a basis of constant gain and time delay, with the actual gain equal to 19 per stage, for a tube having a mutual conductance of 9000 $\mu mhos$.

Use of $p = 0.85$ and $s = 0.3$ would require a resistor of 1800 ohms and a coil inductance of 33.5 microhenries, with a gain of about 16 per stage. In general, in practical cases, the value of L_1 may vary somewhat from the prescribed values, as may also the value of R_L. When several stages are used in cascade it is very important, however, to obtain as nearly uniform gain and time-delay curves as possible for each stage, since, if L_1 and R_L are not chosen according to the prescribed values, it will be difficult to make counteracting changes in alternate stages to produce an overall characteristic which does not depart too much from the theoretical values (i.e., over and under compensation in successive stages).

CIRCUIT 3—π-TYPE LOW-PASS FILTER EMPLOYED AS COUPLING ELEMENT BETWEEN PLATE OF T_1 AND GRID OF T_2 (SERIES PEAKING)

This type of circuit possesses certain advantages over the shunt-peaking arrangement because it effectively separates C_2 from C_1 by means of L_2. (They would normally be in parallel in the shunt peaking circuit, appearing across R_L and L_1.) It affords a greater gain per stage with a smaller departure from constant time delay. The action of the circuit in preserving the high-frequency response of the amplifier may be described briefly as follows: A voltage $eg\,GmR_L$ is considered to exist across R_L (with C_1, L_2 and C_2 removed). C_1 is next considered to exist across R_L, which causes attenuation of the higher frequencies and produces a voltage

$$\frac{eg\,GmR_L}{\sqrt{1 + (2\pi f C_1 R_L)^2}}$$

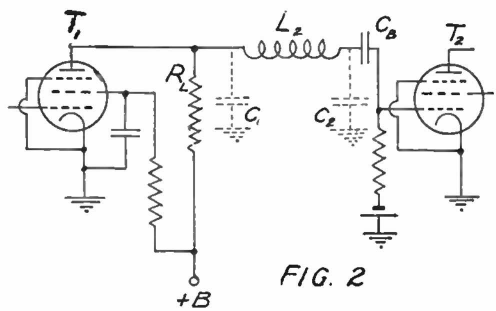


FIG. 2

across R_L and C_1, in parallel. This voltage is applied to the voltage divider consisting of L_2 and C_2 in series, and the resultant drop across C_2 is maintained constant by resonant rise effects in L_2C_2, which counteract the attenuation produced by C_1.

The performance of the circuit depends upon a number of factors. One of these is the ratio of the two capacitances, C_1 and C_2, which appear at the terminals of the low-pass filter. Let this ratio of C_2/C_1 be m. C_1 includes the output and stray wiring capacitances associated with tube No. 1. C_2 includes the input and wiring capacitances of tube No. 2, as well as the stray capacitance between the blocking condenser, C_B, and ground. Note, from Figure 2, that the blocking condenser may be connected at either end of L_2 to assist in adjustment of the value of m.

The value of total capacitance $(C_1 + C_2)$ may be determined experimentally by the methods described for use with the shunt-peaking circuit (L_2 is shorted in this measurement). To measure C_1 open L_2

and find the frequency at which the gain of T_1 is 70.7 per cent of its low-frequency value. A vacuum-tube voltmeter of known input capacitance may be used across R_L as an indicating device, and its contribution to C_1 must be taken into account.

An alternative method of measuring C_1 makes use of the "Q" meter, as described in connection with shunt-peaking circuits.

It has been pointed out by Albert Preisman of R.C.A. Institutes that, for best performance, C_2 should be at least twice C_1, i.e. $m \geqq 2$. This condition is fulfilled in most practical applications. If m is found to be less than 2, the ratio may be adjusted by proper disposition of the circuit components (such as the d-c blocking condenser for T_2) or by connecting small capacitances across the filter input or output terminals. It should be noted, however, that the use of additional capacitance at either end of the filter to produce the desired value of m will result in a loss in gain, for the absolute gain is inversely proportional to the total capacitance in the load circuit.

With C_2 and C_1 known from measurement, the first step in designing the coupling network is to select f_o, the top frequency in the video band for which constant gain is desired. This value of f_o, in conjunction with C_1 determines the inductance L_2 of the series-peaking coil.

To find L_2, let f_r be the resonant frequency of L_2 and C_1, i.e.,

$f_r = \dfrac{1}{2\pi\sqrt{L_2C_1}}$. The value f_r is chosen to be $\sqrt{2}$ times the top video frequency, f_o. Therefore, L_2 is determined from

$$L_2 = \frac{1}{2(2\pi f_o)^2C_1}$$

The inductive reactance of L_2 at f_o is equal to one-half the reactance of C_1 at the same frequency.

The value of R_L, the plate-load resistor, with $m = 2$ and $f_o = f_r/\sqrt{2}$, is equal to one-half the reactance of C_1 at the top video frequency, f_o. Since $m = 2$, R_L also equals one and one-half times the total load-circuit, capacitive reactance at f_o.

The procedure for compensating a stage may be itemized as follows: (1) Measure C_1 and C_2 and, if necessary, adjust C_2/C_1 to be at least 2; (2) make the terminating resistor R_L equal to one and one-half times the reactance of $(C_1 + C_2)$ at the top video frequency, f_o and connect the resistor across the plate end of the filter network; (3)

obtain a coil which resonates with C_1 at $\sqrt{2}$ times the top video frequency, or use the relation $L_2 = \frac{2}{3}(C_1 + C_2)R_L^2$. The resistance of coil L_2 is immaterial as long as the coil Q is greater than 20.

Under some conditions it might be necessary to work out of a high plate-circuit capacitance into a low grid capacitance. In such a case the value of C_2/C_1 may be more nearly ½ instead of 2. In that event, the values of L_2 and R_L are the same as those calculated for $m = 2$, but the load resistor is connected across the output terminals of the network, i.e., across the smaller terminating capacitance. A reciprocal action permits interchanging the point of resistor termination in this special case, and results in operating characteristics which are the same as for the more likely case discussed previously. Specifically, the coupling network may be turned end for end without affecting its operation.

The basic design equations, to be used for any value of m, with the top video frequency chosen to be 0.707 times the resonant frequency of L_2 and C_1, are

$$R_L \doteq \frac{1}{\sqrt{2m}\,\omega_O C_1}$$

$$L_2 = \frac{1}{2\omega_O^2 C_1}$$

where $\omega_O = 2\pi$ times the top video frequency. If the values suggested above $\left(R_L = \frac{3}{2}\,\frac{1}{(C_1 + C_2)\omega_O} \text{ and } L_2 = \frac{2}{3}(C_1 + C_2)R_L^2\right)$ are used in the video stage, the gain and time-delay characteristics are essentially flat out to f_o. The absolute value of gain is 50 per cent greater than the gain experienced in a shunt-peaking circuit having the same total load-circuit capacitance and the same value of f_o. The departure from constant time delay is $0.0113/f_o$ seconds. For a 3-Mc band the variation in time delay is 0.004 microseconds, which is somewhat smaller than Δ_T in the shunt-peaking case. The total time delay is greater with series peaking, but, of course, this is relatively unimportant, since, within reason, the magnitude of T is of no consequence, provided the departure from a constant value is small.

The series-peaking circuit merits serious consideration on the basis of these results. It may be expected to exceed the shunt-peaking circuit in performance in cases where the capacitance distribution is favorable, or when the ratio of capacitances can be adjusted to the desired value without causing a decrease in gain below the value experienced with shunt peaking. Note that operation with values of m less than 2 will cause the gain characteristic to peak at the high end. While this effect is not desirable generally, it may find some utility for peaking purposes in amplifiers in which the high-frequency gain in other stages of the chain is deficient. Such a condition might exist by virtue of the high-frequency attenuation experienced in a concentric transmission line, wherein a drop at the top end of the video band must be overcome by subsequent peaking stages.

CIRCUIT 4—COMBINATION OF CIRCUIT 2 AND CIRCUIT 3

This circuit provides certain advantages over either No. 2 or No. 3 used singly. As described by E. W. Herold[3] it has the following char-

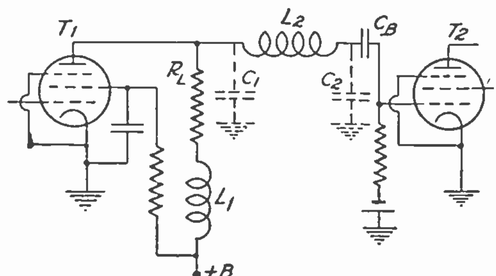


FIG. 3

acteristics: for a given total load-circuit capacitance C_T and prescribed top video frequency f_o, the load resistor which may be used (maintaining constant gain up to f_o) is approximately 80 per cent greater than in the case of simple shunt peaking. This means, of course, 80 per cent higher gain per stage, for the gain is $Gm\ R_L$, when the circuit is properly compensated. The departure from constant time delay is roughly equal to that experienced in a simple series-peaking circuit.

The disposition of circuit components required to produce the 80 per cent increase in over-all gain are as follows:

$$m = C_2/C_1 = 2 \qquad L_1 = 0.12\ (C_1 + C_2)\ R_L^2$$

$$R_L = \frac{1.8}{\omega_O (C_1 + C_2)} \qquad L_2 = 0.52\ (C_1 + C_2)\ R_L^2$$

To design a stage similar to that shown in Figure 3 the procedure is as follows: (1) Select the top frequency f_o to be passed with uniform gain; (2) make $m = C_2/C_1$ equal to 2; (3) determine the total load-circuit capacitive reactance at the top frequency; (4) choose a load resistor equal to 1.8 times this total load-circuit reactance at f_o: (5) calculate L_1 and L_2 from the formulas given above.

The following table itemizes the performance characteristics of the several types of circuits described in this section.

Cir. No.	Type of H.F. Comp.	$\frac{R_L}{2\pi f_o C_T}$	ΔT μsecs.	L_1	L_2	$\frac{C_2}{C_1}$
1	none	1	$\frac{.035}{f_o \text{ Mc}}$			
2	shunt	1	$\frac{.0231}{f_o \text{ Mc}}$	$.5C_TR_L^2$		
3	series	1.5	$\frac{.0113}{f_o \text{ Mc}}$		$.67C_TR_L^2$	2
4	shunt and series	1.8	$\frac{.015}{f_o \text{ Mc}}$	$.12C_TR_L^2$	$.52C_TR_L^2$	2

SECTION II

Low-Frequency Considerations

The presence of the unavoidable shunt capacitance in the output circuit of a video stage impairs the operation of the device only at high frequencies. Below 100 kilocycles the shunt reactances have negligible effect. Consequently, for frequencies ranging between 100 kilocycles and 200 cycles, probably all types of standard video stages perform creditably whether they are compensated or not.

In the frequency range extending below 200 cycles, the gain and time-delay characteristics of a video stage are also subjected to variations from the ideal conditions. These are caused by the inability of the d-c blocking condenser in the grid circuit, in combination with the grid leak, to pass the low video-frequency signal components with their proper amplitude and phase composition, or they may be due to inadequate by-passing of a cathode bias resistor.

It will be recalled that departure from constant gain and time delay in the high-frequency portion of the video band causes imperfect reproduction of horizontal detail. Insofar as the horizontal detail is concerned the video amplifier could be cut off at 10 kc (passing no signals below this frequency), since the lowest frequency involved in reproducing a picture along any horizontal line is equal to the line repetition rate, 13,230 cycles per second.

The function of the very low video frequencies is to supply the background of the reproduced picture. Failure of a video stage to pass these frequencies in their original wave form will generally cause the background to vary in intensity from top to bottom of the picture. As an example of this effect, consider the situation existing when an all-white screen is to be transmitted. If the low-frequency characteristics of the amplifier are inadequate, the background of the picture will be non-uniform, i.e., there will be a gradual variation in shading in the vertical direction. This effect is least pronounced when an all-white or black screen is transmitted. Maximum departure from the desired background conditions occurs when the screen is half-black and half-white, about a horizontal center line. This matter will be discussed in greater detail later in this section.

Insofar as circuit performance at the low-frequency end of the band is concerned, it can be said that maintenance of proper phase characteristics is more important than maintenance of constant gain. However, even though the phase and gain characteristics are known, it is only with considerable difficulty that the response of the system to a low-frequency pulse may be predicted; that is, the performance of a video stage at low frequencies cannot be judged readily from measurements taken on a monotone basis. Consequently, the low-frequency performance of the amplifier may be more readily evaluated on a square-wave basis. This can be accomplished experimentally by applying a low-frequency square wave and by observing the amount of distortion of the output wave form. Or, the problem can be approached analytically, as shown below.

Let Figure 4 represent a grid-coupling circuit and Figure 5 a square wave (60 cycles base frequency) to be passed through it. We wish to establish some means of determining the effect of the grid-circuit time constant on the tilt appearing in the square wave.

The voltage drop across C, due to the application of a voltage E, may be written rigorously as $E_c = E(1 - \varepsilon^{-t/CR})$, where t is the time interval following the application of E to C and R.

For relatively large time-constant circuits, in which the current through R is essentially constant for a short interval following the application of the pulse, we may write

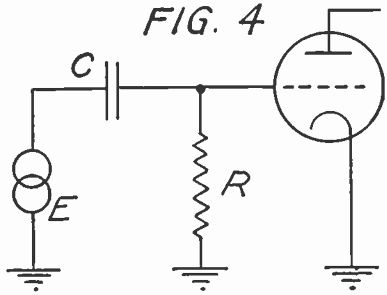
$$E_c/E = t/CR$$

$E_c/E \times 100$ is the percentage drop in amplitude of the rectangular wave during the duration of the pulse, t. E is measured from average value to peak value of the wave. Figure 6 shows the wave after having passed through the grid-coupling circuit. A 10 per cent drop in voltage amplitude is assumed for illustrative purposes.

Note that the amplitude of the pulse approaches the average value of the wave, that is, if the first pulse were allowed to decay indefinitely the total fall in voltage would not exceed the peak value of the wave. If the wave form of the pulse is changed, so that the positive and negative loops are of unequal time duration, the slope of the wave top becomes less pronounced for the long pulse and more pronounced for the short one. Equal positive and negative pulses in a square wave impose the most severe requirements on the grid-coupling circuits, for a given permissible wave-top tilt.

The formula given above may be used to advantage in determining the values of C and R in the grid circuit for a given percentage drop

FIG. 4



in amplitude of a square wave, by letting t equal the time duration of the pulse and by calculating C and R. Note that any values of C and R which produce a given time constant will result in the same low-frequency response. Practically however, extreme values of C and R should not be used, for large grid resistors will influence the d-c bias considerably in the event of grid current (due to gas or grid emission) in the tube. On the other hand, low values of R will require large values of C, which may cause an increase in total load-circuit capacitance (effective at the high frequencies) due to the stray capacitance from the physically larger blocking condenser to ground.

As an example of the application of the square-wave analysis to a grid-coupling stage, consider a square pulse, of 60 cycles base frequency, applied to a grid circuit containing a 0.25-μf blocking condenser and 0.5-megohm resistor. Since the duration of a single pulse is 1/120 second, the slope in wave top may be computed from

$$\frac{E_c}{E} = t/CR = \frac{1}{120} \times 0.25 \times 0.5 = 6.7 \text{ per cent.}$$

If, on the other hand, we calculate the relative voltage response at 60 cycles (on a monotone basis), it is found that the response at 60 cycles is better than 99.9 per cent of the response with an infinite time constant. This indicates the necessity for examining the low-frequency characteristics of a video stage on a rectangular pulse basis.

There are several arrangements available for compensating for deficiencies in the grid-coupling circuits of video amplifiers. Some involve rather complicated resistor-capacitor networks, placed in the plate and grid circuits, which provide equalization of frequency response and cancellation of phase shift down to quite low frequencies. The intricacies and mathematical analysis of these circuits will not

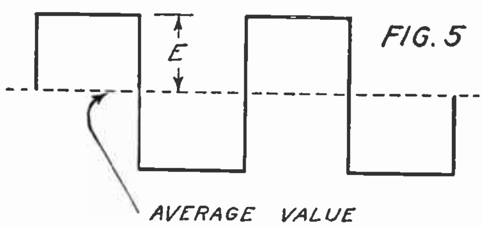


FIG. 5

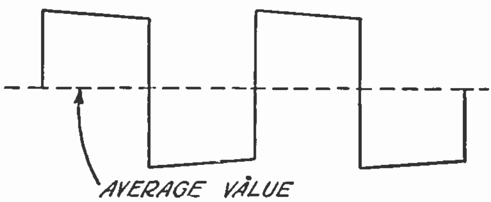


FIG. 6

be included here, for it is felt that one simple form of correction circuit, used widely in video amplifiers, and discussed here, should be all that is required for proper operation.

The simplest arrangement includes a resistance and capacitance in parallel, connected as shown in Figure 7. It can be shown that satisfactory low-frequency response can be achieved with this type of compensation provided the time constant in the grid circuit is approximately equal to the time constant of the video load resistor, R_L, and the decoupling condenser, C_F. This is very nearly true for all frequencies at which the value of R_F, the decoupling resistor, is greater than ten times the reactance of C_F. As a practical example, let R_L be 2000 ohms (video load), $C_F = 16\ \mu f$ and R_F 2500 ohms. Then the grid time constant must be equal to $R_L C_F = 0.032$ seconds. This would require a 0.25-megohm leak and only 0.125-μf grid-blocking condenser.

Note the appreciable reduction in grid-circuit time constant below the uncompensated value. The low-frequency response, in this case, is satisfactory down to 60 cycles. To extend the range to 30 cycles, the only change required is to double the size of R_F, the decoupling resistor.

This circuit has advantages over and above its low-frequency-response compensation. One of these is its filtering action against hum originating in the B supply. Another advantage, also due to filtering action, is the suppression of motor-boating tendencies at very low frequencies.

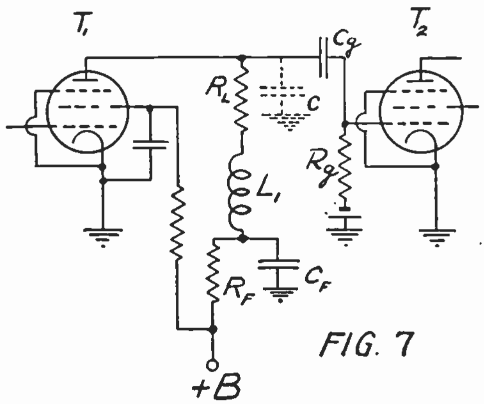


FIG. 7

The general procedure in compensating a stage at low frequencies is to select a value of C_F, say 16 μf. This, in combination with R_L, which is determined by high-frequency considerations, gives the time constant which must be obtained in the grid circuit. The decoupling resistor should be made as large as possible, consistent with obtaining the required d-c plate voltage from a normal B supply.

It should be noted that this type of plate compensation, when used to counteract deficiencies in the grid-coupling circuit, will not give perfect response down to very low frequencies (and to d.c.) except in the theoretical (and, practically, not applicable) case in which the decoupling resistor is infinite.

One of the major problems encountered in video-amplifier design is that of obtaining the required d-c grid biases. Three methods are available: (1) Battery bias; (2) bias obtained from a bleeder resistor, whose voltage is obtained from the plate power supply, generally by the insertion of a small resistor in the B-return lead; (3) cathode or self-bias. Cathode bias is, for several reasons, to be preferred. One of its advantages is that it permits the use of larger grid-leak resistors than in the case of fixed bias. A second advantage is that in involves only one small resistor and a by-pass condenser. Principally, however, its utility lies in the fact that deficiencies in the cathode by-pass condenser may be compensated in the plate circuit by the insertion of a parallel *RC* network at the low-potential end of the video load. In this case, the compensation may be made exact at all frequencies down to direct current, with practical values of circuit components. This is to be contrasted with the compensating effect of the same type of plate network when used to counteract deficiencies in the low-frequency response of grid-coupling circuits. In this case compensation is exact only for frequencies at which the feeding or decoupling resistor is very large compared to the reactance of the plate-filter condenser.

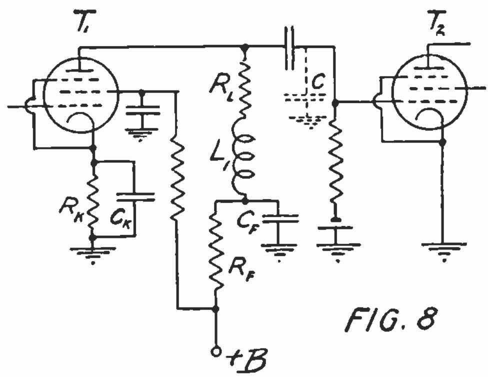


FIG. 8

The circuit for compensating for cathode-bias-network effects is that of Figure 8. In the cathode-bias case, the relation between the various resistors and condensers for compensation at all low frequencies is as follows:

$$C_K R_K = C_F R_F$$
$$R_F = R_K (Gm\ R_L)$$
$$C_F = C_K / Gm\ R_L$$

Note the presence of the GmR_L term. This product is the gain of the stage at frequencies in the mid-range of the video band. Its presence suggests a method of compensating the stage at low frequencies without a specific knowledge of *Gm* or the high-frequency gain. The procedure is as follows:

Connect the stage as shown in Figure 8 with both C_K and C_F sufficiently large to have negligible reactance at a frequency of 10 kc. Adjust the cathode bias resistor to produce the desired grid bias. Apply to the grid a voltage having a 10-kc frequency and of sufficient magnitude to produce a readable deflection on an oscilloscope or vacuum-tube voltmeter connected from plate to ground. With R_L set at its proper value to give the desired high-frequency performance note the reflection of the indicating device. This is a measure of GmR_L, the high-frequency gain. Now remove both C_K and C_F and adjust R_F (with R_K fixed at its correct bias value) to produce the same indicator deflection (constant-input volts maintained at the grid). Then shunt the cathode-load resistor with a by-pass condenser of the value to be used in circuit (say 25 μf). This being done, the only remaining step to achieve complete compensation is to shunt the decoupling-resistor R_F with a condenser which makes the time constant of the cathode circuit equal to that of the plate-filter circuit, i.e., $C_F R_F = C_K R_K$.

Values which might be employed in an 1851 video stage are:

$R_L = 2000$ ohms (depending upon video-band width)

$R_K = 150$ ohms

$R_F = 2500$ ohms

$C_K = 25\ \mu f$ electrolytic

$C_F = 1.5\ \mu f.$

It will generally not be necessary to compensate in the plate circuit of one stage, for deficiencies in grid coupling and cathode-bias circuits employed in the same stage, for the cathode-bias operation permits the use of a larger grid leak than in the case of fixed bias. This will aid in preserving the low-frequency characteristics of the grid-coupling circuit.

It is difficult to prescribe exactly the minimum values of grid-circuit time constants which may be used in a video stage, for the choice of time constant will depend upon the permissible slope of the wave tops, and upon the number of stages in the chain. Generally, in video stages containing no low-frequency plate compensation (for deficiencies in grid circuits) the time constant of each grid circuit should be from 10 to 15 times the period of the lowest frequency to be transmitted.

One should not attempt to compensate for deficiencies in a number of grid or cathode-circuit time constants in one plate-compensating network, because the results will be unfavorable unless the departure from flat-top performance on a square-wave basis is small in each stage. Best results are obtained, in a multi-stage amplifier, by compensating in each successive plate circuit.

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[1] S. W. Seeley, C. N. Kimball, "Analysis and Design of Video Amplifiers", *RCA REVIEW*, Vol. II, No. 2, October, 1937.

[2] Freeman and Schantz, "Video Amplifier Design", *Electronics*, August, 1937.

[3] E. W. Herold, "High Frequency Correction in Resistance-Coupled Amplifiers", *Communications*, August, 1938.

A PRECISION TELEVISION SYNCHRONIZING-SIGNAL GENERATOR*†

By

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Summary—The Radio Manufacturer's Association (R.M.A.) standard television synchronizing signal consists of three types of rectangular pulses in a single wave. The timing of all pulses is such that the leading edge of each pulse is either 1/26460-second or 2/26460-second from the leading edge of its adjacent pulses. The three types of pulses differ only in the duration (or width) of the individual pulses, which is not extremely critical.

In the synchronizing-signal generator described a frequency-regulated master oscillator and multivibrator generate a master wave of uniform pulses which occur at 1/26460-second intervals. This master wave as produced contains only pulses which have the width specified for the narrowest type of pulses of the R.M.A. synchronizing wave, but of course, it has many extra pulses which are not required in the R.M.A. signal. Also many of the pulses should be made wider as the R.M.A. signal contains two other types of pulses of greater widths. These two other types of pulses are produced by "lap-joining" other suitable auxiliary pulses to the trailing ends of the narrow pulses of the master wave, without altering the critical leading edges of the pulses. The particular pulses of the master wave which are undesired in the final wave are "blanked" out by other auxiliary-keying pulses.

Since some of the auxiliary-keying waves are produced by 60-cycle pulse waves, it is necessary to lock the 26,460-c.p.s. and the 60-cycle pulse waves together. This is done by a chain of three pulse-counters (acting as frequency dividers) which derive the 60-cycle wave from the master 26,460-cycle wave. Stability is obtained since the tubes involved are employed substantially as lower-resistance keys.

The advantage of frequency dividing over frequency-multiplication is discussed.

The entire chain of frequency-dividers is kept in synchronism with the 60-cycle power system by comparing the 60-cycle pulse wave to the power wave to obtain a control voltage for regulating the 26,460-cycle oscillator. The circuit used is such that the derived-control voltage is made free of 60-cycle components without sacrifice of quick response.

For economic reasons involving reliability of the transmitting system, maximum attention was given to stability and inherent accuracy of performance in all critical respects.

Introduction

MANY of the standards which are necessary to specify a television system pertain directly or indirectly to the characteristics of the synchronizing-signal generator. The shapes of the synchronizing pulses, the type of interlacing, number of lines, return periods, and picture-repetition rate, all affect the design of this

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generator. The Radio Manufacturers Association has adopted certain standards covering these points, but has not set tolerances on many of the standards. In order to obtain acceptable performance extremely high and stable accuracy is necessary in certain respects, as indicated in the description below. An entirely-electronic generator has been developed which employs circuits for obtaining high reliability and inherent precision of the output waves in all critical respects.

PRINCIPLES OF METHODS USED

In odd-line interlaced scanning as adopted by R.M.A. the horizontal-deflecting frequency, 13,230 cycles-per-second, is precisely a whole number plus one-half times the field frequency, 60 cycles-per-second. This results in interlacing since each field scan then contains a whole number plus one-half scanning lines. (This number is 220.5 lines for the R.M.A. standard 441-line television.) To meet these conditions it is essential that the generator of the 60 cycles-per-second synchronizing pulses be rigidly interlocked with the 13,230 cycles-per-second pulses. Since the two frequencies differ by such a large ratio and also do not have a whole number relation, a single-stage stable direct interlock is not feasible. The exact number of lines was chosen such as to permit interlocking by several stages where each step differs in frequency by a small whole-number ratio. A regulated master oscillator produces 26,460 cycles-per-second signal for driving a frequency-divider circuit producing half that frequency, i.e., 13,230 cycles-per-second. Similar frequency-divider circuits operating in cascade also divide the frequency of the 26,460 cycles-per-second oscillator in whole-number odd steps of 7, 9, and 7 producing frequencies of 3,780, 420, and 60 cycles per second, respectively.

For several reasons it is desirable (though not essential) that the nominal 60-cycle-per-second output of the synchronizing generator be accurately synchronized with the main power system of the community being served by the transmitter.[1] Due to the cost of filtering and shielding, television receivers will generally have some residual 60-cycle and 120-cycle ripple in their deflection systems and beam-modulating amplifiers. If these spurious influences are synchronous with the picture deflection they are much less annoying as the small power-frequency waves of displacement of the picture subject and the modulation shadows will not move vertically over the screen. Also when motion picture films are used as program material, the film projector should be synchronous and phased with the iconoscope-deflecting system within approximately 4 degrees. This condition is conveniently

[1] R. D. Kell, A. V. Bedford, and M. A. Trainer, "Scanning Sequence and Reptition Rate of Television Images", *Proc. I.R.E.*, Vol. 24, pp. 559-576; April, 1936.

obtained by driving the projector with a synchronous motor on the 60-cycle power system. Then in order to keep the "60-cycle" signal produced synchronous with respect to the local 60-cycle power supply, the two are compared in a special improved circuit which produces a controlling voltage for regulating the frequency of the 26,460-c.p.s. master oscillator.

An inspection of the R.M.A. standard synchonizing-signal in Figure 1 shows that it consists of various time-mixtures of 26,460 and 13,230 cycle-per-second pulses. The 26,460-cycle pulses are of two kinds and occur in small groups at regular sixtieth-second intervals. Each group consists of six narrow "equalizing" pulses, six much wider pulses and six more narrow "equalizing" pulses, all occurring in the order named. (The six "wider" pulses mentioned, acting as a unit, comprise a single serrated "vertical" or field synchronizing pulse.)* Each such group occupies only about 4 per cent of the one-sixtieth second.

The remaining 96 per cent of the time is occupied by the normal 13,230-cycle-per-second horizontal or "line" synchronizing pulses. Their width is $0.08H$ (where H is 1/13230 second) which is twice as wide as the equalizing pulses.

When used in the television receiver the leading edges of these pulses by their abrupt rise cause "firing" of the horizontal-deflecting oscillator. The duration of the pulses or shape of the trailing edge of the pulses do not appreciably affect the horizontal-deflecting circuit. According to the standard the 26,460-per-second equalizing pulses and the serrated-vertical pulses have alternate rising leading edges which are timed with the 13,230-per-second pulses such as to provide continuous uniform rising edges at intervals of 1/13,230 second. These rising edges provide horizontal synchronization in the receivers which is uninterrupted by the vertical synchronizing pulses.

In one of the methods tested while developing the synchronizing generator, the three different kinds of pulses which comprise the entire synchonizing wave were generated continuously in separate circuits. The output of each circuit was then keyed by an amplifier which was intermittently driven to cut-off by certain "keying" waves. The final wave was obtained by adding the several keyed outputs. This simple method had a serious fault in that permanently accurate relative timing of the three types of pulses was not obtained except by frequent adjustment. The three circuits which generated the three kinds of

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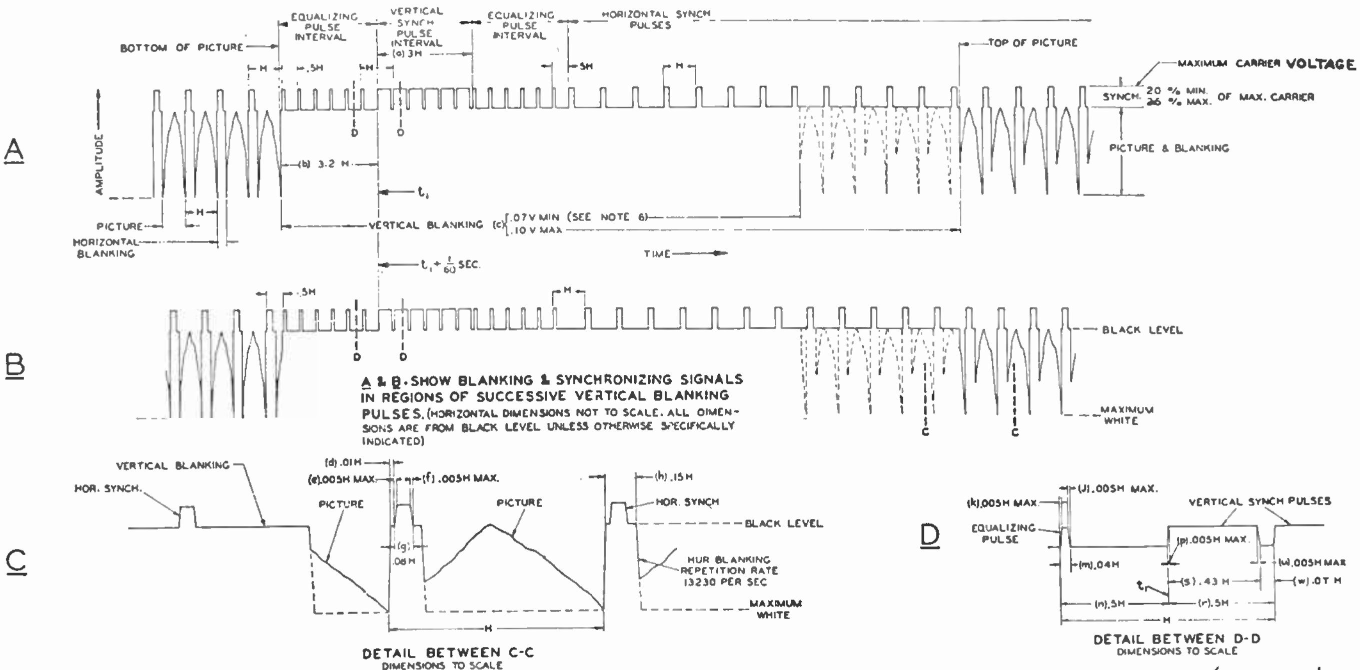
* Receivers can be made to operate on a synchronizing signal which is somewhat less complicated than the R.M.A. standard signal. However, to do so requires either some sacrifice in performance or the use of additional complication in the receiver. Since the number of receivers will be much greater than the number of transmitters, the generation of the R.M.A. signal is economically preferable.

RMA STANDARD T-111 TELEVISION SIGNAL

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



1- DIAGRAM C SHOWS ENLARGED DETAIL VIEW OF SIGNAL IN VIEW B BETWEEN LINES C-C

2- DIAGRAM D SHOWS ENLARGED DETAIL VIEW OF SYNCH. SIGNAL IN VIEW A BETWEEN LINES D-D

3- H-TIME FROM START OF ONE LINE TO START OF NEXT LINE = 1/13230 SEC.

4- V-TIME FROM START OF ONE FIELD TO START OF NEXT FIELD = 1/60 SEC. = 220½ H

5- LEADING AND TRAILING EDGES OF BOTH HOR. AND VERT. BLANKING PULSES HAVE SLOPES (NOT INDICATED IN A & B) WHICH SHOULD BE KEPT AS STEEP AS POSSIBLE

6- RECEIVER VERTICAL RETRACE SHALL BE COMPLETE AT END OF .07 V

Fig. 1—*A* and *B* are portions of the same wave. The Synchronizing-signal Generator produces the portion (25 per cent) of the R.M.A. Standard Television Signal above the "black level". It also produces the blanking portion of the signal shown below "black level".

pulses were synchronized by the common 26,460-cycle pulses, but due to variable degrees of "firing resistance" and finite slope of the synchronizing pulses, timing errors, between the several kinds of pulses in the output wave, of several per cent of H frequently occurred. Further difficulty was experienced due to the three kinds of pulses having different wave shapes of their leading edges and different rates of rise due to different constants in the three generating and keying circuits. These errors sometimes resulted in loss of horizontal synchronism in the receiver at those points of non-uniformity which would require an abrupt increase in oscillation frequency and at other times caused a slight displacement of a few scanning lines at the top of the picture screen. For operation beyond reproach it seems that the error between any two adjacent horizontal-synchronizing leading edges should be considerably less than that corresponding to one picture element, which is of the order of 0.002 H.

In the present synchronizing generator, uniformity of both timing and wave shape of all leading pulses is inherent due to a single 26,460-c.p.s. multivibrator (which is driven by rectangular waves obtained from a tuned oscillator and limiter) producing the leading edges of all pulses in the finished R.M.A. synchronizing wave. These pulses as produced have a width of 0.04 H and without alteration become the "equalizing pulses" in the final wave. A section of the 26,460-c.p.s. pulses are made wider by having other pulses added to their trailing ends in order to widen them to the 0.43 H as specified for the vertical synchronizing pulses. During the region which is to contain only the normal 13,230-per-second horizontal pulses the alternate unwanted pulses are keyed out and the remaining pulses are widened to 0.08 H by adding other suitable pulses to their trailing ends. In each case the leading edges of the original pulses are not altered by the additions. Since any slight change in the shape or slope of the leading edges or delay in their transmission through the circuits will be the same for all types of pulses, no relative errors will be introduced. The widths of the various composite pulses will vary somewhat with unavoidable changes in the timing of the added pulses since they are produced in separate circuits, but considerable tolerance is permissible in this respect. The widening pulses are always added to the 26,460-c.p.s. pulses with an appreciable overlap in time so that no gap in the completed pulses is ever present after a final step of limiting amplification. The principles involved will become more evident when the specific apparatus is described.

Frequency Division vs. Frequency Multiplication

Considered casually it would seem better to begin the frequency chain with the 60-cycle power supply and use frequency multipliers in the various steps to obtain the 26,460-c.p.s. signal, since it would avoid the indirect method of obtaining 60-cycle synchronism and the relatively complex frequency dividers as were used in the later signal generator to be described below. The objection to this simpler method resides in an inherent weakness of the frequency multipliers themselves, namely that the instantaneous-output frequency of a frequency multiplier is not necessarily correct and in close agreement with the average frequency at all times as will be explained. In this type of device the lower frequency sine-wave signal is greatly distorted by an amplifier tube to produce harmonics, and the desired harmonic is then presumably selected and isolated in the plate circuit by a tuned circuit. The operation is imperfect because the tuned circuit can not readily be made to have sufficiently low loss to be adequately selective to isolate the desired harmonic completely. Furthermore even if it were adequately selective it would not be capable of following the slight frequency changes in the 60-cycle power system. (If a flat-top band-pass filter were used the change in phase with frequency might be objectionable if sharp cut-off is obtained.)

From a physical point of view the lower-frequency input signal merely shock-excites the tuned-plate circuit once for each lower-frequency cycle and leaves the tuned circuit to generate say 7 or more cycles by free damped oscillation. Between shocks the frequency may be slightly off the ideal since it is determined only by the tuned circuit. Figuratively speaking it may be said that a 60-cycle per second wave, for example, measures or divides time into 1/60-second units. Then in attempting to produce by frequency multiplication a frequency of say 420 cycles per second we have the difficult task of measuring 1/420-second intervals by a "ruler" calibrated only in units seven times as large. The experimental efforts of others seem to support these conclusions.

Frequency Dividers

Two very different types of frequency-dividing circuits are available: the multivibrator* which, when driven by a higher frequency, may have its natural frequency adjusted so that firing occurs only on say every seventh pulse, and the pulse-counter circuit † which accumulates the effect of several consecutive cycles of pulses without regard to their frequency and fires producing a single pulse output when the accumulated effect is adequate.

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* The blocking oscillator or other forms of self-running relaxation oscillator may be used instead of the multivibrator. The multivibrator is however, the preferred form of this general type.

† The Electric Music Industries, Limited of Great Britain is credited with developing the pulse-counter circuit for synchronizing-signal generators.

Certain advantages in stability lead to the adoption of the "counter" type of circuits for the synchronizing generator in spite of its greater complexity.

With reference to Figure 2(a) the "counter" circuit for frequency dividing may be explained as follows: Since the first frequency divider in the chain has been chosen for the explanation, V_1 is shown amplifying rectangular waves derived from the master 26,460-c.p.s.

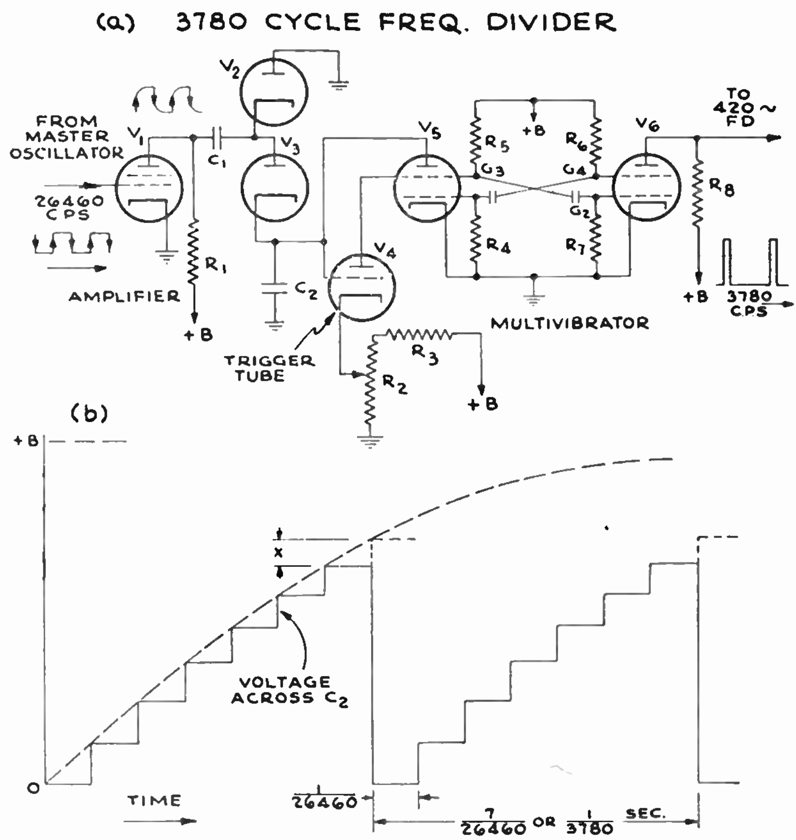

Fig. 2—(a) Frequency Divider. (b) Shows how C_2 is charged in steps through V_3 and C_1 and discharged at intervals by V_5.

oscillator by limiting amplifiers (as shown further in Figures 3 and 4). The plate resistance R_1 is high and the grid swing of V_1 is of saturating magnitude so that the output-rectangular wave has amplitude limited by and approximately equal to the voltage of the power supply + B. Starting with no charges on C_1 and C_2, the plate voltage swings to + B and the capacitors C_1 and C_2 charge in series through diode V_3 to substantially the entire voltage. The + B voltage of say 250 volts is divided between the two condensers inversely as their respective capacitances. C_1 is small and C_2 large so that C_2 has the lower voltage, say 1/20 of + B which is 12.5 volts.

On the negative stroke diode V_2 conducts, discharging C_1 to ground, but not altering the charge on C_2. On the next positive stroke the conditions are repeated except that this time only approximately 225 net volts are available for adding *new* charges to the two condensers. Condenser C_2 then will obtain a second incremental charge of only about 10 volts. On the next negative stroke of the plate, diode V_2 will again discharge C_1. The "stair-step" rise of voltage across C_2 is shown at (b) in the figure. Note that the voltage across C_2 would asymptotically approach the +B voltage as the voltage increments decrease in amplitude if not interrupted. During the "build-up" time trigger tube V_4 is biased beyond "cut-off" due to a definite portion of the positive voltage from the +B supply applied to its cathode. When the "stair-step" voltage across C_2 reaches slightly higher than the cut-off condition for the trigger tube, depending upon the setting of the potentiometer R_2, the trigger tube conducts, driving the control grid of the multivibrator tube V_6 negatively. Tubes V_5 and V_6 are connected in a conventional multivibrator circuit using the screen grids as anodes for the multivibrator action so that the plates are available for other purposes to be explained. The multivibrator constants are such that tube V_5 would remain in the cut-off portion of the cycle for extremely long intervals if it were not for the pulses received from the trigger tube V_4. When the trigger tube conducts, the multivibrator is triggered, tube V_5 conducts and its plate circuit quickly discharges condenser C_2 substantially to zero potential. Then the multivibrator re-sets to its initial condition where V_5 is cut-off and V_6 conducts. The "stair-step" charging cycle starts again and the charge due to the next seven cycles is metered by the trigger tube, and so on. The plate output of tube V_6 is a 3,780-c.p.s. rectangular wave of suitable amplitude to charge a similar "C_1 and C_2" of the next "counter" circuit (or frequency divider).

This frequency-dividing circuit is entitled to be called a "pulse-counter" only on the grounds that it produces one output pulse for every certain number of input pulses over a wide-frequency range. In the present application the ability of the output to "follow" the input for a large frequency change is of no great value, except that it simplifies changes for experimental purposes. The real advantage of using the "counter" circuits instead of the multivibrators alone is their very great stability in counting accurately with changes in tube characteristics and +B voltage. The magnitude of each step in the stair-step charging of condenser C_2 tends to vary in proportion to the +B voltage. The bias voltage on the cathode of the trigger tube, which responds to the accumulated voltage across C_2, also varies in proportion to the +B voltage so that approximate cancellation of the effects of B—voltage changes is obtained. Some tendency to error in "counting" occurs due to the variable resistances of the diodes, which prevent complete charging of C_2 through V_3 and the complete alternate discharging of C_1 through V_2. Also tube V_5 does not discharge C_2 to completion due to tube resistance. However, these effects may be made negligible

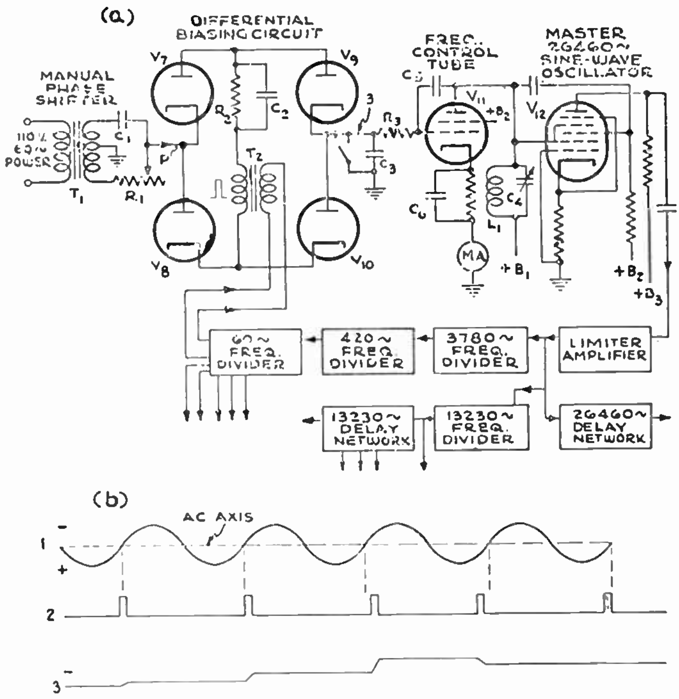

Fig. 3—(a) 60-cycle Locking Circuit. (b) Wave 1 from the Phase Shifter and wave 2 from the 60-cycle Frequency Divider produce frequency control wave 3 at 3 in the circuit. Note wave 3 is uniform except when correction is required.

by using circuit constants of relatively high impedance. Also the cut-off voltage of tube V_4 may vary incorrectly with plate voltage and from tube to tube, but this voltage is relatively small, especially in the high-mu-type tubes. Another and more-serious source of irregularity was found to be due to gas current and leakage in tubes V_4 and V_5 when certain unfavorable types were tried for these circuit positions. The merit of this type circuit is due to the fact that to a certain extent the tubes act only as switches so that stability is largely determined by the relatively-stable condensers and resistors.

Sixty-Cycle Locking Circuit

The frequency-regulating circuit for maintaining the entire synchronizing generator in synchronism with the 60-cycle power system is shown in Figure 3(a), in which each frequency-dividing circuit is represented by a block having the lower or output frequency indicated. The 26,460-c.p.s. oscillator is of the negative-transconductance type and electronic coupling to the plate circuit is used for output. The frequency is determined by the tank circuit which includes the constants L_1 and C_4 and the automatically adjustable impedance due to the plate current of the frequency-control tube V_{11}. Since the grid of this tube is excited from the tank circuit, through a small condenser C_5, which provides phase shift, the plate current is largely in quadrature so the mutual conductance of this tube will affect the resonance frequency. The bias of tube V_{11} and, hence, its mutual conductance is controlled by the "differential biasing circuit" in accordance with the relative phases of the 60-cycle power circuit and the "60-cycle" pulse output of the last frequency-divider of the chain.

The differential-biasing circuit can best be understood by considering the entire bridge circuit comprised by the four diodes, V_7, V_8, V_9 and V_{10}, the condenser C_2, the resistor R_2 and transformer T_2, as merely a key or switch that momentarily connects the 60-cycle power supply through the manual phase shifter to the condenser C_3. The narrow 60-cycle pulses as shown at 2 in Figure 3(b) introduced by transformer T_2 into the bridge cause the four diodes to conduct briefly for each pulse. This conduction charges condenser C_2 so as to retain the diodes entirely biased off between pulses while resistor R_2 continually discharges the condenser slightly so that the diodes will continue to conduct during the pulses. The pulses from the frequency divider as shown at 2 in Figure 3(b) occur near the time the sine-wave power-supply voltage (wave 1) impressed at P crosses the a-c axis from positive to negative and has a maximum rate of change. Hence, slight changes in the relative timing of the frequency-divider circuit and the 60-cycle power line will cause the voltage accumulated upon condenser C_3 to vary considerably as shown by wave 3. Since the sine-wave voltage at P is changing negatively during these pulses, a lagging condition of the frequency divider for example, will adjust the bias on condenser toward the negative, making the frequency-control tube less conducting. This in turn tends to cause the master oscillator to increase its frequency, thereby decreasing the lag of the entire chain of frequency-dividers. Conversely an advancement of the phase of the pulses produces a retarding influence upon the master oscillator. Hence, when the tank, L_1, C_4, is set manually such that its natural frequency would be approximately 26,460-cycles per second, the synchronizing generator will be automatically adjusted to synchronism with the 60-cycle power supply. Also if the manual setting is made such that balance is obtained with the 60-cycle pulses occurring very nearly the time the sine-wave passes through zero, the locking phase relation will not vary appreciably with the amplitude of the power wave. The adjustment necessary for this condition may be determined practically by the milliammeter in the cathode circuit of the frequency-control tube.

The time required for a frequency adjustment to occur after it is needed is very short, but no appreciable 60-cycle changes in voltage occur across C_3 since C_3 can neither lose nor gain charge between the narrow 60-cycle pulses as indicated by Wave 3 being uniform between pulses of Wave 2. Note also that if no frequency correction is needed the control-voltage Wave 3 is uniform d.c.

These advantages were not present in an early form of regulating circuit used experimentally in which the 60-cycle power wave was beat with the 60-cycle pulse wave and rectified to produce variable height controlling pulses. In that circuit the pulses were filtered by R-C circuits to change the pulses to a variable d-c control voltage. Difficulty was experienced in that when the filtering was made adequate to avoid excessive 60-cycle frequency modulation, the control action was sluggish and subject to hunting or over-swing.

Wave-Shaping Circuits for Auxiliary Output Waves

Figure 4 shows the entire synchronizing generator in block diagram. In the left-hand portion of the figure the chains of frequency dividers are indicated by blocks containing the letters FD and a number corresponding to the cycles-per-second output. The block marked "L-26,-460" is the limiter amplifier shown in Figure 3 and supplies the delay network DN-26,460 with 26,460 rectangular pulses per second. Similarly the frequency divider FD-13,230 is a source of synchronized 13,230-cycle rectangular pulses which supplies another delay network DN-13,230 having several different output taps. The frequency divider FD 60 (shown in Figure 3 also) is a 60-cycle synchronous pulse source for synchronizing five separate multivibrators shown as blocks "MV". These 60, 13,230, and 26,460-cycle-per-second sources are used to produce all the synchronizing-generator output signals.

For example the output signal known as "video blanking" is a mixture of 60-cycle-per-second pulses and 13,230-cycle-per-second pulses with the 13,230-cycle pulses eliminated during the occurrence of each 60-cycle pulse. The letter "*d*" in Figure 4 indicates the conductors for this wave and the wave shape is shown at "*d*" in Figure 5. The component waves *a* and *b* are mixed in the mixer-limiter, block ML-1 of Figure 4, to provide the sum wave *c*. Wave *c* is limited in the same block at the level of the broken line to produce the wave d which is supplied and transmitted at 75-ohm impedance by the line amplifier, block LA-1. In each case the small letter adjacent the conductors in Figure 4 corresponds to the wave shape present in that portion of the circuit as shown in Figure 5. Figure 6 shows the essential circuit elements (which are

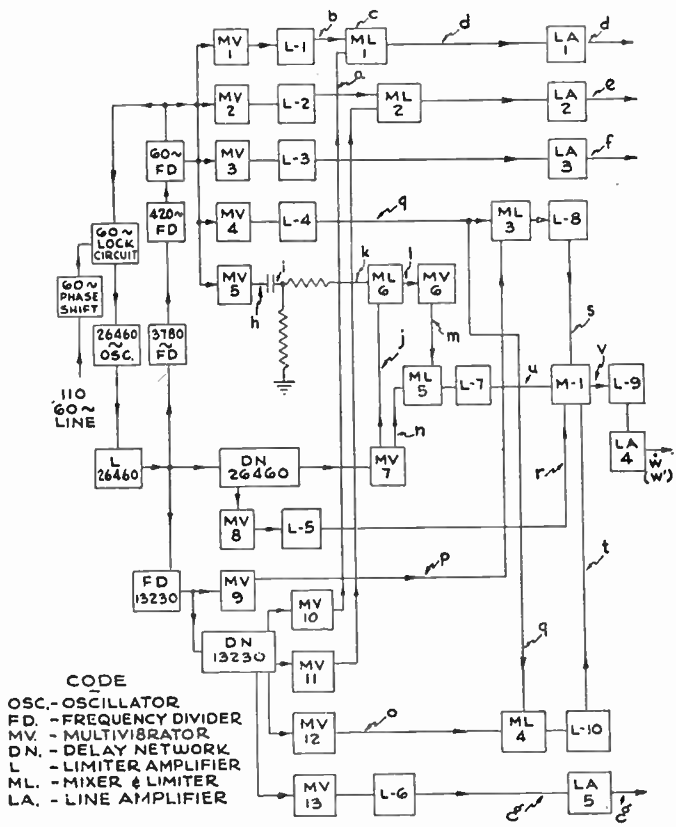

Fig. 4—Block diagram of entire Synchronizing-signal Generator. Letters a, b, c, etc. refer to wave shapes in Figure 5. Wave w (w') represents R.M.A. Synchronized output wave while d, e, f, and g are used for auxiliary equipment.

represented by the blocks in Figure 4) for producing the "video blanking" wave d as explained. Tube V_{13}, the output stage of the 60-cycle frequency divider (FD 60) synchronizes the multivibrator (MV-1) via the amplifier-buffer stage V_{14}. The time constant (RC) in the grid of the first triode of Tube V_{15} is very long compared to that of the second triode of the tube. Hence, the first triode remains cut-off for longer periods and is caused to conduct by negative pulses applied upon the grid circuit of the second triode. The brief conduction period of the first triode of V_{15} can be adjusted by the potentiometers which controls the positive bias on the second triode so that the duration of the positive pulses impressed upon tube V_{16} can be made as long as desired in the wave *b* of Figure 5. Tube V_{16} serves as a mixer since its plate is in parallel with the plate of the second triode in tube V_{18}. Tube V_{16} also limits its positive plate swing by cut-off and its negative plate swing by drawing grid current.

The multivibrator MV-10 which includes Tube V_{17} is synchronized by 13,230-cycle-per-second pulses from the frequency-divider, FD 13,-230, which have been properly delayed by the delay-network DN-13,230. The multivibrator output-wave *a* is combined with wave *b* by means of the second triode of V_{18} to form voltage wave *c*. (Actually waves *a* and *b* are present as current waves in the plate leads while the voltage on the plate of either Tube V_{16} or the second triode of V_{18} is wave *c*.) The first triode of Tube V_{18} and the line-amplifier Tube V_{19} serve as limiting amplifiers to reduce wave *c* to wave *d* by saturating-off the portion of wave *c* above the broken line. The line-amplifier LA-1 is connected for output from its cathode circuit as this provides a lower impedance for operation into a 75-ohm distribution cable.

The "Iconoscope blanking" output wave and the apparatus for producing it is essentially the same as for the "video blanking" wave except that the pulses are shorter and are delayed different amounts in the synchronizing generator due to the Iconoscope blanking being subjected to additional delay by transmission through the television camera cables. The wave is shown at *e* in Figure 5 without the additional delay and the circuit apparatus involved in its generation may be determined by following backward along the lines from the letter *e* in Figure 4.

The "vertical-driving signal for the Iconoscope" is a simple 60-cycle-pulse wave as shown by wave *f* in Figure 5. Its generation involves only one multivibrator MV-3 (synchronized by the 60-cycle frequency-divider), a limiter and a line amplifier as shown in Figure 4.

The "horizontal driving signal for the Iconoscope" is a similar kind of wave involving a similar type of apparatus except that the frequency of the pulses is 13,230 per second as shown by the wave *g* in Figure 5 and the letter "*g*" in Figure 4.

Circuits for Generating the R.M.A. Synchronizing Wave

The last signal output of the synchronizing generator is the "R.M.A. synchronizing wave" as indicated at *w* and *w'* in Figure 5 for the intervals near the even and odd vertical pulses respectively. (*W* and *w'* are views of the same voltage wave taken 1/60-second apart.) For clearness the number of equalizing pulses and the duration of each vertical pulse shown in w and w' has been halved as can be seen by comparing with the R.M.A. standard drawing T-111 of Figure 1. Since several unusual methods are employed to insure a very high degree of accuracy in the wave at its critical points, a brief review of the steps is of interest.

All of the waves from h to v in Figure 5 are generated and used in various combinations to produce the final wave w. In the last step, wave w is obtained from wave v alone by simply limiting or "clipping" at the positive and negative levels indicated by the broken lines on wave v. Wave v on the other hand is derived by adding the four waves

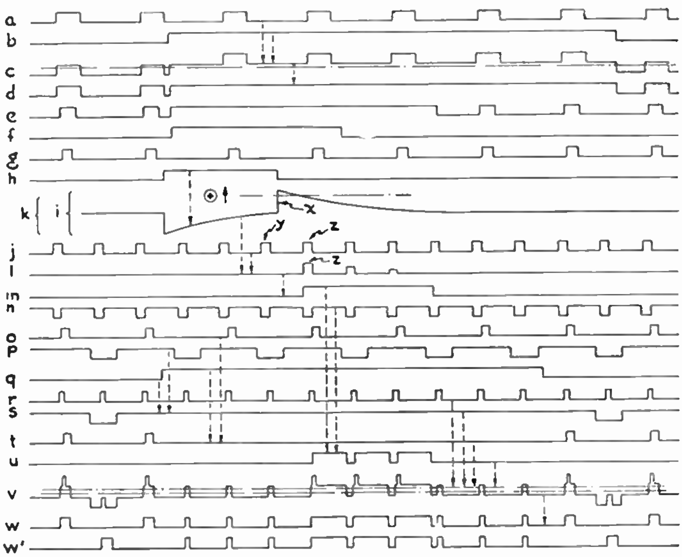

Fig. 5—Wave shapes present in circuit of Figure 4 at points indicated. Wave v, from which R.M.A. synch wave w is derived, is the sum of r, s, t, and u as shown by dotted arrows.

r, s, t, and u in mixer M-1 as indicated in Figure 5 by the vertical dotted arrows connecting the several waves. Wave r is a simple 26,460-cycle pulse wave produced by a delay-synchronized multivibrator, MV-8, and a limiter, L-5. Wave s is the 13,230-cycle pulse wave p after having a group of the pulses keyed-out by each pulse of the 60-cycle wave q. The keying is accomplished in the mixer-limiter, ML-3, of Figure 4 by applying the two waves p and q respectively to the first and third grids of a tube of the type commonly known as a pentagrid converter. When wave p allows electrons to pass the first grid, the wave q modulates their flow to the plate output circuit whereby one wave modulates the other. Wave t is produced in a similar manner by allowing wave q to key wave o in the mixer-limiter ML-4. The waves p, q, and o originate in multivibrators synchronized by suitably-delayed pulses.

Wave *u* is obtained by the 60-cycle pulse wave *m* keying the 26,460-cycle wave *n* in the mixer-limiter, ML-5, with such polarity that the high-frequency pulses are passed only *during* each 60-cycle pulse.

The pulse of wave *m* must be delayed with considerable accuracy with respect to the output of the 60-cycle frequency-divider in order to insure the leading pulse of each 60-cycle group of pulses in wave *u* being a whole pulse. The main delay is obtained indirectly by using the back or trailing edge of the pulses of wave *h* instead of a delay network as will be explained. Wave *h* is distorted to a shape such as shown at *i* by transmission through a small condenser with a resistance

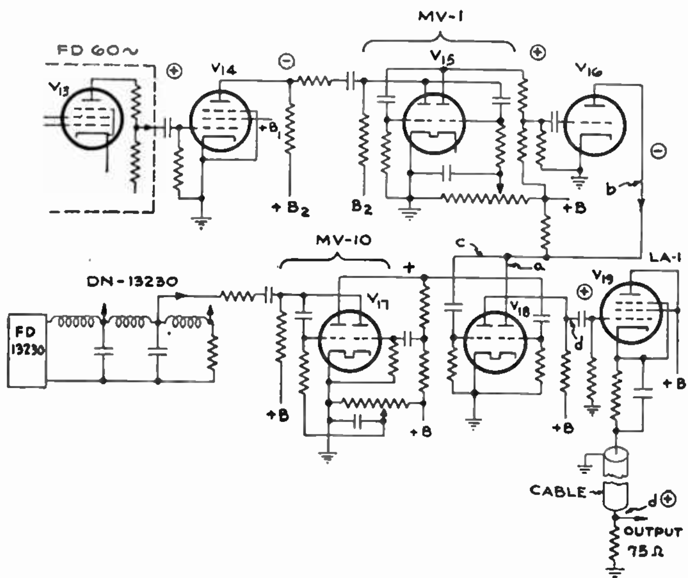

Fig. 6—Multivibrators, mixer, limiter, and line amplifier for combining waves *a* and *b* to form wave *c*, which is limited to produce *d*.

output load. It is then applied to the first grid of a pentagrid tube through a series resistor which limits wave *i* at the broken line due to grid current, producing wave *k*. Wave *j* is applied to the third grid of the tube and the limited wave *k* holds the tube cut-off except when pulse *x* of wave *i* occurs. Then the plate-output circuit contains small isolated 60-cycle groups of the 26,460-cycle pulses as shown in wave *l*. The leading pulse of each 60-cycle group of wave *l* then accurately triggers the multivibrator MV-6. This multivibrator produces the 60-cycle-pulse wave *m* which is used as described above. Wave *m* is, therefore, accurately timed with respect to the 26,460-cycle-pulses of wave *n* since waves *n* and *j* are outputs of the same multivibrator. It should be noted that the trailing end of the pulse of wave *h* could occur any time be-

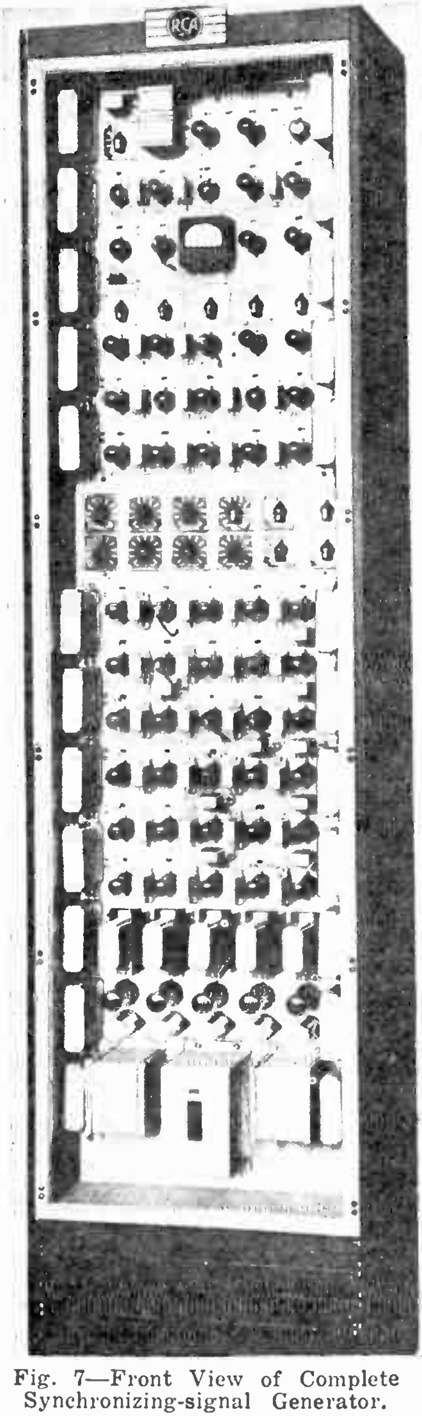

Fig. 7—Front View of Complete Synchronizing-signal Generator.

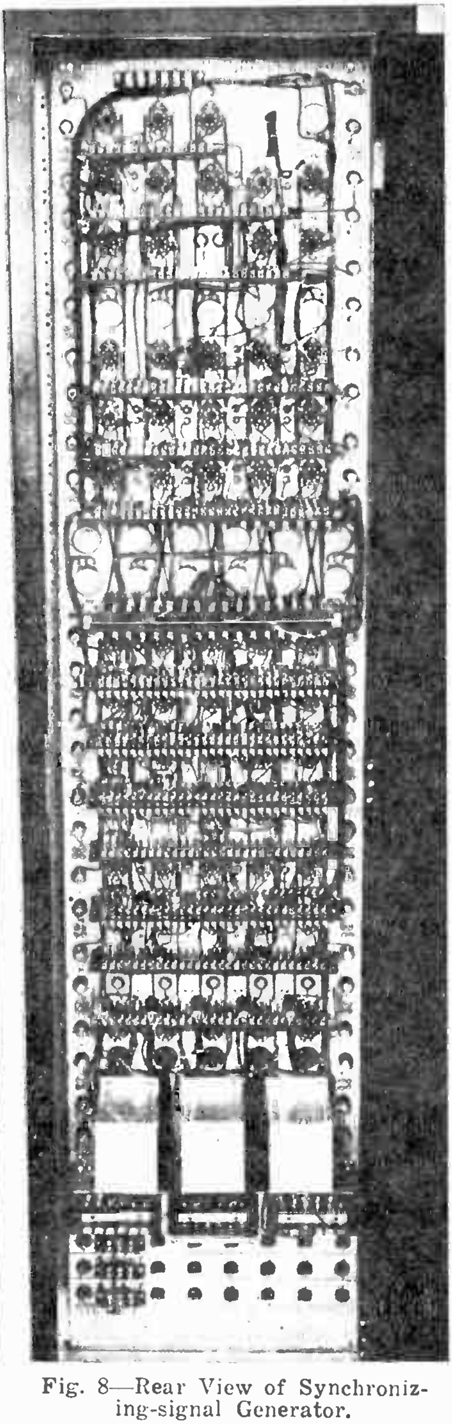

Fig. 8—Rear View of Synchronizing-signal Generator.

tween the pulses *y* and *z* of wave *j* and still select the same pulse *z* as the leading pulse in the group in wave *l*. Such tolerance in operation allows reliable accuracy in performance without critical adjustments.

As mentioned before, the extreme uniformity of all leading edges in the finished synchronizing wave *w* is due to the fact that the same oscillator produces all pulses of wave *r*, which provide the leading edges of all the pulses of various widths in the finished wave. The unwanted pulses of wave *r* are extinguished by wave *s* and certain other pulses are made wider by adding to their trailing end the pulses of waves *t* and *u*. If it had been attempted to add these pulses so that no over-lap was caused, i.e., as a sort of "butt point" the relative timing and wave shapes would have to be adjusted to extreme accuracy to avoid notches and gaps in the sum pulses. Study of Figure 5 shows that by using a "lap-joint" between adjoining pulses and by subsequent clipping, considerable tolerance in the timing and shape of the various component waves may be allowed.

Practical Features

Figures 7 and 8 respectively show front and rear views of the synchronizing generator mounted in a cabinet rack. All of the 62 tubes (most of which are double tubes) and the controls are on the front of the single chassis, and are made accessible by opening the hinged door. Most of the smaller components are mounted on bakelite terminal boards and are readily accessible as seen in the rear view. The wiring has been greatly simplified by carefully grouping the apparatus so that leads are very short. This also avoids the need of shielded wires except in a very few connections. Adequate electric isolation of the various parts from one another is obtained by locating the parts in suitable groups on the chassis. Of course all circuits which contain pulse signals having steep wave-fronts, corresponding to several megacycles, are relatively low-impedance circuits in order to provide fidelity and prevent cross-talk.

The tube heaters are supplied by the five 60-cycle transformers near the bottom of the chassis, two in front and three in the rear. The plate supply required is 770 milliamperes at 250 volts, which is usually supplied from two external regulated power rectifiers operating on 110 volts a.c. (320 milliamperes of this current is used in the five line amplifiers for distribution about the studios.)

Since a failure of the synchronizing generator in a commercial television installation might cause a serious interruption or impairment of service, great attention was given to attain reliability and *inherent* accuracy (rather than accuracy which is dependent upon critical adjustments of controls). The use of relatively complicated circuits in the signal generator which require the use of many tubes (with the resulting increased number of chances for tube failure) might seem to decrease the reliability. However, the circuits were chosen to permit the operation to be unaltered by very large changes in tube characteristic without readjustment of controls. Therefore, routine checking of tubes should avoid failure by the normal deterioration of tube characteristics. Furthermore, abrupt structural failure of tubes is relatively rare when, as in this case, all tubes are operated conservatively within their rating.

Similarly, a number of variable controls are provided as an aid to reliability as they allow easy periodic adjustment to optimum mean positions using an oscilloscope for an indicator. This insures that gradual changes in the circuit elements will not likely cause failure. They also permit some changes for experimental purposes.

Considerable experience with the several factory-built synchronizing generators of the type described, indicates that excellent results may be expected if reasonable care is used in routine maintenance.

ACKNOWLEDGMENT

The authors express appreciation to Messrs. K. R. Wendt and A. C. Schroeder for many valuable suggestions in connection with the development of this signal generator. Also Messrs. C. H. Vose and F. E. Cone contributed many desirable mechanical features.

A VESTIGIAL SIDE-BAND FILTER FOR USE WITH A TELEVISION TRANSMITTER*†

By

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Summary—The television transmitter standards adopted by the Radio Manufacturers Association (R. M. A.) place the carrier of the picture transmitter at a point which is 1.25 megacycles above the lower edge of a six-megacycle channel. The television receiver characteristics permits reception of the carrier, the upper side-bands, and those lower side-bands which lie within 0.75 megacycle of the carrier. Any other lower side-bands which might be transmitted would not be accepted by the receiver, and would therefore play no part in furnishing picture information. Since these side-bands which are not accepted by the receiver lie outside of the assigned channel, they must be suppressed at the transmitter in order that there may not be any interference caused to other services operating on adjacent channels. This paper describes a filter which has been built for this purpose.

The filter is placed in the transmission line between the power amplifier and the antenna. In order to insure the absence of reflected energy on the transmission line leading to the filter at any frequency generated by the transmitter, the filter is so designed that the input impedance of the filter is practically a constant for both the pass and rejection band, while the reactance remains essentially zero.

In order to secure the constant-resistance feature throughout the rejection band, the rejected energy is dissipated in water-cooled resistors of a special type which have zero reactance and constant resistance throughout the required band.

The equivalent circuit of the filter is shown. Because of the high carrier frequency as well as the extreme selectivity required, the circuit elements are sections of concentric transmission line. The factors governing the lengths and diameters of these sections are discussed.

The vestigial side-band filter described here has been placed in a practical television installation and has been operating satisfactorily since March, 1939. The tests and observations made at the time of this installation are described.

I. INTRODUCTION

THE channels allotted to television stations in this country are limited in width to six megacycles. An early form of channel layout was as shown in the top diagram of Figure 1. The sound carrier was located one-quarter megacycle below the upper edge of the band. The picture carrier was located 2.5 megacycles above the lower edge of the band. With the picture carrier located at this point, the maximum possible modulation frequency was 2.5 megacycles. Any higher frequency modulation would cause side-bands which would lie

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outside the assigned channel. The receiver was operated in a semi-single-side-band fashion, having the response characteristic shown in the top diagram of Figure 1. It is to be noted that side-bands which lie more than 0.75 megacycle below the picture carrier are not accepted by the receiver and hence serve no useful purpose in forming the picture. This point is of importance in explaining the new channel arrangement.

In the new arrangement, the sound carrier remains unchanged in its position, but the picture carrier is now placed 1.25 megacycles above the lower edge of the band, thus providing an additional space of 1.25 megacycles between picture and sound carriers. The receiver characteristic is modified as shown in the lower sketch of Figure 1. If the transmitter is now modulated with frequencies up to four megacycles, all the frequencies in excess of 1.25 megacycles will cause lower side-bands to be formed which will lie below the lower edge of the assigned band. These side-bands may easily cause interference with other services which may operate in this region. Since these side-bands are not accepted by the receiver, it seems obvious that they may be filtered off at the transmitter.

It is the object of this report to describe a filter which has been built for this purpose. The filter is placed in the transmission line somewhere between the antenna and the power amplifier of the transmitter. In a conventional high-pass filter, without losses, the input impedance would become practically a pure reactance in the rejection band. If this filter were placed directly at the transmitter, this build-up of reactance in the lower side-band region would cause unsymmetrical operation of the final stage, with ensuing generation of transients. Further, if the filter were placed some distance from the transmitter, the reactance in the rejection region would cause reflected energy to be present on the transmission line leading to the filter, with consequent multiple images in the received picture. The filter discussed here has been so designed that the input resistance of the filter is practically a constant for both the pass and rejection band, while the reactance remains essentially zero.[1] Because of this fact, the transmission line between transmitter and filter is terminated in its characteristic impedance for all frequencies, pass and rejection. Then the transmitter looks into a pure resistance over both upper and lower side-bands, and the transmitter generates a double side-band signal, quite unaware of the fact that much of the lower side-band energy is not radiated.

[1] For an excellent mathematical treatment of constant resistance networks, the reader is referred to E. L. Norton, "Constant Resistance Networks with Applications to Filter Groups", *Bell. Sys. Tech. Jour.*, April, 1937, and to U. S. Patent 2,076,248 issued to E. L. Norton.

The lower side-band energy that is not radiated is absorbed in water-cooled resistors which form an integral part of the filter.

The receiver characteristic for the new arrangement is shown in the lower sketch of Figure 1. (Curve *R*.) Here also is shown the desired characteristic of the transmitter filter. (Curve *T*.) Roughly, the requirements are that the filter shall pass all frequencies above a point which is itself 0.5 megacycle above the lower edge of the band,

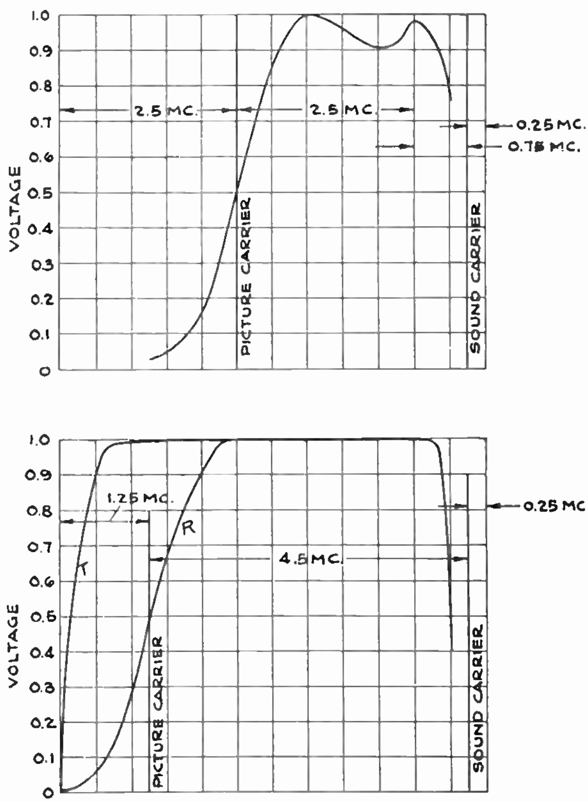


Fig. 1.

and that the attenuation of all frequencies below the lower edge of the band be as great as possible.

It was found that it was not practical at the time to secure this characteristic in a single filter stage. Actually two distinct types of filter are used, with three sections of each type.

II. Type A Filter

The Type A filter is shown schematically in Figure 2. In discussing this filter, we will consider the channel to extend from 44 megacycles to 50 megacycles, since that is the channel for which the first filter was built. Turning then to Figure 2, we may specify the values of the inductances and capacitances shown. It is assumed that the resistance of the antenna (or the input to the antenna transmission line) is 70 ohms. Then the dissipative resistor is made to be 70 ohms. The inductance L_1 is so chosen that it has a reactance of 70 ohms at 44 megacycles, while C_1 is a capacitor having a reactance of 70 ohms at 44 mc. L_2 and C_2 are so proportioned that the combination of the two in a

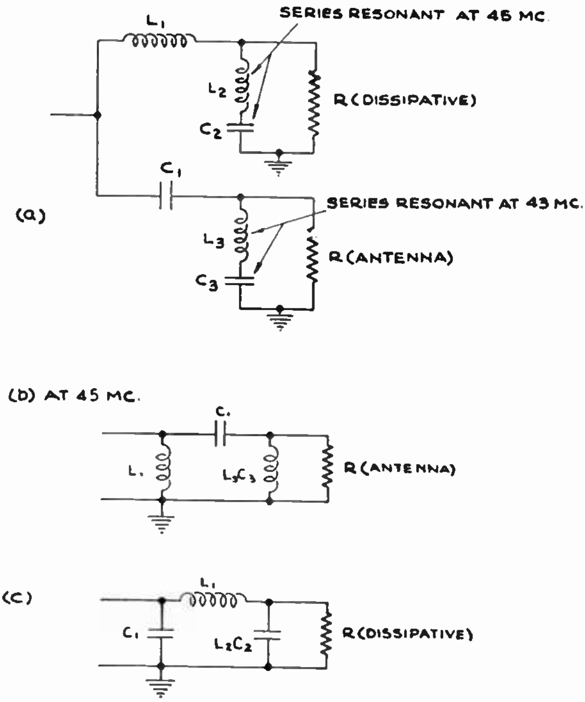


Fig. 2.

series circuit becomes resonant (zero reactance) at 45 megacycles, and at the same time, reaches a capacitive reactance of 70 ohms at 43 megacycles. L_3 and C_3 are chosen so that they yield zero reactance at 43 megacycles, and rise to an inductive reactance of 70 ohms at 45 megacycles.

Let us now examine the circuit at 45 megacycles. The dissipative resistor is short-circuited by the combination of L_2 and C_2, so that no energy goes to the resistor. The reactances of L_1 and C_1 have changed only slightly from 70 ohms. The combination of L_3 and C_3 has placed an inductive reactance of 70 ohms in parallel with the antenna resistance, yielding the equivalent circuit of Figure 2b. Since each of the arms of the Pi network has an impedance of 70 ohms, we have simply a one to one impedance transfer, with an input impedance of 70 ohms, pure resistance, and with a 90-degree phase advance between output and input voltages. All the energy put into the system comes out at the antenna terminals.

At 43 megacycles we have exactly the reverse procedure. The antenna resistance is short-circuited, and the Pi section of Figure 2c is obtained. The input impedance is still 70 ohms, pure resistance, but all of the energy is passed to the dissipative resistor.

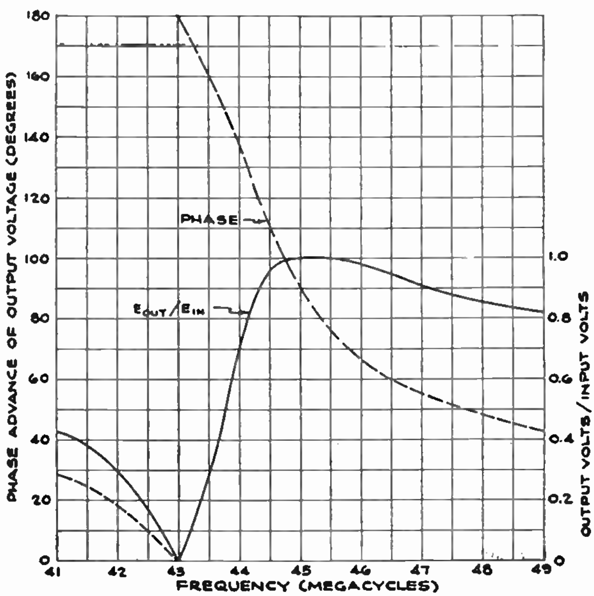


Fig. 3.

We have thus illustrated that the input impedance of the filter is a pure resistance of 70 ohms at 43 and 45 megacycles. As a matter of fact, calculations show that this condition exists for all the frequencies in question.

Figure 3 shows the phase and amplitude characteristics of the Type A filter. We see that we get very great attenuation in the region of 43 megacycles, but that the curve rises rapidly so that at 41 megacycles the attenuation is not very great. To pull down the output in the region from 41 to 43 megacycles, it becomes necessary to place three Type A filters in series. This point will be taken up later in the paper.

Let us turn now to the actual construction of the filter. If we were to use lumped inductances and capacitances, the required values of each element would be as shown in Table I. The construction of elements of this magnitude, accurately tuned to the proper frequency, and able to stand the transmission of power of the order of 10,000 watts, becomes a serious problem, which was briefly considered before turning to the use of concentric-transmission-line elements.

TABLE I

L_1	0.253	microhenries
C_1	51.75	micromicrofarads
L_2	2.72	microhenries
C_2	4.6	micromicrofarads
L_3	2.845	microhenries
C_3	4.81	micromicrofarads

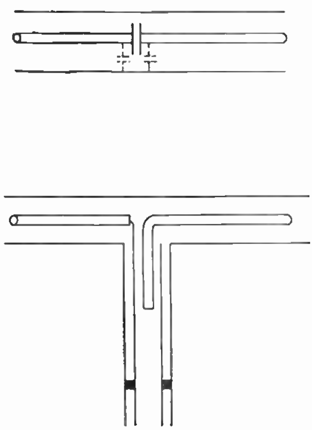


Fig. 4.

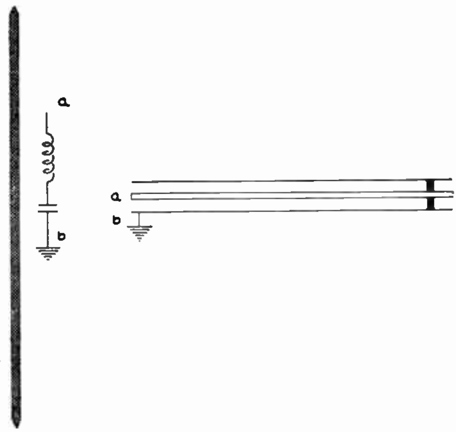


Fig. 5.

The capacitance, C_1, is to be inserted as a series element, that is, both terminals are above ground potential. It is well known that a section of concentric transmission line, of 70 ohms characteristic impedance, open-circuited at the far end, and of length equal to one-eighth wave, has an input impedance which is capacitive and has a value of 70 ohms. However, such an element could not be placed in the hot lead because of the capacitance from the outer conductor to ground. To get around this difficulty, the outer conductor of this capacitance element is extended until its length is one-quarter wave, and the remote end is connected directly to ground. This procedure makes the impedance of this outer conductor to ground very high and effectively floats the series capacitance element. The construction is shown in Figure 4.

The inductance element, L_1, is obtained in exactly the same manner except that a shorting plug is placed at the end of the eighth-wave rod.

The series resonant elements which shunt the antenna resistance and the dissipative resistor are formed by single sections of concentric lines, with shorting bars at the far end, as shown in Figure 5. It is known that the input impedance of a transmission line which is shorted at the remote end becomes zero when the length of transmission line is any integral multiple of a half-wave length. As the frequency increases from this critical value, the input impedance becomes

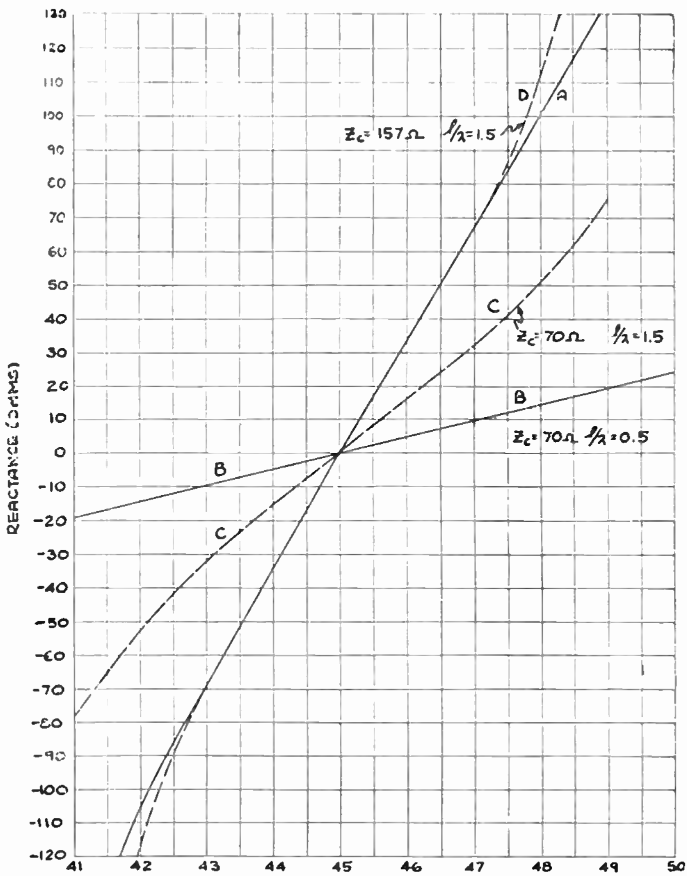


Fig. 6.

inductive, and at frequencies below the critical frequency, the input impedance becomes capacitive. We will consider the case of L_2 and C_2, which have zero reactance at 45 megacycles and reach a capacitive reactance value of 70 ohms at 43 mc. Figure 6 shows several reactance curves, one of which (Curve *A*) shows the calculated curve obtained when L_2 and C_2 have the lumped constants shown in Table I. Curve *B* is the input reactance curve obtained when the line section has a char-

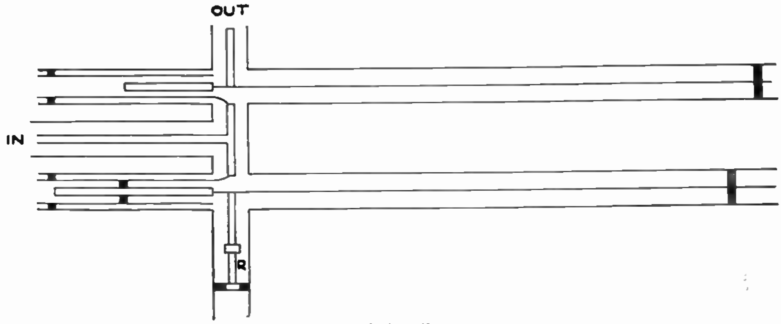


Fig. 7.

acteristic impedance of 70 ohms, and has a length equal to one-half wave. We see that the capacitive reactance at 43 megacycles is only 9.85 ohms. Curve *C* shows the input reactance when the line section has a characteristic impedance of 70 ohms, and is 1.5 waves long. The capacitive reactance at 43 megacycles is now 31.2 ohms. Curve *D* is the input reactance when the line section has a characteristic impedance of 157 ohms and is 1.5 waves long. This curve coincides almost exactly with the desired curve. Therefore, the line element chosen to simulate L_2 and C_2 consists of a shorted transmission line which is 1.5 wave lengths at 45 megacycles (32.8 feet) and which has a characteristic impedance of 157 ohms. The inner diameter of the outer pipe is 4.5 inches, while the inner rod has a diameter of 0.34 inch.

The constructional layout of a single Type A filter is shown in Figure 7. When three of these units are used in series, the point marked "out" on Figure 7 is connected to the input of the following filter.

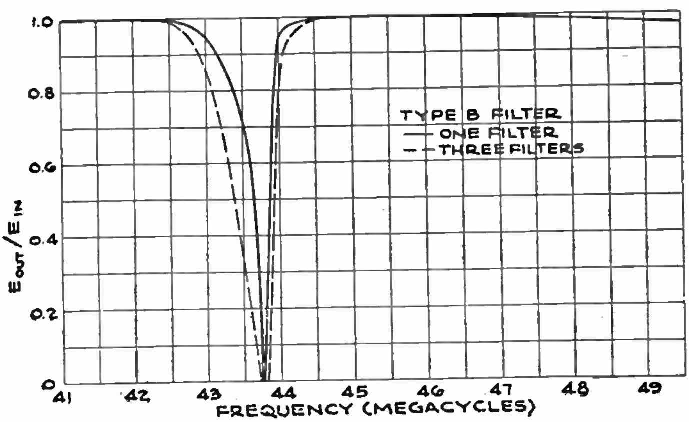


Fig. 8

III. Type B Filter

When three Type A filters are connected in series, the resulting attenuation curve is satisfactory except for the small region around 43.75 megacycles, where it was felt that the attenuation was not great enough. Accordingly, it becomes desirable to use a filter which has a maximum attenuation point at 43.75 megacycles. In fact, it seemed desirable to simply cut a notch in the pass characteristic at this point.* Figure 8 shows the attenuation of such a notching filter (Type B) for the case where a single Type B filter was used and where three such filters were placed in series. These are experimentally determined curves.

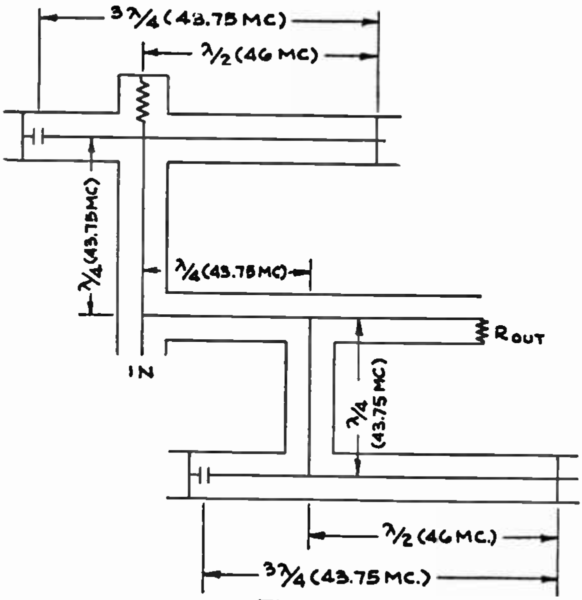


Fig. 9.

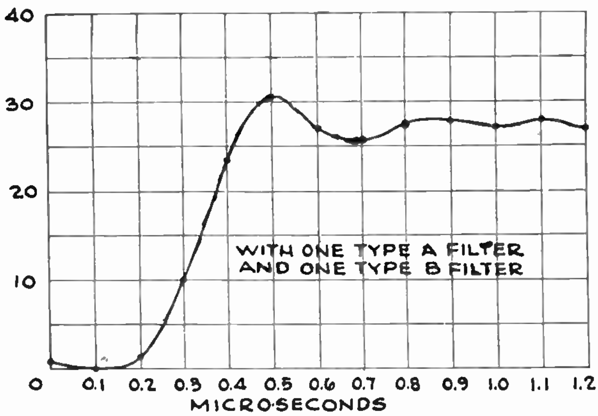
The Type B filter is likewise constructed of concentric transmission lines. The filter utilizes a water-cooled resistor in each unit, and each unit possesses the constant input resistance feature. The filter arrangement is shown in Figure 9. The element lengths are shown on this sketch.

IV. Filter Tests

During the early development work, a single Type A and a single Type B filter unit were built and tested in Camden. The measured

*This notching filter becomes increasingly important when used on the 50 to 56 megacycle channel, for then the maximum attentuation point would occur at 49.75 megacycles, the sound carrier of the 44 to 50 megacycle television channel.

amplitude characteristic of the Type A filter was so close to the calculated characteristic shown in Figure 3 that it seems unnecessary to repeat it here. As stated above, Figure 8 shows the measured response of the Type B filter. The input impedance of these two units in series was found to be quite satisfactory. Tests were made in which the filter



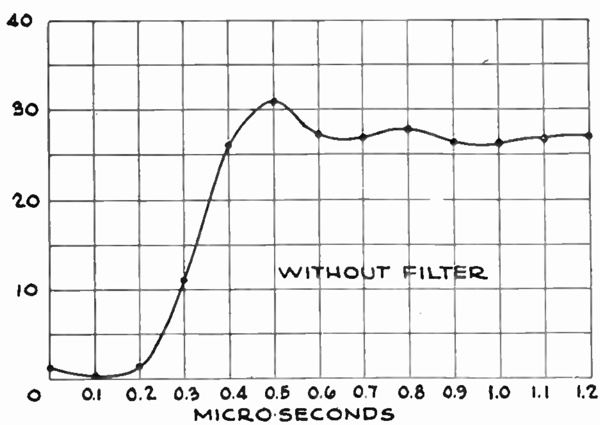


Fig. 10.

combination was fed with a signal generator which was modulated by means of a standard test pattern. A television receiver was placed at the output of the two filters. Observations were made of the received test pattern. The filter was next removed from the system, and the signal generator fed directly to the receiver. The test pattern without the filter was then observed. Most observers agreed that there was no essential difference between the two pictures.

The signal generator was modulated with a square wave, and the response at the second detector of the receiver recorded on a cathode ray oscilloscope. The filter was then placed in the circuit, and the square-wave response again observed. The results for the two cases are shown in Figure 10.

The tests on these single filter units were satisfactory in every respect. Of course, the attenuation, while in strict agreement with the calculated values, was not as great as necessary for the problem at hand. Accordingly, two more Type A filters and two more Type B filters were constructed.

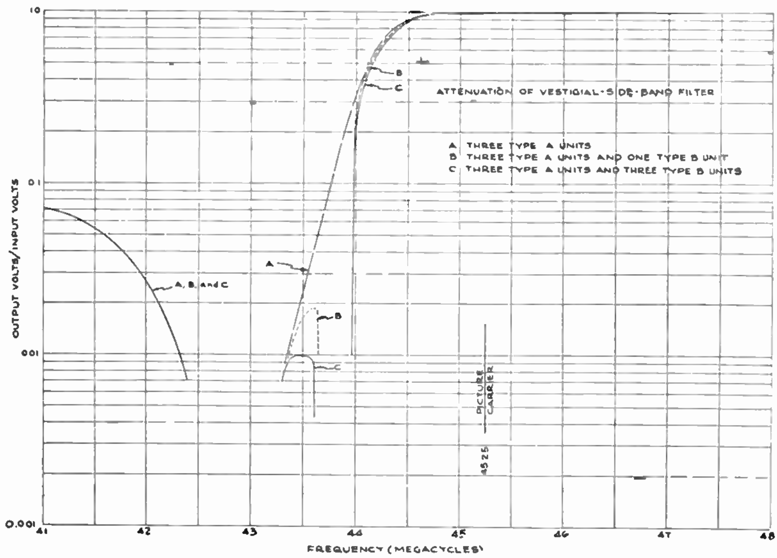


Fig. 11.

With three filters of each type, the response to a square wave was again determined and the test pattern was observed. These tests were again satisfactory. The overall attenuation characteristic as measured is shown by Figure 11. The results shown by this curve were considered to be adequate.

Observations of standing waves on the 70-ohm feed line leading from the test oscillator to the filter showed that the filter system offered a very good termination to the transmission line. The observed reflection on the feed line is shown as a function of frequency in Figure 12.

The entire set of filters as set up in Camden is shown in Figure 13. The input to the filter is at the extreme left rear and is not visible in the picture. Three Type A filters, stacked one above the other, are shown on the left, while the three Type B filters are on the right.

After a thorough series of tests in Camden, the filters were dismantled and reassembled on the eighty-fifth floor of the Empire State Building, New York City, where tests could be made using the television transmitter of the National Broadcasting Company. Many tests

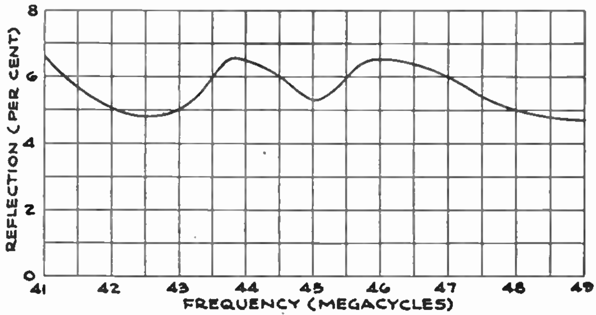


Fig. 12.

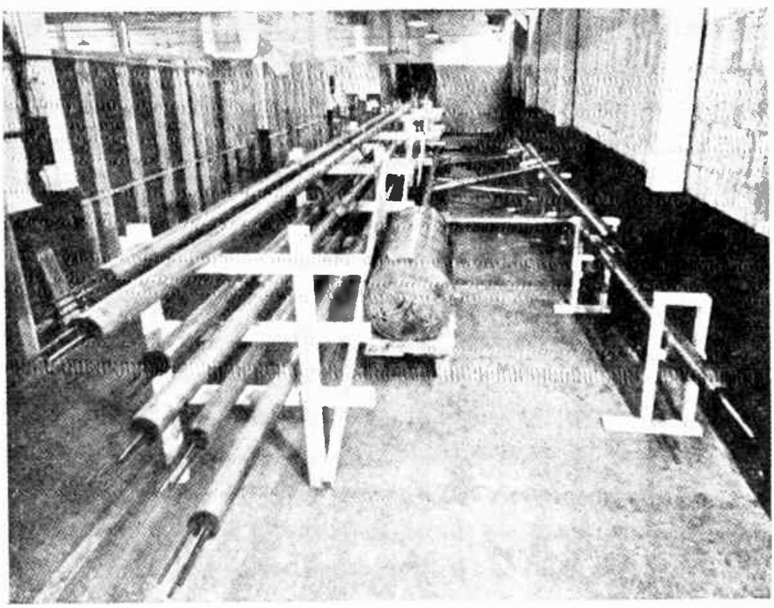


Fig. 13.

were made in 1938 and early in 1939. Field observations were made for a time using a temporary antenna located on the side of the building at the same level as the eighty-fifth floor. During March, 1939, the filter system was connected to the new antenna[2] which had just

[2] Nils E. Lindenblad, "Television Transmitting Antenna for Empire State Building," *RCA REVIEW*, Vol. III, No. 4, April, 1939.

been constructed on the top of the Empire State Building. On April 30, the new system went into regular operation.

The new top antenna is fed by two 55-ohm lines operated in push-pull. However, the output of the filter is arranged to feed into a single 70-ohm line. Accordingly, a matching drum was constructed to transfer from a single-ended 70-ohm line to the two push-pull 55-ohm lines. A sketch of this drum is shown on Figure 14. We see from the left hand (a) sketch that it would not be possible to drive the two push-pull lines with equal and opposite voltages unless means are provided to effectively isolate the end of the outer conductor of the 70-ohm

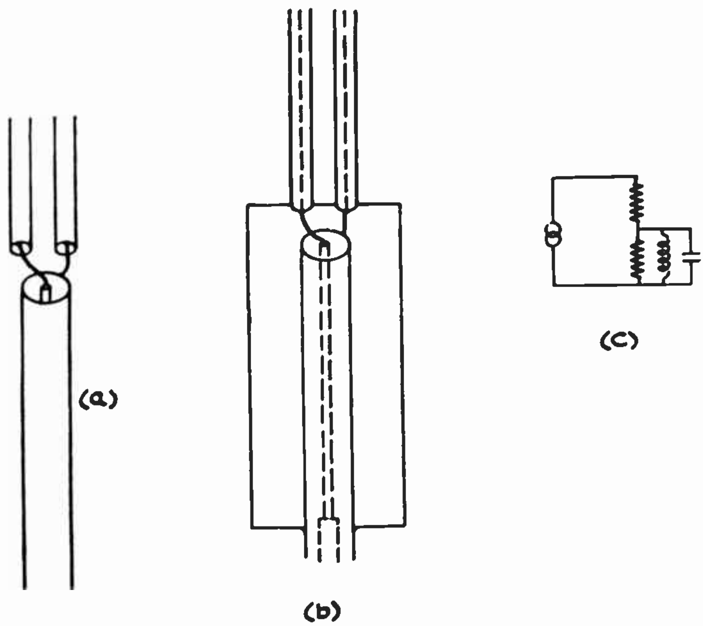


Fig. 14.

line. A quarter wave isolating section is placed around the 70-ohm line, as shown in the middle (b) sketch of Figure 14. The equivalent circuit, which determines the degree of equality of the voltages on the two 55-ohm lines, is shown on the right side (c) of Figure 14. This type of converting drum is described by Lindenblad.[2] In our particular case, the diameter of the outer drum was made large in order to secure high impedance for this shunting or isolating circuit over a wide band of frequencies. Figures 15 and 16 show views of the drum which is used at the Empire State Building. The drum may also be seen in

Figure 13. Our tests showed that this method of going from single-ended to push-pull was quite satisfactory, even for the wide range of frequencies required in a television system.

Reference has been made earlier in the paper to the water-cooled resistors used in the filters. These resistors were designed especially for this application. Essentially, the resistors form the inner con-

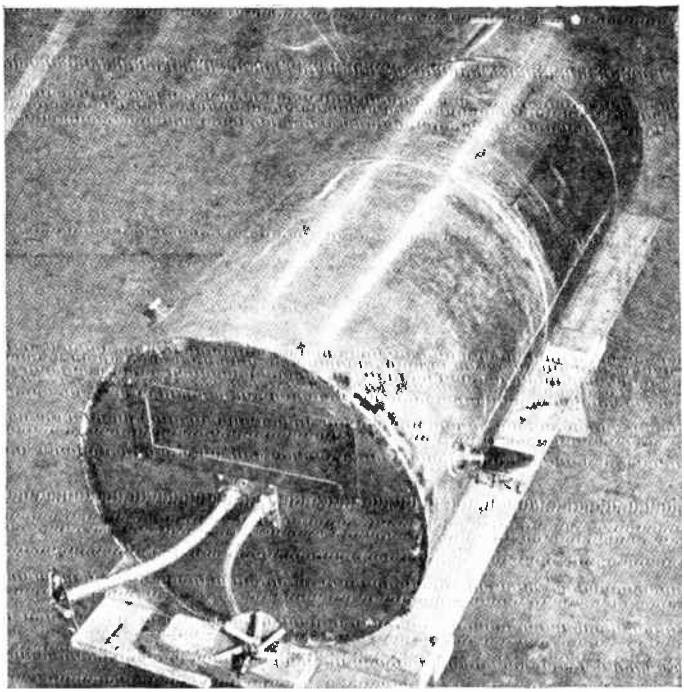


Fig. 15.

ductor of a transmission line. One end of the resistor is shorted to ground. Water is sent through this shorting plug, through the middle of a ceramic tube which supports the resistive film, and back over the surface of the resistor tube. The resistor unit, together with the shorting plug, is shown in Figure 17. As may be seen, the resistor is encased in a glass tube so that the water will return over the surface of the resistor. The theory behind the design of this type of resistance termination is given elsewhere.[3] These resistors form a very good termination for the 70-ohm line. In addition, a rather large amount of power may be dissipated with a small amount of water flow. Tests

[3] G. H. Brown and J. W. Conklin, "Water-Cooled Resistors for Ultra-high Frequencies," *Electronics*, April, 1941.

have shown that the resistors will safely handle a power of one kilowatt with a water flow of slightly more than one gallon per minute. Six of these resistors have been in regular use at the installation in the Empire State Building, without any failures to date.

Early in March, 1939, the transmitter was operated into the top antenna with the filters removed from the circuit. A short time later,

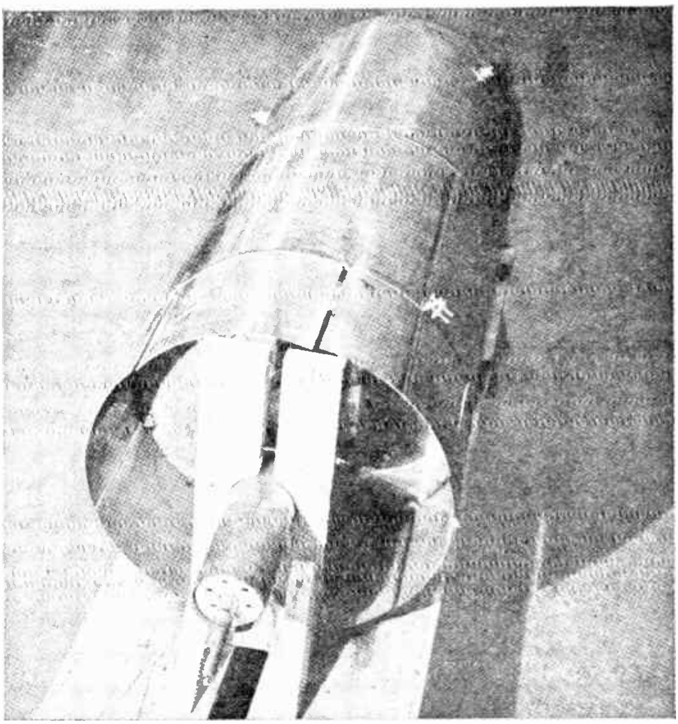


Fig. 16.

the filter was placed in the system, with a resulting transmitted picture that was unimpaired by the insertion of the filter.

V. Conclusion

The filter systems described in this report make possible the use of higher modulation frequencies, that is, the limited assigned band is utilized to the best advantage. The filter was built and tested in the laboratory, where almost ideal conditions existed. Later, the filter was installed in a high power television system, and is now in operation in a practical system.

APPENDIX I

A Consideration of the Energy Distribution in the Side-Bands of a Television Signal

While our general experience and observations indicate that the major portion of the radiated energy of a television picture lies in the frequency spectrum very close to the carrier frequency, and that the energy dissipated in the filter resistors is small, it seems desirable to examine this distribution for a few typical cases. We will consider the case of a picture which consists of alternate vertical black and white bars of equal width. The superimposed synchronizing signals will be neglected. Then the radio frequency signal will be as shown in Figure 18. This signal consists of a sine-wave voltage which has a

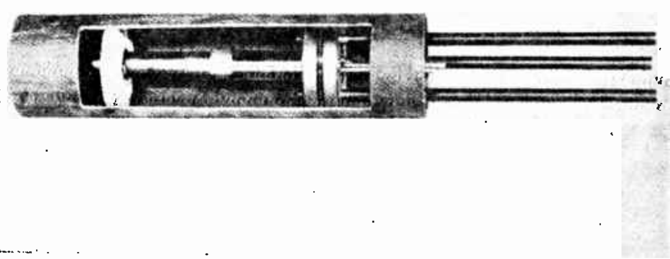


Fig. 17.

frequency equal to the frequency of the carrier. This sine wave of voltage exists for a period of time T, becomes zero for an equal period, and then repeats. This signal may be analyzed by the method of Fourier. It will then be found[4] that the signal may be represented by a carrier frequency and upper and lower side-bands equally distributed on both sides of the carrier. The amplitudes of these side-bands are represented by a simple relation.

If the voltage of the intermittent sinusoidal voltage is E, the carrier voltage, of frequency f, has a magnitude of $\frac{E}{2}$. The first upper side-band has a frequency of $f + \frac{1}{2T}$ and a magnitude of $\frac{E}{\pi}$. The next

[4] L. J. Peters, THEORY OF THERMIONIC VACUUM TUBE CIRCUITS, McGraw-Hill, 1927. Pages 124-127.

upper side-band has a frequency of $f + \frac{3}{2T}$ and a magnitude of $\frac{E}{3\pi}$

The lower side-bands follow similar relations. The pertinent relations are shown in the following table. (If T is expressed in microseconds, the frequencies will be given in megacycles.)

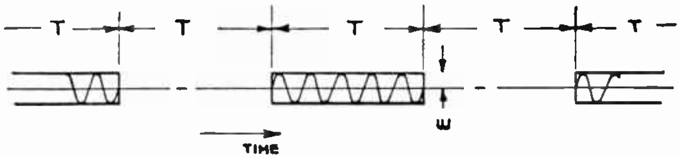


Fig. 18.

TABLE II

Frequency	Voltage Magnitude	Watts in each side-band when peak power is 10,000 watts (see note)
$f - \frac{15}{2T}$	$\frac{E}{15\pi} = -0.0212E$	4.5
$f - \frac{13}{2T}$	$\frac{E}{13\pi} = 0.0245E$	6.0
$f - \frac{11}{2T}$	$\frac{E}{11\pi} = -0.0289E$	8.35
$f - \frac{9}{2T}$	$\frac{E}{9\pi} = 0.03535E$	12.5
$f - \frac{7}{2T}$	$\frac{E}{7\pi} = -0.0455E$	20.7
$f - \frac{5}{2T}$	$\frac{E}{5\pi} = 0.0637E$	40.6
$f - \frac{3}{2T}$	$\frac{E}{3\pi} = -0.1062E$	112.8
$f - \frac{1}{2T}$	$\frac{E}{\pi} = 0.3185E$	1012.0
f (carrier)	$\frac{E}{2} = 0.5E$	2500.0

$f+\frac{1}{2T}$	$\frac{E}{\pi} = 0.3185E$	1012.0
$f+\frac{3}{2T}$	$\frac{E}{3\pi} = -0.1062E$	112.8
$f+\frac{5}{2T}$	$\frac{E}{5\pi} = 0.0637E$	40.6
$f+\frac{7}{2T}$	$\frac{E}{7\pi} = -0.0455E$	20.7
$f+\frac{9}{2T}$	$\frac{E}{9\pi} = 0.03535E$	12.5
$f+\frac{11}{2T}$	$\frac{E}{11\pi} = -0.0289E$	8.35
$f+\frac{13}{2T}$	$\frac{E}{13\pi} = 0.0245E$	6.0
$f+\frac{15}{2T}$	$\frac{E}{15\pi} = -0.0212E$	4.5

Note: By a peak power of 10,000 watts, we mean that E is of such a value that the power in the wave would be 10,000 watts if the sinusoidal wave were continuous, not interrupted for the period, T. Because the wave is cut off half of the time, the actual total power for the wave shown in Fig. 2D will be 5000 watts for a peak power of 10,000 watts. If, in Table II, we sum the power in the carrier, the first eight upper side-bands, and the first eight lower side-bands, we find a power of 4939.9 watts.

The power absorbed by the filter resistors may be readily computed from Table II. We assume that the carrier frequency is 45.25 megacycles. Then we assign a definite value for T. We then turn to Table II and see which side-bands lie between 41 and 44 megacycles and sum up the power which is assigned to these side-bands. The results of this summation are shown in Figure 19 as a function of the time interval T. We see particularly that for vertical bars which last more than 0.4 microsecond, the filter resistors must handle less than 200 watts for a peak power of 10,000 watts.

Evidently, the severest condition occurs when the transmitted picture consists of vertical bars which are between 0.36 and 0.4 microsecond wide. Then, the filter resistors must be able to handle 1125 watts. It seems reasonable to assume that average picture conditions will be closer to the 200-watt region.

It is interesting to note that 100 per cent modulation of the 2500-watt carrier, at a modulating frequency of 2.25 megacycles, causes only 625 watts to be sent into the filter resistors.

We now have sufficient data to examine the interesting case of two television transmitters operating in adjacent bands. Specifically, the No. 1 channel* lies from 44 to 50 megacycles with picture carrier at 45.25 megacycles, while No. 2 channel lies between 50 and 56 megacycles with picture carrier at 51.25 megacycles. From Table II we can

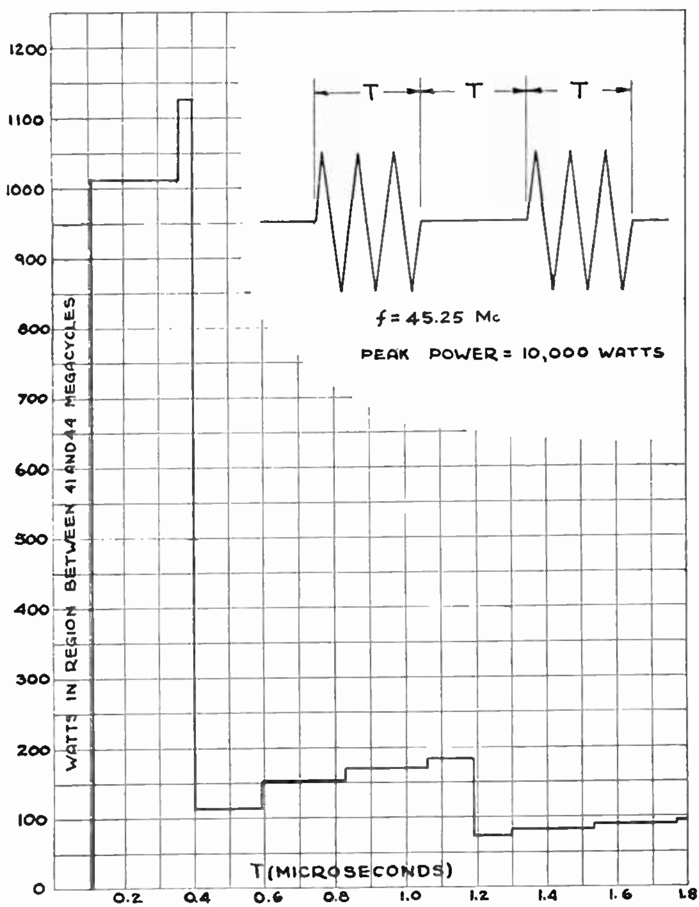


Fig. 19.

easily obtain the side-band energy distribution of either transmitter. We will consider each transmitter to have a peak power of 10,000 watts. In constructing Figure 20 we assumed each transmitter to be

* Since this paper was written, the television channel assignments have been changed. All references to channels in this paper are based on the old channel assignments that existed in 1939.

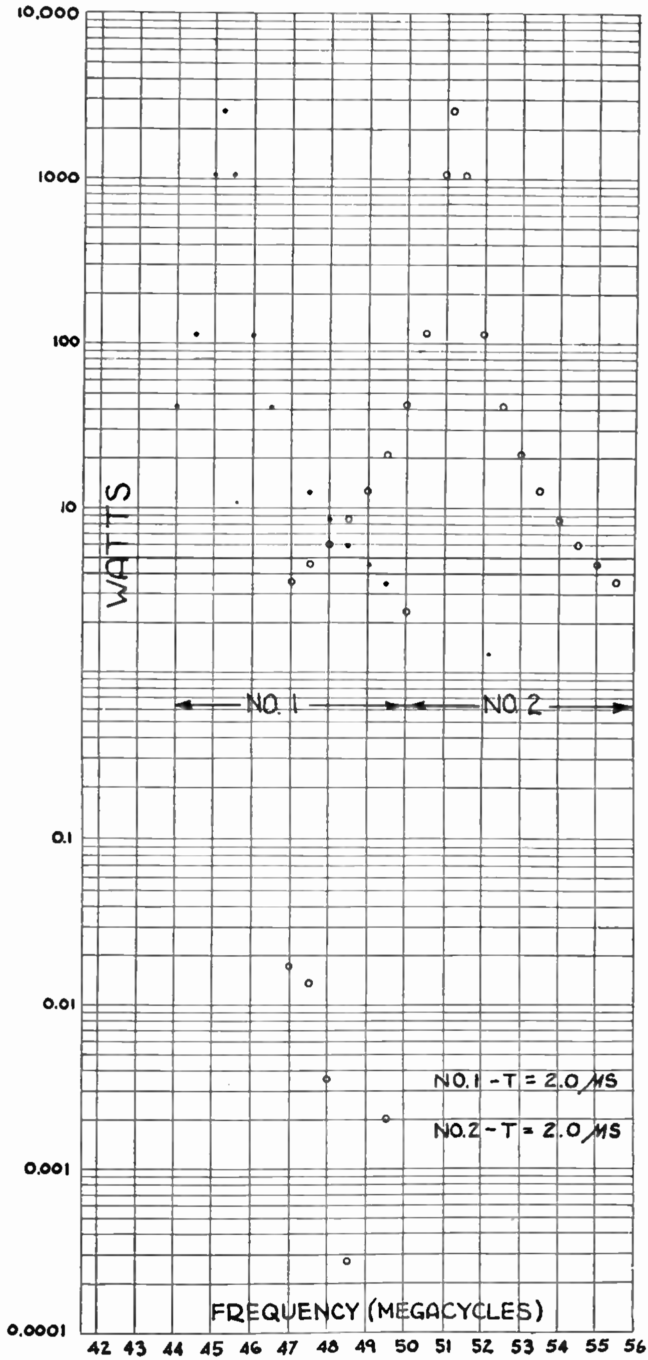

10,000
1000
100
10
0.1
0.01
0.001
0.0001
WATTS
NO. 1
NO. 2
NO. 1 - T = 2.0 μS
NO. 2 - T = 2.0 μS
FREQUENCY (MEGACYCLES)
42 43 44 45 46 47 48 49 50 51 52 53 54 55 56

Fig. 20.

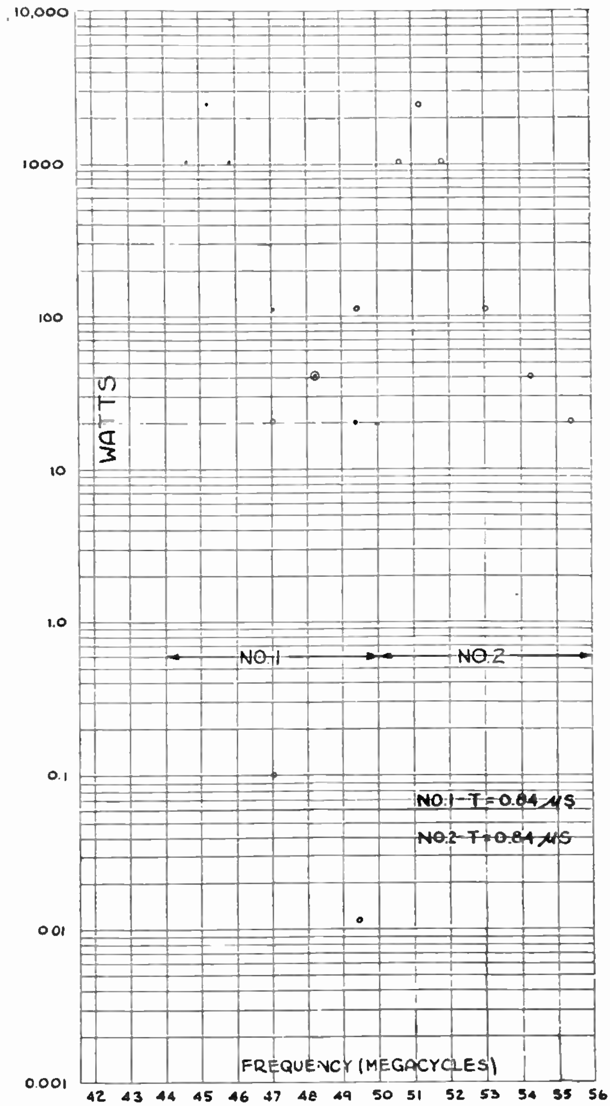


Fig. 21.

sending out a picture made up of vertical alternate black and white bars, the duration of each bar being 2.0 microseconds. The dots on Figure 20 indicate the energy distribution throughout the No. 1 channel due to the No. 1 transmitter. The circles show the energy distribution due to the No. 2 transmitter. We see that the No. 2 transmitter would radiate a strong signal in the No. 1 channel. The lower set of

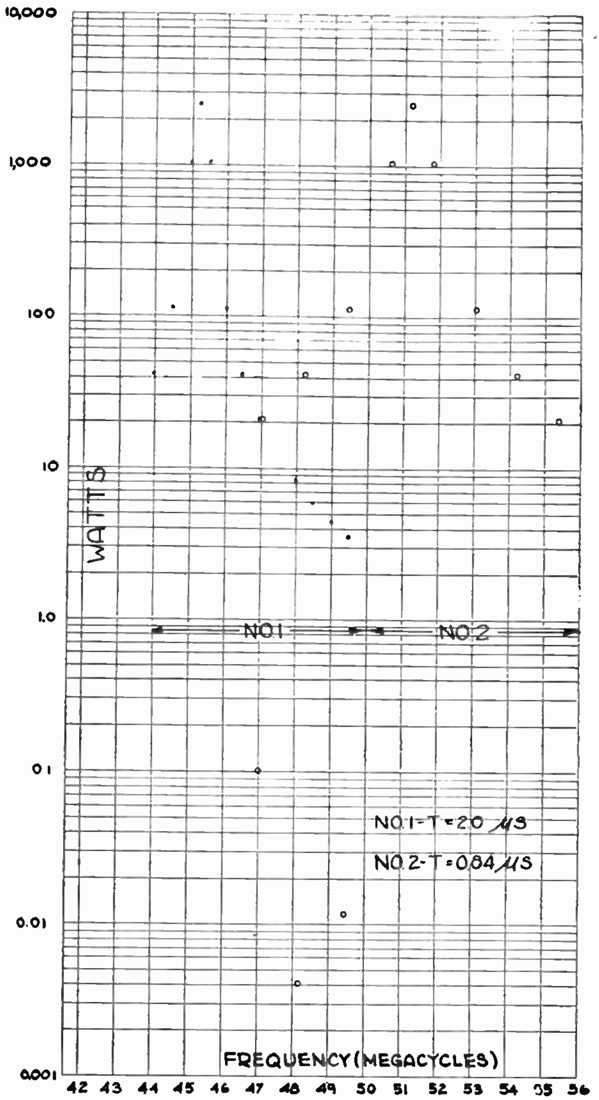


Fig. 22.

circles on Figure 20 shows the disturbing energy when the side-band filter is in operation. In this case there seems to be ample protection.

Figure 21 is similar to Figure 20 except that the width of the modulating bars is now 0.84 microsecond. In Figure 22 the width of bars for the No. 1 transmitter is 2.0 microseconds while No. 2 transmitter is modulated with 0.84 microsecond bars. The need for a side-band filter is quite evident from these figures.

APPENDIX II

INPUT IMPEDANCE OF TYPE A FILTER

For ease of reference, the Type A filter schematic diagram is repeated in Figure 23. We have already shown how transmission line elements may be used to take the place of the lumped elements shown in Figure 23. To compute the various circuit relations we use the following equations.

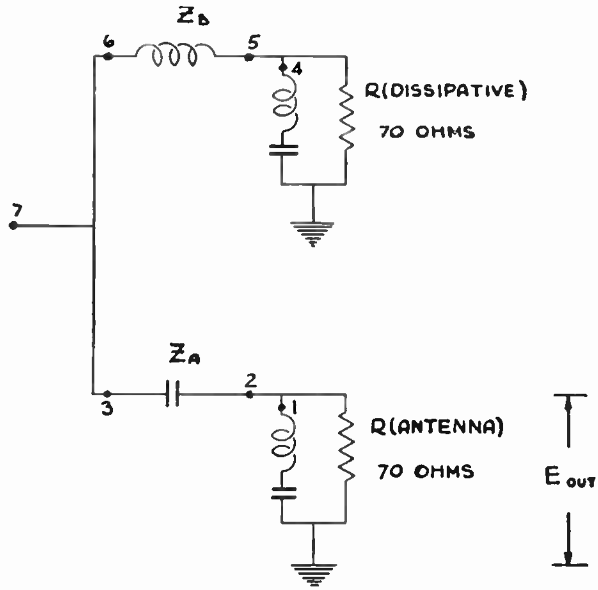


Fig. 23.

$$Z_A = -j70 \cot\left(\frac{f_{mc}}{44} \times 45°\right)$$

$$Z_B = +j70 \tan\left(\frac{f_{mc}}{44} \times 45°\right)$$

$$Z_1 = +j157 \tan\left(\frac{f_{mc}}{43} \times 540°\right)$$

$$Z_4 = +j157 \tan\left(\frac{f_{mc}}{45} \times 540°\right)$$

We may then compute the impedance at the input or at any other point. At the same time we may compute the attenuation through

TABLE III

f_{mc}	Z_A	Z_B	Z_1	Z_2	Z_3	Z_4	Z_5	Z_6	Z_7	Per Cent Reflection	$E_{out}/\sqrt{P}$
41	—j77.8	+j63	—j70	35—j35	35—j112.8	—j174.5	60.4—j24.2	60.4+j38.8	70—j2.73	1.95	0.42
42	—j75.2	+j65.2	—j35	14—j27.9	14—j103.1	—j119	52.2—j30.7	52.2+j34.5	68.5—j3.5	2.7	0.30
43	—j72.5	+j67.6	0	0	—j72.5	—j70	35—j35	35+j32.6	65.5+j1.9	3.65	0
44	—j70	+j70	+j35	14+j27.9	14—j42.1	—j35	14—j27.9	14+j42.1	70.2+j0	0	0.707
45	—j67.6	+j72.5	+j70	35+j35	35—j32.6	0	0	+j72.5	65.5—j1.9	3.65	1.0
46	—j65.2	+j75.2	+j120.5	52.4+j30.25	52.4—j35	+j35	14+j27.9	14+j103.1	68.5+j3.5	2.7	0.98
47	—j63	+j77.8	+j186.5	61.8+j23	61.8—j40	+j70	35+j35	35+j112.8	71.6+j3.6	2.8	0.91
48	—j61	+j80.5	+j295	66.2+j15.8	66.2—j45.2	+j114	50.7+j31	50.7+j111.5	73+j2.08	2.4	0.85
49	—j59.7	+j82.2	+j588	69.5+j8.2	69.5—j51.5	+j174.5	60.5+j24.2	60.5+j106.4	75+j1.2	3.5	0.815

TABLE IV

f_{mc}	Z_1	Z_2	Z_3	Z_4	Z_5	Z_6	Z_{in}	Per Cent Reflection	$E_{out}/\sqrt{P}$	$E_{res}/\sqrt{P}$
41	0.0428∠+87.5°	18.55∠+88°15′	+j0.0428	+j18.55	1∠+3°	1∠—1°	1.0∠+2°	1.7	1.0	0.04
43	0.259∠+74°50′	3.86∠—74°50′	+j0.269	+j3.72	0.965∠—15°	1.035∠+15°	1.0∠—5°	4.3	0.965	0.26
43.5	0.638∠+50.5°	1.565∠—50.5°	+j0.825	—j1.21	0.77∠—39.5°	1.298∠+39.5°	1.0∠0°	0	0.775	0.64
43.75	1.0∠0°	1.0∠0°	∞	0	0	∞	1.0∠0°	0	0 (0.025)	1.0
44	0.717∠—44°10′	1.392∠+44°10′	—j1.03	+j0.97	0.697∠+45°50′	1.435∠—45°50′	1.0∠0°	0	0.63	0.73
46	0	12.7∠—90°	0	—j12.7	0.998∠—4.5°	1.0∠0°	0.998∠—4.5°	3.9	1.0	0
49	0.135∠+82.5°	3.16∠—86.5°	+j0.135	—j3.14	0.952∠—17.5°	1.17∠+15°35′	1.18∠—6°	9.65	0.99	0.14

the filter in a very simple manner. The power into the antenna resistor is

$$P_{out} = \frac{E^2_{out}}{70}$$

Also, this power is equal to the power fed to this network at Point 3 so that

$$P_{out} = E^2_{in}\, G_3$$

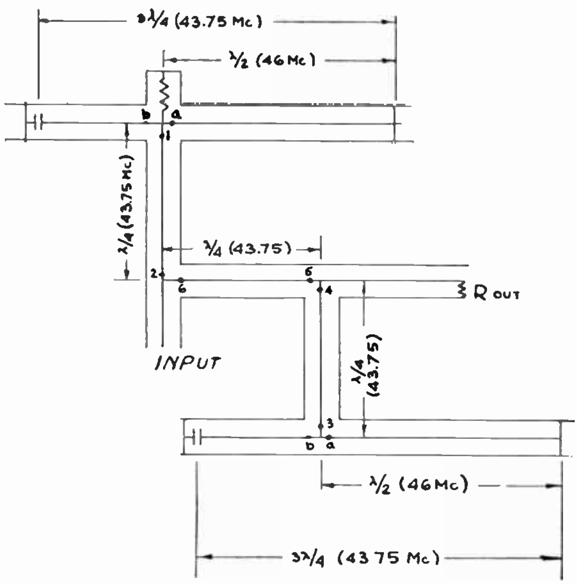


Fig. 24.

where G_3 is the conductance at Point 3. Then

$$E_{out}/E_{in} = \sqrt{70G_3}$$

The power into the total sytem is

$$P_{in} = E^2_{in}\, G_7$$

For a constant power input,

$$\frac{E_{out}}{\sqrt{P}} = \sqrt{\frac{70G_3}{G_7}}$$

We see that Z_7 is practically a pure resistance of 70 ohms, so that G_7 is close to 1/70 in value. Under this condition a constant input voltage corresponds to a constant input power so we may use either equation to determine the output voltage. The results of these calculations are given in Table III. Also, we have included in this table, a column labelled "Per Cent Reflection." This is the amount of reflected voltage wave given in terms of the incident voltage wave on the concentric feed line to the filter.

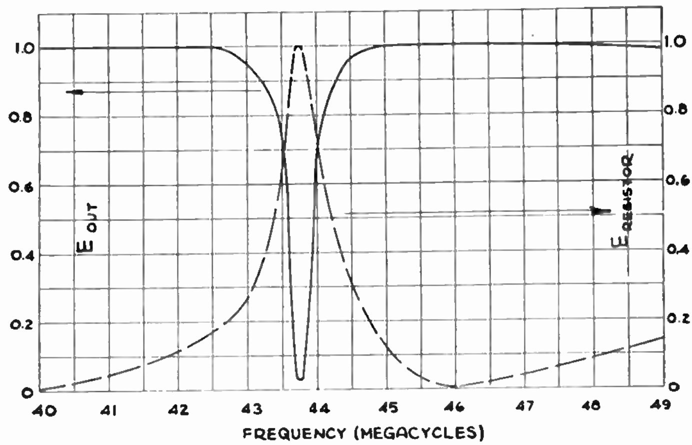


Fig. 25.

APPENDIX III

Input Impedance of Type B Filter

The Type B filter is shown in Figure 24. In computing the impedances of this filter, we may simplify the work by remembering that the resistors have the same value as the characteristic impedance of all the line elements. Then we may take the resistor value and the characteristic impedance as unity. Table IV shows the various impedances involved, in terms of unity starting resistance. For 70 ohm lines and resistors we simply multiply the values in Table IV by 70. Figure 25 shows the calculated values of the voltage on the output and the voltage on the dissipative resistor as a function of frequency. In making these computations, we made use of the relations

$$Z_a = +j1.0 \tan\left(\frac{f_{mc}}{46} \times 180^\circ\right)$$

$$Z_b = -j1.0 \cot\left(\frac{f_{mc}}{43.75} \times 99^\circ\right)$$

CONTEMPORARY PROBLEMS IN TELEVISION SOUND*†

BY

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New York, N. Y.

Summary—The present rapid development of television is introducing new problems in sound pickup and operation. As the art progresses, engineering tools and methods must not only keep pace with, but generally anticipate, the needs of the program-producing staff in the production of more and more intricate material. The nature of the acoustic problems so raised, and their solutions, are treated in this paper. New tools necessary to proper operation and the methods of their employment are discussed. For a better understanding of television requirements, the methods normally employed in motion pictures and standard radio broadcasting are compared with those in use in the present television studio. Some indications as to what may be required in the near future are discussed and possible developments suitable for such use are described.

IN THE history of every new activity, problems and concepts peculiar to itself arise. Certainly television is no exception to this rule nor is that part of television which we are to consider. There may have been many who felt in the earlier days of the art that television's sound accompaniment could well be expected to care for itself, for much had been done to perfect a technique of sound pickup with action in progress in the motion-picture studios of Hollywood. But very shortly, marked departures from the accepted methods were found desirable, and gradually it became clear that good television sound required not only different treatment but also different tools than were used at first. As the show-producing workers in television become familiar with their picture-making equipment, more and more is being demanded of it, and the sound accompaniment must keep pace. No consideration of the sound portion of a problem arising in a television studio is permitted to interfere with the picture technique, since the production staff has come to rely upon the sound engineer to find a way around his difficulties. This paper discusses these difficulties, and considers what may be done to overcome them.

A consideration of the mechanics of television studio operation will disclose some of the problems arising in sound pickup associated with

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visual programs. The National Broadcasting Company's studio is equipped at present with three television cameras and normal set lighting requires the use of four floor broads of about 3 kilowatts each. All of this equipment must be positioned for best advantage as to camera angles and lighting effects. If no sound equipment were used at all, the portion of the studio in use would be crowded enough, but it is necessary for the microphone boom to find a place also. The boom operator chooses his position with regard not only to his own best sound requirements but also considers the possible camera movements. If it is likely that a camera-dolly movement will find him in its way, he must be able to move the base of the boom sufficiently in advance of the dolly to clear the necessary space. Thus, the boom operator must

Fig. 1—For good pictures, television cameras require most of the space. Sound equipment must operate in what remains.

not only follow closely the action on the set but must also bear in mind the exact pattern of off-stage activities. The present operators have become adept at maintaining the position of the microphone correctly above the heads of the persons on picture, while stepping from the boom platform and moving it bodily a sufficient distance to permit passage of a camera. Often, too, only a few seconds can be allowed for a complete change from one set to another, necessitating accurate planning of movements and precise co-operation between sound- and sight-equipment personnel. To aid in this the boom used is as small as is presently practicable, having a maximum extension of 14 feet and being about 4 feet wide across the base. A unidirectional microphone is used to aid in reducing off-set noise, but this adds to the precision necessary in operation, for if close-ups are being used, the microphone must be aimed at the person being televised. This means that the boom operator must watch the camera-switching lights and position the microphone to suit the camera as well as the actor, being careful to discriminate against off-camera sounds only.

To facilitate scene transitions, or to provide a second pickup in a set where two widely spaced sound sources act concurrently, a method of hanging microphones has been devised. The studio ceiling carries a network of pipes of approximately 2½ inches in diameter. A special clamp has been made to fit these pipes. Connected to each clamp is an adjustable length of light conduit, designed to accept a standard

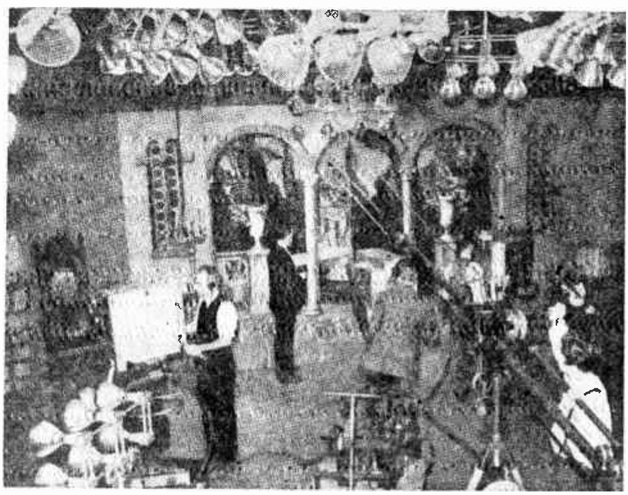

Fig. 2—Efficient utilization of floor space is a necessity in television.

microphone coupling. The clamp can be operated by twisting the conduit making it unnecessary to climb ladders to hang microphones; this greatly increases the all-important factor of speed.

Three types of microphones are normally used in the National Broadcasting Company's television studio. The unidirectional type with a cardioid pickup pattern is used for dialogue, mainly because of its ability to reduce the effect of off-stage noise. Television, unavoidably, has rather more of this than is used on a motion-picture sound stage, since following scenes must be prepared, equipment moved continuously to new locations, and the show kept running generally. This contrasts markedly with the complete stopping of all other activity when a scene is made in motion pictures. Regular velocity microphones

Fig. 3—A transition from one scene to another may require the use of a fixed-position microphone for opening the new scene. Action will be restricted until the arrival of the boom microphone.

are used in cases requiring more reverberation, or when convenient to use both sides for pickup. Usually this occurs when music is used on the set, and an acoustically bright effect is desired. A pressure microphone is used when its nondirectional characteristic is advantageous. The production staff at NBC recognizes that in recent years a micro-

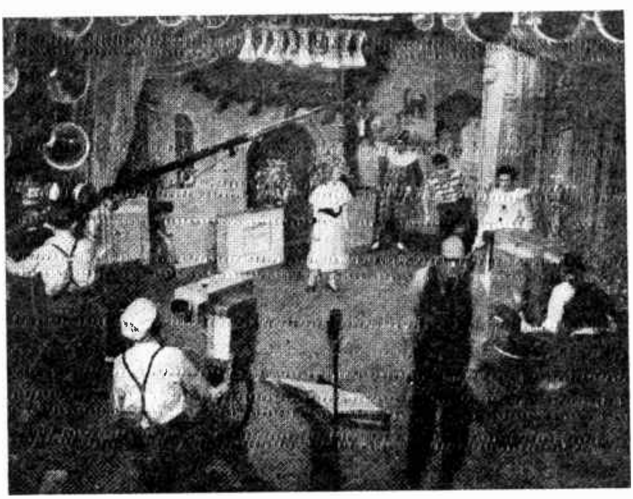

Fig. 4—Musicians must be close to the set for good musical coordination, introducing problems in balance and overlapping pickups. Unidirectional microphones aid greatly in such situations.

phone has become an integral part of some scenes. A supper-club set may call for several microphones, and if a grouping dictated by picture requirements is too wide for other types, a pressure microphone will solve the problem. As these microphones are relatively small, they are also most suitable for use in positions where they might tend to obscure the picture, or when a microphone must be held in the hand.

Even with the above variety of tools, situations arise that defy ordinary "on-the-spot" pickups. These cases generally can be classified into those in which high scenes limit the possibility of bringing a microphone close to the action, and those in which the action is too fast or too complicated to permit its being followed by the micrphone boom. Both occur usually in the musical production type of scene. It may be that a large and decorative background has been erected for a solo song, center stage and low. Obviously, no reasonable balance can be obtained between voice and accompaniment if the microphone must be far enough away to be out of the picture when it includes so large a backdrop. In the second case, trouble usually is encountered when performers not only sing but also move through a routine of action not suited to sound pickup. This may include singing while facing away from the camera, or while moving through a doorway, or perhaps next to percussion instruments of an orchestra where maintenance of balance would be impossible. All of these situations call for prerecording, a technique developed in Hollywood and happily adaptable to television. Two methods of procedure are available. In the first type mentioned above, the microphone is located in a suitable position for the making of the record, usually several hours before show time. The action is carried out as usual and the timing of the record automatically fits the scene as it will be broadcast. When the actual show takes place, a cue from the production director will start the record and kill all sound pickup in the studio. The record is then not only put on the air, but also fed back into the studio, where the performers can hear it, and synchronize their actions to it. When the recorded portion ends, the studio microphones are opened and the show continues normally. In the second type mentioned, the action is too detrimental to sound pickup to permit recording with it in progress even though no picture is required. Hence the action is carefully timed and cues noted. The recording will then be made without action, the setup being entirely to suit the sound situation. Such a record is then checked for synchronism on another rehearsal, and used "on the air" as described. A lacquer disk recording with the NBC Orthacoustic characteristic is used, resulting in transitions from direct pickup to record and back again with practically no noticeable change in sound quality. With such satisfactory matching of sound quality available, prerecording is a very useful tool in television.

Another angle of the studio-mechanics problem is in peculiar contradiction to the case in motion pictures. In some instances, the motion-picture-making equipment causes some trouble through making noise which may interfere with the desired sounds. In television the reverse is true. Sound in the studio may be of such intensity and frequency that it will cause spurious signals due to microphonics to appear in the picture. These generally consist of horizontal bars across the picture, and result from vibration of elements of vacuum tubes used in the video preamplifier in the camera. It is necessary to treat the television camera to keep sound out, rather than in. A heavy material, similar to roofing felt may be cemented to the inside surfaces of the camera housing to reduce sound transmission, and particularly to damp vibration occurring in the large plane sections of the present camera's sides and top. Without such damping, these parts will vibrate very heavily at their natural periods, making sound crossover almost a certainty. With sufficient loading the tendency to vibrate disappears almost completely, permitting operation with any normal studio sound level.

The very nature of television is that its appeal must be in the intimate manner. As long as the present methods of picture reproduction are being used extensively, this will continue to be the case, for picture size and detail make best use of close-ups and penalize the extreme long shots. The sound that accompanies these pictures should partake of the same quality, heightening the tone and mood of a scene. The methods adopted and the tools used must, then, be suitable for such work.

The National Broadcasting Company's live-talent studio is a room 30 x 50 x 17 feet. Its acoustic treatment differs radically from what a motion-picture engineer might expect to find on a sound stage, in that the reverberation constant is not as short as it could be made, but rather a variable quantity, being in some cases as long as 1¼ seconds, and in others as short as ½ second (over the essential range of frequencies). The reasons for this are close to the heart of the television problem. In the usual sound-stage case, the studio is a large acoustically dead room, in which relatively permanent sets are erected. It is normally the intention to permit those sets to exhibit their own characteristic reverberation without much, if any, artificial reinforcement. The case in television is somewhat different. Our sets are designed for rapid scene changes, and efficient use of personnel. They are made of linen stretched on wood frames in the manner of legitimate stage scenery. Instead of adding a lifelike reverberation to the sound originating in them, such sets produce undesirable low-frequency resonance effects, and add large amounts of high-frequency absorption in their unpainted surfaces. If the studio itself were very "dead" these effects would add detrimentally. Dialogue equalizers are used, which help to avoid this trouble, but the less equalization that can be employed, the better will be the average sound quality.

Studio acoustics also play an important part in television sound for other reasons. The volume of the sets in use is always a very large portion of the total studio volume, since many scenes must be set up at once to provide a continuous performance. Under usual conditions almost the whole studio is used in a show to run an hour and a half. With so much absorption added in the sets, much of the original treatment of the studio must be removed to produce anything like normal reverberation. Most television shows will also present music as well as speech in the same studio, without a pause between the two portions of the program. Such a case in motion-picture production would call for the use of a scoring stage, or a set especially treated for music. In television, the problem is attacked by making large sections of the acoustic treatment on the studio walls movable. These panels can be opened to expose a hard, reflective surface, increasing the reverberation to an acceptable level. Should an outdoor scene be required, however, all the absorbing panels would be closed, and equalization added to produce an essentially reverberation-free pick-up.

It has often been remarked that television even now should use large studios of the motion-picture sound-stage variety. There are however certain mechanical and acoustical considerations that make this doubtful. Present television practice, which demands many close-ups and rather restricted action during most of the show, means that even with a relatively large set, for the major portion of a "take," the cameras, lights, and sound equipment must be crowded together to serve best the particular portion of it used at the moment. It is a provoking fact that although most of the studio may be empty, the television equipment must be worked in close quarters. Consequently, additional room would not materially increase the freedom of action of the cameras as far as any one set is concerned. Mechanical considerations, then, indicate that the size of the studio is determined by the number of sets which reasonably can be used on one show, or can be served by one group of equipment. Under present production conditions, this would result in a studio considerably smaller than the larger motion-picture sound stages. Acoustically, the smaller studio is desirable, because of the requirement mentioned above that studio acoustics be adjustable to compensate for set absorption. If the studio becomes too large, it cannot contribute usefully to the over-all sound quality, for reverberation as a desirable enhancement is replaced by what is commonly called room-slap, or echo. Hence, if a very large studio is to be used, it *must* be very "dead," which inflexibility seriously limits its usefulness as an acoustic tool. The answer to this problem seems to be that television studios should be of a size between those used in radio broadcasting and the large stages of Hollywood if all the mechanical and acoustical advantages are to be realized.

The excellent work currently done in the broadcasting studios has raised to a high stage of refinement the art of producing mood and atmosphere with sound. Television must offer at least as much facility for creation of these effects and at the same time must not limit in any way the freedom of action necessary to good pictorial effect. Some of the problems encountered in this blending of sound technique and sight productions are worthy of consideration.

In the television studio, both close-up and long shots must be taken at the same time. The accompanying sound must not only suit the apparent distances shown in the picture, but may also be required to produce an effect complementary to it. At times, perspective in television sound is so important that what would normally be only a medium long shot can be made to *seem* very long, if the sound which accompanies it carries sufficient reverberation to suggest great distance to the mind of the listener. Since actual long shots are not usually permitted for long periods of time, such an aid in producing the effect of distance is a valuable tool. Close-ups, of course, require intimate sound, and often the change from a distant view to a close-up occurs too quickly to permit any actual change in microphone placement or acoustic treatment, so the *effect* of a change must be produced electrically. Reverberation once added cannot be deleted; consequently, the pickup conditions must be set to produce close-up sound. It is then possible to add reverberation through the use of the standard echo-chamber method. In the studio, close-up cameras are provided with long-focal-length lenses, and long-shot cameras have either normal- or short-focus lenses. Switching between the cameras actuates a set of relays so connected that amounts of reverberation and volume level can be adjusted to suit the lens of the camera in use. This is accomplished by providing for each camera a separate volume control, and a separate reverberation control. If a camera is to be used to take a long shot, the volume control associated with it is turned down an amount calculated to produce the proper psychological effect, and the reverberation control is opened to accept a large portion of the output of the echo chamber. Another camera having an intermediate focal length would use more direct sound and less reverberation, while the close-up camera would use full volume and no feed from the echo chamber at all. When the technical director switches from one camera to another, the sound is also switched from one set of controls to another, producing instantaneous changes in sound quality to suit the picture requirements. Of course such artificial correction is confined in its application to interior shots which would normally exhibit acoustical characteristics similar to those available from the echo chamber. Corrections can be applied

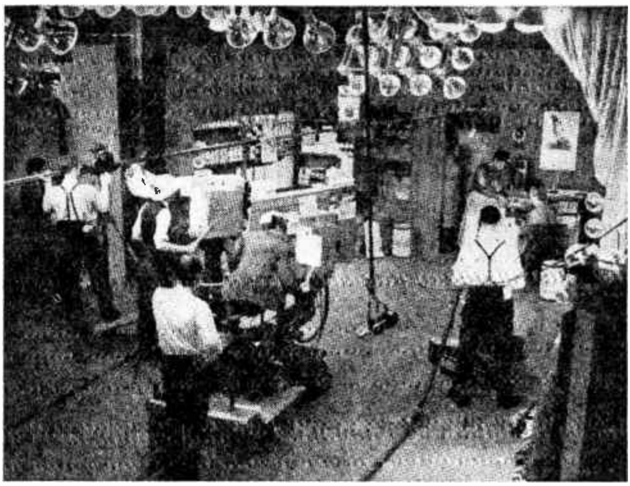

Fig. 5--Making close-ups and long shots simultaneously introduce problems of acoustic realism.

to outdoor scenes by changing the volume level and low-frequency response of the system to match the camera switching. Thus an exterior long shot would be accompanied by a reduction in volume-control setting and an increase of equalization designed to remove low and high frequencies, thus simulating the conditions obtaining in nature. Without such processing, the sound accompanying a television picture would not only lose valuable contributing effects, but at times might give an almost ludicrous effect, for the human eye and ear have been trained to expect a certain correlation between sight and sound perspective, and violations of their normal relationships are not acceptable.

Another problem peculiar to television sound is the result of a demand for realism in its dynamic range. In motion pictures, the acceptable range of loudness is a strictly measurable and controllable quantity. The lowest modulation permitted is a function of track hiss, frequency response, audience noise, etc., and the highest sound output is determined by the 100 per cent modulation point and reproducer power. No one in the audience is able to change the volume reaching him, nor does he expect to hear sound which is not dimensioned to fit the picture on view. In radio broadcasting, almost the exact opposite is true. With no visual program, the listener demands the maintenance of a relatively constant level, and often writes to his station complaining that he has to adjust his receiver volume control during the progress of a show. Television encounters portions of both of these troubles, and has had to evolve its own operating procedures to combat them. Since television is broadcast, a reasonably high average modulation should be maintained, in order that receiver noise levels may be low. Maximum deviation is determined, of course, by channel width. Within these two extremes must be confined sound to suit anything from the scraping of a pen across paper, to the crashing thunder of a modern blitzkreig. Such matching of sound and sight is necessary, for if the eye sees what would in nature produce a loud sound, but the ear hears only a small, muted version of what is expected, the mind rejects both sight and sound as being counterfeit. Thus a dynamic range is required of television sound which is greater than absolutely necessary in sound broadcasting. Here the home receiver enters the problem. If dialogue, and other relatively quiet sounds, are broadcast at their proper level over a period of time, it is likely that the volume of the home receiver will be increased by the listener to match what has come to be expected of broadcast sound. Then, if full dynamic range is employed, the louder passages will exceed reasonable living-room power, or perhaps overload the receiver. Hence, some compression must be employed, yet without producing the above-mentioned unconvincing mis-match. Treatment of this problem has evolved into a skillful handling of audio levels in such a way as to produce changes in apparent loudness which are greater than those actually broadcast. If it is known in advance that some particular point in a performance will require a large increase in volume, the loudness of the passages preceding the expected increase in level is gradually lowered, the process sometimes extending over several minutes. This decrease in loudness is accomplished so slowly that it does not come to the attention of the listener and is in some degree compensated by what appears to be an increase in the listeners aural sensitivity. Then, when the large amplitude is required, an increase to maximum deviation is sufficient to produce an admirable effect. Of course, such a loud period causes the listener's hearing again to be reduced, and care must be exercised in returning to a medium or low level of modulation.

The television sound problems which have been discussed are a few of those that have already been encountered in television broadcast operation. They have increased in complexity as television program production has advanced its techniques. In an art developing as rapidly as television, no one can be certain that indicated trends will be followed or that present methods and materials will be adequate, or even useful, in the future. It is only by continuing the present close cooperation between the studio and the development laboratory that television's sound problems can be solved.

IMAGE ORTHICON CAMERA*†

By

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Summary—One of a series of developmental television cameras using the image orthicon is described. The complete camera weighs less than forty pounds. The input power required by the camera is 300 watts. This power may be supplied by a non-regulated power supply or generator. A unique regulated high voltage supply was developed for the electron multiplier and image section of the camera tube. The camera circuits include the deflection system, voltage regulators, black-level setting, blanking circuits, and video amplifiers. A total of seventeen tubes is used in the camera. An extremely high-sensitivity version of the camera, using reflective optics, is also described.

I. INTRODUCTION

THE development of the image orthicon[1] provided a camera tube for an extremely sensitive television camera. In addition, due to its high output signal level, it permitted a substantial reduction in the number of tubes used in the video amplifier, and thus permitted the incorporation of other circuits within the camera that were built into auxiliary equipments in previous types of cameras. While certain operating features of the image orthicon provide simple blanking and black level setting, the photo-cathode image section and multiplier electrodes require potentials of such values and stability that new circuits had to be designed to permit the incorporation of these supplies in the camera.

II. RESOLUTION

During the early part of the development the major effort was applied to improving the resolution of the image orthicon. In the course of the investigation, it was observed that by scanning only a portion of the target considerably better resolution was obtained than when the whole target area was scanned. In order to determine whether the lack of resolution was due to limitations in the scanning or in the image section of the tube, or possibly in the coupling section between the camera tube and the amplifier, a variable frequency signal was applied to the target. It was found that a 5-megacycle signal was satisfactorily passed by the scanning section and the amplifier, indicating

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* Decimal Classification: R583.12.

† Reprinted from *RCA REVIEW*, March, 1946.

[1] Paper on the image orthicon was presented by A. Rose, H. B. Law, and P. K. Weimer, at the I.R.E. Winter Technical Meeting, on January 24, 1946.

that the limitation was caused by the image section of the tube. The fact, however, that scanning a small portion of the target provided a well-resolved picture indicated that the image section itself formed a picture of satisfactory resolution. These experimental results tended to show that an interaction between the scanning and the image section was degrading the picture.

Upon the assumption that the horizontal scanning field was vibrating the electron image on the target, and thereby blurring the picture, a portion of the horizontal deflecting current was applied to an auxiliary coil located over the image section. This current had a direction opposite to that in the deflecting coil in order to cancel the variable component of the magnetic field. The experiment resulted in considerable improvement in resolution.

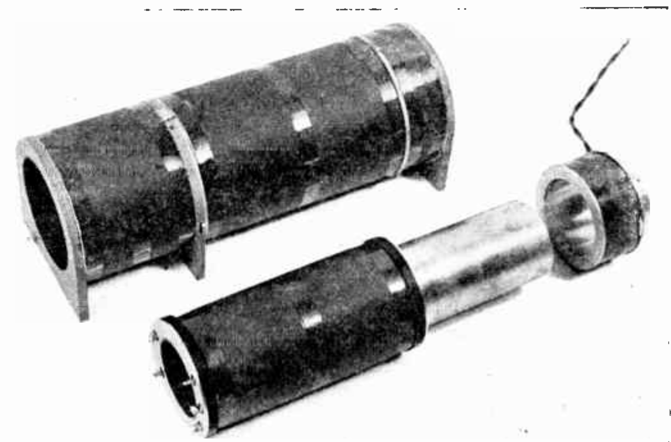

Fig. 1—Focusing, Deflection, and Alignment Coil Assemblies.

As another approach to the problem, this crosstalk effect was reduced by careful shielding. The problem was pursued further, since it was known that a reduction of the deflection power would proportionally reduce the effect. A simple reduction of the focusing field intensity allowed a reduction in deflection power, but it degraded the resolution around the edges of the picture, and hence was not permissible. However, it was found that by reducing the focusing field over the deflection coils and reinforcing it over the gun and the target, better resolution in the corners was obtained because of better electron landings. It was further found that the desired field distribution could be obtained with a uniformly-wound focusing coil and a magnetic shield (of iron wire) over the focusing coil. This method, with the addition of electrostatic shielding, was finally adopted. The arrangement considerably reduced the required deflection power and provided a resolution in excess of 450 lines under high light conditions. In cameras where a maximum resolution is required, the image section bucking coil was also provided.

Figure 1 shows the focusing coil assembly, the deflection coil assembly, and the alignment coil in the usual order with the shields. The shield which extends from the deflection coil over the gun end is to prevent pickup from the deflection coil by the signal lead.

III. The Circuits

A block diagram of the camera is shown in Figure 2. The video

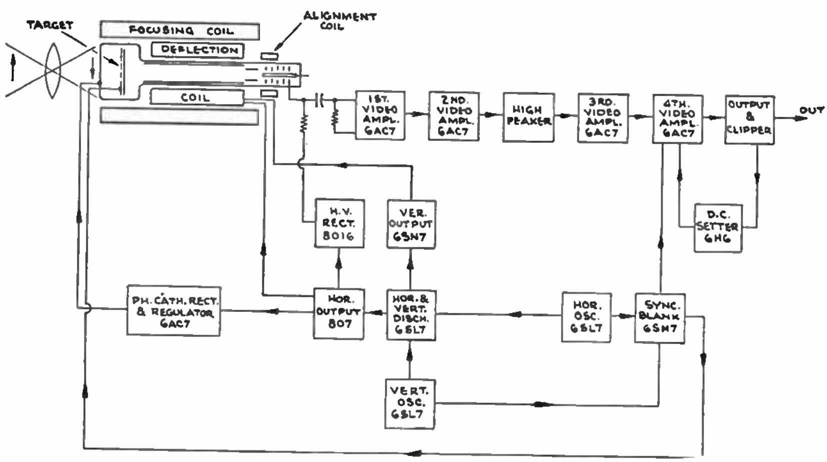

Fig. 2—Camera Block Diagram.

output is taken from the last, or fifth, multiplier of the orthicon across a 33,000-ohm resistor, through which a high potential of approximately 1500 volts is fed to the multiplier. This load resistance is about one-tenth of the conventional value used with iconoscopes. It is permitted by the higher signal current output of the image orthicon. The lower signal output resistance also permitted the use of a correspondingly reduced amount of equalization in the high peaker circuit[2] in the second video amplifier plate circuit. With the five stage multiplier image orthicon, substantially all the noise generated is due to the scanning beam, and with the reduced equalization in the high-peaking circuit there were no noticeable microphonics due to the amplifier system.

[2] U. S. Pat. No. 2,151,072—A. V. Bedford, March, 1939.

By using a clamping circuit at the fourth video amplifier stage to reinsert the low video frequencies, further assurance was taken to keep the camera free from microphonics generated in the amplifier.

The clamping circuit is shown in Figure 3, and it functions as follows:[3] At the input to the amplifier the video signal is given a reference level, such as black, during the horizontal return time. This reference level is readily obtained by applying pulses to the target of the image orthicon during the horizontal blanking interval. These pulses cause all of the scanning beam to return to the multipliers. This is a signal which is equivalent to black level. After this reference level is inserted in the signal, the low frequency response of the video amplifier can be reduced to the point where it will just pass a square wave corresponding to line frequency. At a high signal level, where all danger of microphonic disturbance in the amplifier tubes is passed, the signal at

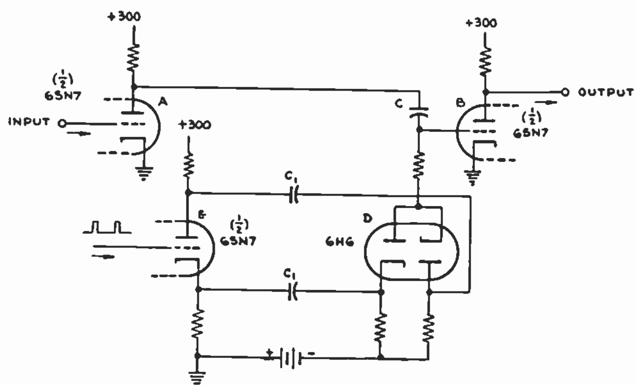

Fig. 3—Direct Current Setting Circuit.

the time of the black reference (which has become variable in level due to the presence of picture signal) is again established at a fixed value. With black level representing a fixed bias on the amplifier stage, it follows that the low frequency and direct current component of the signal are again present. Referring to Figure 3, the video signal which has lost the direct current and all low frequency components, passes from the plate circuit of tube A to the grid tube B through the small coupling condenser C. The grid leak on tube B is replaced by the two diodes of the 6H6 type. The push-pull pulses obtained from the tube E are applied to the diodes. The pulses cause both diodes to conduct. This is equivalent to connecting the grid of tube B to the battery through a switch. This makes the potential of the grid corresponding to black equal to the battery voltage.

[3] U. S. Pat. No. 2,299,945—K. R. Wendt, October, 1942.

The reconstructed signal is mixed with a blanking signal in the plate circuit of the fourth amplifier stage, then fed to a cathode follower output stage, which provides a complete video signal of approximately one volt peak-to-peak value.

The deflection circuit consists of the horizontal and vertical oscillators, the two discharge tubes in one envelope, and class A_1 type deflection output stages for both the vertical and horizontal deflection.

The high voltages for the image orthicon were obtained by rectifying the return sweep voltage of the horizontal output stage. Any change in the deflection voltage then tended to upset the operating conditions of the image orthicon tube. Since the photo-cathode voltage, in particular, is very critical, a simple voltage regulator was devised.

Owing to the fact that the current required was exceedingly small, the constant current property of a pentode was considered the simplest method of providing a constant voltage. A further improvement in regulation was obtained by applying a portion of the rectified potential

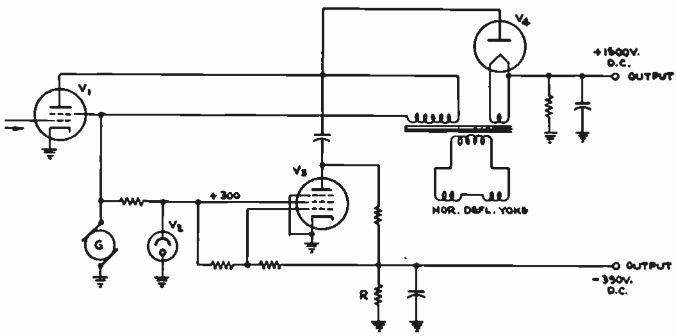

Fig. 4—High Voltage Power Supply and Regulator

to the control grid of the pentode rectifier and degenerating any change that might occur.

The circuit is shown in Figure 4. A portion of the high alternating current pulse voltage across the horizontal deflecting output transformer is rectified by the pentode V3. The useful direct current voltage supply then occurs at the negative terminal shown and is regulated by suitably controlling the grid voltage of the pentode. A portion of the output of the power supply G is regulated by the glow discharge tube V2 and is used for the screen supply to V3. This regulated voltage also serves as a reference potential for the control action, in that a portion of the rectified output voltage is subtracted from it and applied to the control grid. This arrangement will produce a large potential change of the grid voltage with small percentage change of the output voltage. When the negative potential tends to increase across the load resistance R, the grid becomes more negative and the resistance of the circuit increases, thereby reducing the potential across the load. The high voltage for the multipliers is supplied by the rectifier V4. The wall coating and persuader voltages are obtained from the voltage regulator V2 which is actually two VR-150 tubes in series.

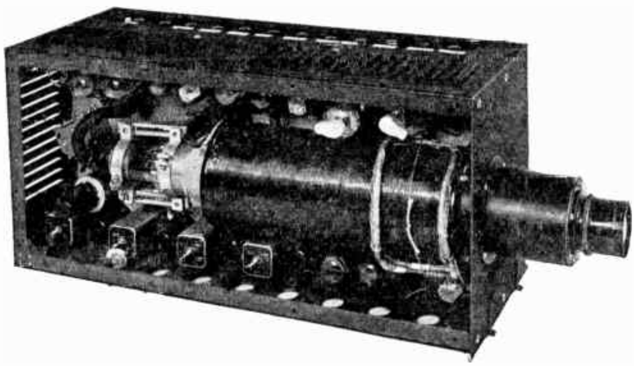

Fig. 5—Top View of Camera Chassis

Figure 5 shows the top of the camera chassis. The high voltage signal coupling capacitor may be seen in the left side of the picture. The video amplifier is located in the bottom row. The voltage regulators, high voltage supplies, and deflecting circuits occupy the top row.

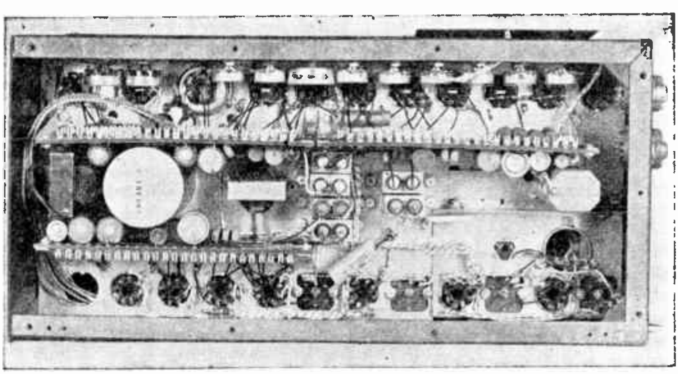

Fig. 6—Bottom View of Camera Chassis

The bucking coil to eliminate the image jiggling is on top of the focusing coil. A bottom view of the chassis, showing the circuit components, is given in Figure 6. The voltage divider for the electron multipliers is at the left side, the potentiometers at the top, and the deflection transformers at the left side of the picture. Four controls, namely, the scanning beam bias, the scanning section focusing control, the image section focusing control, and the amplifier gain control are readily accessible by the opening of a hinged lid. The other controls are normally covered with a plate fastened with screws.

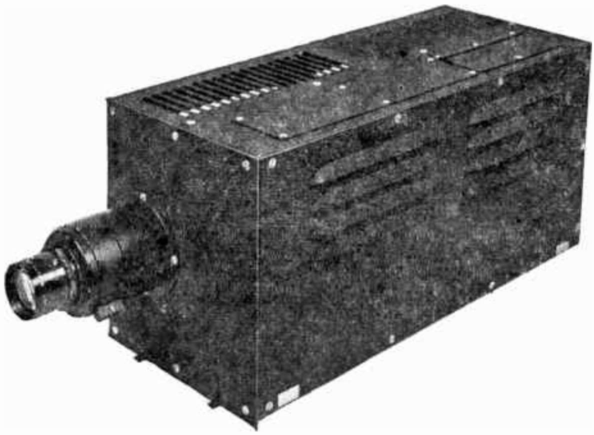

Fig. 7—External View of Camera Assembly with Lens

IV. The Camera Assemblies

Figure 7 shows an external view of the camera assembled with a 12 cm. f 2.7 lens. Figure 8 shows an image orthicon camera assembled with a reflecting Schmidt optical system. The photo-cathode surface of

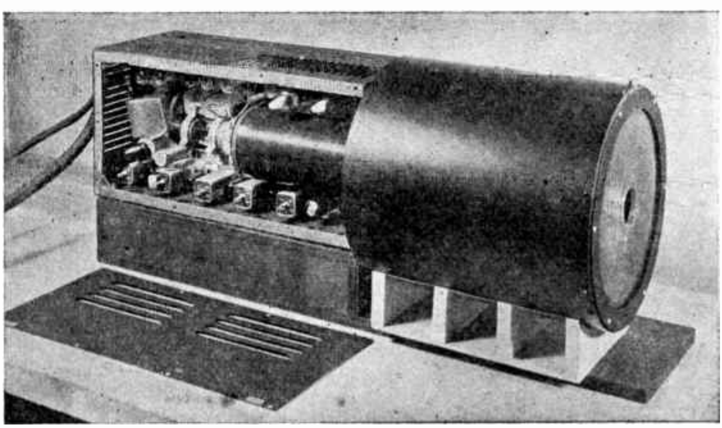

Fig. 8—Camera Assembly with Reflective Optical System

the image orthicon used in this camera was properly curved in order to secure proper focus of the optical image of the entire field of view, and it was placed approximately in line with the spherical mirror. The

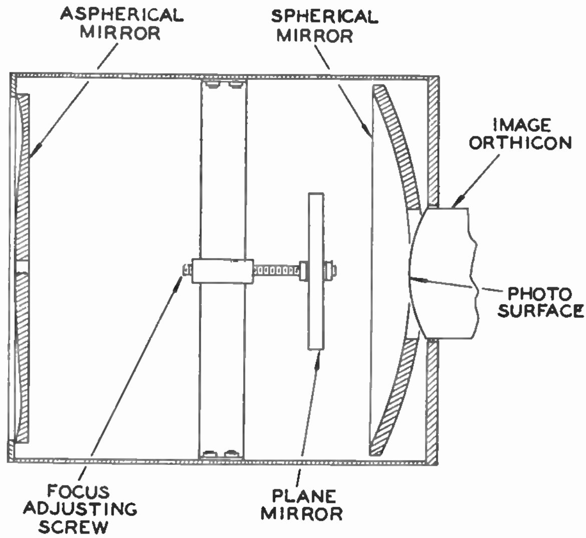

Fig. 9—Construction of the Reflective Optical System

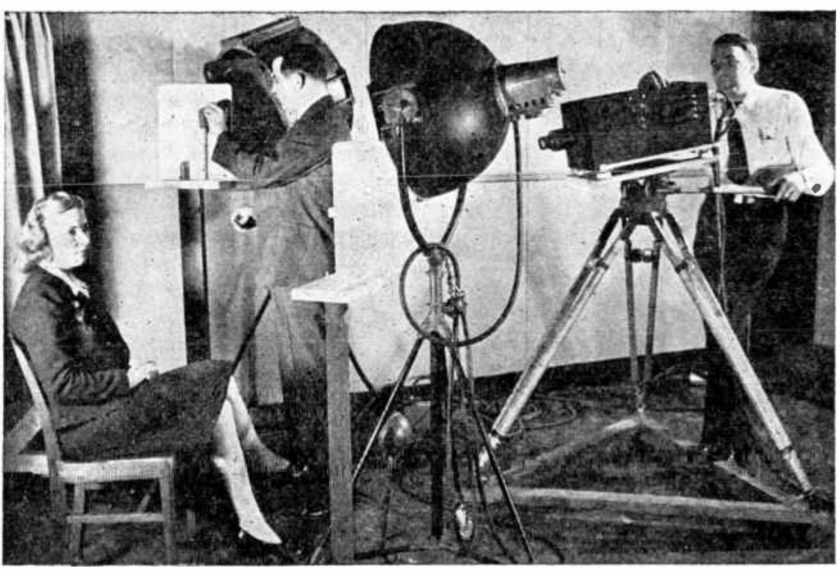

Fig. 10—Camera Demonstration Setup

design of the optical system is shown in Figure 9. A brass barrel provides a rigid structure for the system. The focusing is adjusted by the plane mirror which reflects the image on the photo-cathode of the image orthicon. The system has an (f) power of .7 and an aperture of 10 inches. The completed optical unit has a resolution of better than 1000 lines at the image surface.

Fig. 11—Television Picture Taken with the Subject Illuminated by 3 Kilowatt Incandescent Light.

Fig. 12—Picture with the Subject Illuminated by a 25 Watt Desk Lamp.

Fig. 13—Picture with the Subject Illuminated by One Candle.

V. Performance

Figure 10 shows a typical demonstration setup with the image orthicon camera using the f 2.7 lens. Lighting can be provided by the two one-and-a-half kilowatt reflectors, a 25 watt lamp, or by one to four candles. Figure 11 shows a picture taken from a 12-inch direct viewing monitor when the subject was illuminated by the two one-and-a-half kilowatt lights. Figure 12 shows the same subject illuminated with the 25-watt lamp, and Figure 13 shows the same subject with a single candle at a distance of three feet as the only source of illumination. The main difference between the last two pictures is in the noise present, which can not be seen in the photographs due to the inherent integration of the exposure.

The sensitivity of the Schmidt camera was found to be adequate to detect the presence of a test pattern in an incident illumination of 150 microfoot candles. For 200 line resolution of the test pattern, however, 1.5 millifoot candles were required.

ACKNOWLEDGMENT

The authors wish to acknowledge the help and suggestions given by the members of the television section of RCA Laboratories Division, particularly that of Mr. R. R. Thalner. Part of the work described in this article was carried out under contract between the Office of Scientific Research and Development and the Radio Corporation of America.

THE IMAGE ORTHICON — A SENSITIVE TELEVISION PICKUP TUBE*†

By

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Summary—*The image orthicon is a television pickup tube incorporating the principles of low-velocity-electron-beam scanning, electron image multiplication, and signal multiplication. It closely approaches the theoretical limit of pickup tube sensitivity and is actually 100 to 1000 times as sensitive as the iconoscope (1850) or orthicon (1840). It can transmit pictures with a limiting resolution of over 500 lines and, if properly processed, is relatively free from spurious signals. At low lights, the signal output increases linearly with light input; at high lights, the signal output is substantially independent of light input. The tube is completely stable at all light levels. The signal output is sufficiently high to make the operation of the tube insensitive to many of the preamplifier characteristics that are normally considered significant. The construction, operation, electron optics, and performance of the tube are discussed.*

I. INTRODUCTION

THE importance of sensitive pickup tubes to the success of a well-rounded television service needs little emphasis. One has only to be reminded that, insofar as the television pickup tube is called upon to replace the human observer, the sensitivity of the pickup tube should match that of the human eye. The demands on a television service are often more stringent than on news photography, for example. The latter can, within wider limits, select the times and conditions under which it will record pictures. The pickup tube, once committed to transmitting an event, such as a football game, must steadily transmit pictures under the whole gamut of lighting conditions. It is, accordingly, highly desirable to have a pickup tube which can transmit pictures both at very low and at very high light levels.

The iconoscope[1] has transmitted excellent pictures at high light levels; the orthicon[2] has operated best at medium light levels. The

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* Decimal classification: R583.6.

† Presented at the 1946 Winter Technical Meeting of the I.R.E. in New York, N. Y., on January 24, 1946. Reprinted from *Proc. I.R.E.*, July, 1946.

[1] V. K. Zworykin, G. A. Morton, and L. E. Flory, "Theory and Performance of the Iconoscope," *Proc. I.R.E.*, vol. 25, pp. 1071-1092; August, 1937.

[2] A. Rose and H. A. Iams, "The Orthicon," *RCA REVIEW*, vol. 4, pp. 186-199; October, 1939.

image orthicon extends the range still further toward lower illuminations by a factor of approximately 100. At the same time, the image orthicon can operate *stably* at medium and high light levels. Unlike the orthicon, it is not subject to transient loss of operation caused by sudden bursts of illumination. The use of the image orthicon in the higher light ranges is not, however, emphasized relative to the iconoscope or orthicon. The additional complexity of the tube needed to provide its increased sensitivity has not yet permitted pictures whose quality equals the best that the iconoscope or orthicon can transmit.

The present paper describes the construction, operation, and performance of the image orthicon. It is hoped to treat some of the electron-optical and constructional problems in more detail in separate papers.

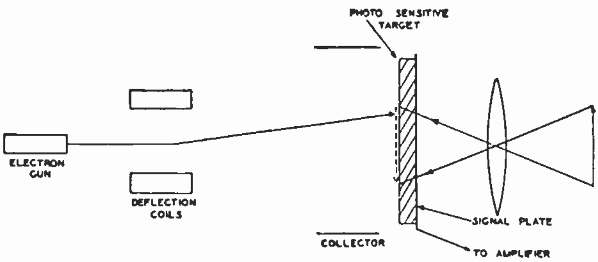

Fig. 1—Typical parts of storage type of pickup tube.

II. General Description of the Image Orthicon

The usual storage type of pickup tube (Figure 1) has an electron gun, a photosensitive insulated surface, referred to as the target, and a means for deflecting the electron-scanning beam. The scene to be transmitted is focused on the target on which it builds up by photoemission a charge pattern corresponding to the light and shade in the original scene. The beam of electrons, generated by the electron gun, is made to scan the charge image in a series of parallel lines. While a constant stream of electrons approaches the target, the stream which leaves is modulated by the charge pattern. A signal plate located close to the target surface picks up the modulation by capacitance and feeds it into the grid of the first amplifier tube. The same video signal, however, appears in the modulated stream of electrons leaving the target, and if these electrons could be collected on a single electrode, the signal could be fed through it into an amplifier.

The image orthicon (Figures 2 and 3) has, in addition to the usual gun, deflection means, and target, three parts that contribute to its

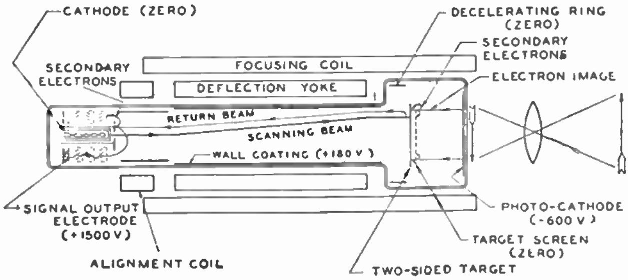

Fig. 2—Diagram of the image orthicon.

sensitivity and stability. An electron multiplier, built into the tube near the gun, multiplies the modulated stream of electrons returning from the target before it is fed into an amplifier. Sensitivity gains of 10 to 100 are thereby made possible. The charge pattern on the target, instead of being generated by photoemission, is formed by secondary emission from an electron image focused on the target. The electron image is released by light from the scene to be transmitted falling on a conducting semitransparent photocathode and is focused on the target by a uniform magnetic field. The combination of the higher photo-sensitivities that can be obtained for a conducting surface than for an insulated surface, together with the secondary-emission gain of the electron image at the target, provides another factor of about fivefold increase in sensitivity. The use of a separate conducting photo-cathode is made possible by a two-sided target in place of the usual one-sided target. The two-sided target allows the charge pattern to be formed on one side and the scanning to take place on the opposite side. Further, it permits the tube to operate stably over a large range of scene brightnesses.

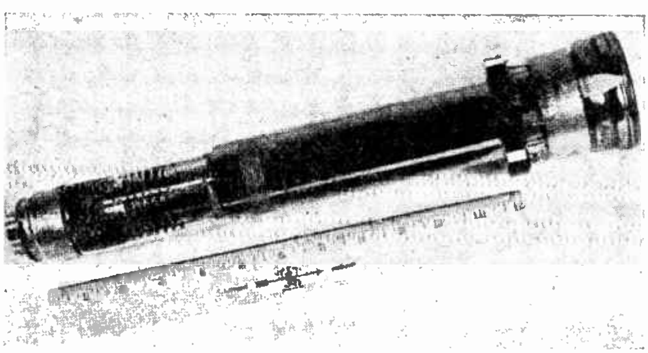

Fig. 3—The image orthicon.

The electron multiplier, two-sided target, and electron-image section will be recognized as elements whose virtues and incorporation into a pickup tube have been discussed frequently in the literature.[1,3-6] The image orthicon represents one way of including all three elements in a useful, sensitive, and stable pickup tube.

III. Typical Operating Cycle

The scene to be transmitted is focused on the semi-transparent photocathode (Figure 2). Photoelectrons are released in direct proportion to the brightnesses of the various parts of the scene. The photoelectrons are accelerated from the photocathode toward the target by a uniform electric field and are focused on the target by a uniform magnetic field parallel to the axis of the tube. The paths of the electrons from photocathode to target are, except for emission velocities, substantially straight lines parallel to the axis. The electron image, accordingly, has unity magnification.

The photoelectrons strike the target at about 300 volts, at which potential the secondary-emission ratio is greater than unity. Because more secondary electrons are emitted than there are incident photoelectrons, a positive charge pattern is formed on the target, the high lights corresponding to the more positive areas. The secondary electrons are collected by the fine-mesh target screen.

At the same time that a charge pattern is being formed on one side of the target, a beam of electrons scans the opposite side. The scanning beam is of the low-velocity type already described for the orthicon.[2] It starts at the thermionic cathode of the electron gun at zero potential and is accelerated by the gun to about 100 volts. From the gun to the target the beam is in an approximately uniform magnetic focusing field. As the beam electrons approach the target they are decelerated again to zero volts. If there is no positive charge on the target, all the electrons are reflected and start to return toward the gun along their initial paths. If there is a positive charge pattern on the target, the beam electrons are deposited in sufficient numbers to neutralize the positive charges. The remaining electrons are re-

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[3] H. A. Iams and A. Rose, "Television Pickup Tubes with Cathode-Ray Beam Scanning," *Proc. I.R.E.*, vol. 25, pp. 1048-1070; August, 1937.

[4] H. A. Iams, G. A. Morton, and V. K. Zworykin, "The Image Iconoscope," *Proc. I.R.E.*, vol. 27, pp. 541-547; September, 1939.

[5] A. Rose, "The Relative Sensitivities of Television Pickup Tubes, Photographic Film, and the Human Eye," *Proc. I.R.E.*, vol. 29, pp. 293-300; June, 1942.

[6] P. T. Farnsworth, "Television by Electron Image Scanning," *Jour. Frank. Inst.*, vol. 218, pp. 411-444; October, 1934.

flected. In this way a stream of electrons, amplitude-modulated by the charge pattern, is started on its way toward the gun.

The return beam not only starts back toward the gun, but it actually arrives at the gun very near the defining aperture through which it emerged. An electron beam will follow closely the lines of a magnetic field under the following conditions: (1) that the beam is initially directed along the magnetic lines; (2) that the beam velocity in volts does not greatly exceed the magnetic field strength in gausses; (3) that electric fields transverse to the magnetic field are small or absent; and (4) that the magnetic lines do not bend sharply. These conditions are approximately fulfilled in the image orthicon. The beam is shot into the magnetic field parallel to its lines. The beam velocity in volts and magnetic field strength in gausses are each in the neighborhood of 100. The only prominent electric field is near the target and parallel to the magnetic field. The bends in the magnetic field caused by the transverse fields of the deflecting coils are well tapered.

The return beam accordingly strikes the gun in an area around the defining aperture which is small compared with the defining aperture disk, but large compared with the defining aperture itself. Also, the return beam strikes this surface at about 200 volts and generates a larger number of secondary electrons than there were incident primary electrons. In short, the defining aperture disk is also the first stage of an electron multiplier. Succeeding stages of the multiplier are arranged symmetrically around and back of the first stage. More will be said of the multiplier in a following section. Meantime, the secondary electrons are drawn from the first stage by suitable electric fields into the succeeding stages. The number of stages, as will be explained, need not be large to exhaust the useful gain of the multiplier. In its present form, the image orthicon uses five stages of electron multiplication.

The output current from the final stage of the multiplier is fed into a wide-band television amplifier in the usual manner. Because this output current is already at a high level, the required gain of the amplifier is small compared with that for an iconoscope or orthicon. The high-level output has other advantages. The performance of the tube, for example, is not critically dependent upon the noise characteristics and input-circuit parameters of the preamplifier, as is the case for the iconoscope and orthicon.

The above operating cycle, while somewhat elaborate, is nevertheless easily traceable. On the other hand, the detailed operation of the parts of the tube does include some interesting and less obvious problems. These will be discussed below.

IV. Electron-Image Section

The semitransparent conducting photocathode is a well-known structure for getting photoemission from the side opposite to that from which the light enters. Photosensitivities several times higher than those for insulating mosaic surfaces can be obtained.

The use of a uniform magnetic field to focus the electron image is not only well known but is also one of the simplest methods of electron-image formation. Unity magnification, erect image, and good definition at low anode voltage are its characteristics.

V. Construction of the Two-sided Target

The two-sided target is perhaps one of the oldest and most frequently proposed structures for improving the sensitivity of a television pickup tube. It makes possible the separation of charging and discharging processes so that the sensitizing procedures and electric fields appropriate to each may be incorporated in the tube without mutual interference. The two-sided target must conduct charges *between* its two surfaces but *not along* either surface. It should have a conducting element nearby to act as the common capacitor plate for the separate picture elements.

Most of the attempts to fabricate two-sided targets have centered on a structure which had discrete conducting elements or "plugs" embedded in an insulating medium. These have been satisfactory for testing the properties of a two-sided target but have failed thus far to provide the uniformity necessary for a commercial tube.

The two-sided target used in the image orthicon is exceedingly simple and capable of a high degree of uniformity. It is a thin sheet of low-resistivity glass. The resistivity is chosen low enough so that charges deposited on opposite sides of the glass are neutralized by conduction in a frame time (1/30 second). It is chosen thin enough so that these same charges do not spread laterally in a frame time sufficiently to impair the resolution of the charge pattern. Thicknesses of five to ten wavelengths of light have been found to be satisfactory.

The thin sheet of glass, about 1½ inches in diameter, is mounted flat to within a few thousandths of an inch and spaced about two thousandths of an inch from a similarly flat fine-mesh screen. The mounting techniques to achieve these tolerances have been the subject of a considerable amount of work. The problem is especially accentuated when it is realized that the assembled structure must go through a standard bake-out schedule at about 400 degrees centigrade. Satisfactory assemblies were obtained only after the glass and screen were each mounted under tension on flat metal rings. The metal ring for the glass had to be carefully chosen so that the 400-degree-centigrade bake-out did not cause the glass either to break or to wrinkle on cooling.

The fine-mesh screen mounted near the glass target to collect secondary electrons and to act as the common capacitive member for all of the picture elements has been, itself, a problem of appreciable magnitude. Because the electron image passes through the screen and impresses the shadow of its wires on the picture, the screen had to be of extremely fine mesh and highly uniform. In addition, for efficient operation, it was desirable to have the percentage open area of the screen 50 per cent or greater. The finest commercial screen available during the early development of this tube which had even reasonable uniformity was a 230-mesh per linear inch, woven-wire, stainless-steel screen. It had 47 per cent open area and could be etched to about 60 per cent open area. The 230-mesh screen was, however, readily resolved in the transmitted picture and limited the resolution objectionably.

In contrast to this screen, a technique was developed for making fine-mesh screens with 500 to 1000 meshes per linear inch, an open area of 50 to 75 per cent, and an accuracy of spacing comparable with that of a ruled optical grating. These screens have made possible the transmission of pictures with high definition and substantial freedom from spurious signals.

VI. Operation of the Two-Sided Target

Figure 4 shows the potentials[7] of the two sides of the glass target during a typical charge-discharge cycle. In Figure 4(a) the tube has been in the dark. The scanned side of the target has been brought to zero volts by the scanning beam. The picture side also is at zero volts as a result of leakage to the scanned side. The fine-mesh screen for collecting secondary electrons is held at + 1 volt. Figure 4(b) shows the target potentials after exposure to light for a frame time. The picture side of the glass has been charged to + 1 volt by the electron image. The scanned side of the target also has been brought up to + 1 volt by capacitive coupling to the picture side. In Figure 4(c),

[7] For simplicity, the emission velocities of the thermionic and secondary electrons are taken to be zero and the contact potentials of all surfaces are taken to be the same. Including finite emission velocities and contact potential differences would merely shift the values of the potentials shown in Figure 4 without affecting the argument.

the beam has just scanned the target, bringing the scanned side down to zero volts and the picture side down *almost* to zero volts by its capacitive coupling to the scanned side. The "*almost*" results from the fact that there is a positive charge on one side of the glass and a negative charge on the other, constituting a charged capacitor. If, therefore, the scanned side is brought to zero volts, the picture side must be positive by an amount equal to the picture charge divided by the capacitance between the two sides of the glass. This turns out to be small compared with the + 1 volt to which the target as a whole has been charged. In particular, it is shown to be 0.01 volts in the illustration chosen. During the next frame time the charges on the

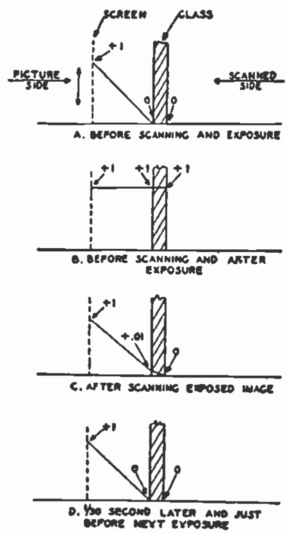

Fig. 4—Target potentials during a typical scanning cycle.

two sides of the glass unite by conduction to wipe out the potential difference between the two sides. Figure 4(d) shows the potentials at this time, and by comparison with Figure 4(a) the target has returned to its initial state ready for another cycle.

In the above cycle, the charging by the picture, discharging by the beam, and leakage between the two sides of the glass were described as events in series. Actually, of course, all three events occur simultaneously and steadily.

It may be remarked, in passing, that the choice of a glass with too high a resistivity (that is, a leakage time constant greater than a frame time) tends to allow charge to accumulate on the picture side.

For sufficiently high resistivities, an objectionable loss of signal, as well as spurious after-images, are encountered.

VII. An Electron-Optical Problem

It has been found that, for good operation over a large range of scene brightnesses, the fine-mesh screen potential should be kept low, about + 1 volt. This means that the glass target potential can swing only between the narrow limits of zero volts, to which the scanning beam charges it, and + 1 volt, to which the picture can charge it as limited by the potential of the fine-mesh screen. The maximum signal output is proportional to the maximum potential swing of the target (e.g., + 1 volt as above). It is important, therefore, in order to insure uniform signal output at all points on the target, to have the limits constant over the target. The upper limit, + 1 volt, as set by the fine-mesh screen, is obviously the same at all points on the target. The lower limit, however, is set by the lowest potential to which the beam can charge the target. If the beam approached the target at all points with normal incidence, the lower limit would be constant over the target and equal to zero volts.[8] The attainment of this "if" is not, in general, a simple task. The ease with which the beam can depart from normal incidence is, perhaps, more suggestive. A few possibilities will be mentioned.

When the beam is shot into the magnetic field by the short electron gun, it is usually not quite parallel with the magnetic lines. The component of the beam's velocity transverse to the magnetic field lines goes into helical motion of the beam. The energy of this helical motion is subtracted from the energy of the beam directed along the magnetic lines. The latter energy, however, determines the potential to which the beam can charge the target. Thus if ½ volt of energy is absorbed in helical motion, the beam can charge the target to only + ½ volt instead of to zero volts. This permits the target to swing only between the limits of + ½ volt and + 1 volt. In other words, the maximum signal output is reduced by half.

Another contribution to the helical motion of the beam may come from the deflection fields. The electron beam, in the process of negotiating a bend in the magnetic field lines, redistributes some of its energy into helical motion.[9] The amount of this energy increases in general for larger angles of deflection, weaker magnetic fields, and

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[8] Again for simplicity, the thermionic-emission energies of the beam electrons are taken to be zero.

[9] A. Rose, "Electron Optics of Cylindrical Electric and Magnetic Fields," *Proc. I.R.E.*, vol. 28, pp. 30-39; January, 1940.

higher beam voltages. Here one expects, and finds, the helical energy, and correspondingly the loss of signal, increasing from the center of the picture out to the edges.

Helical motion introduced into the beam is fortunately a removable defect. One has only to introduce a second source of helical motion of equal amplitude and opposite phase. To correct for helical motion resulting from misalignment of gun and magnetic field, an adjustable, small (in magnitude and physical extent) transverse magnetic field is introduced at the exit end of the gun. To correct for helical motion resulting from the deflection fields, a second source, whose contribution also increases from the center of the picture to the edges, is introduced near the target. This source is the component of the electric field of the decelerating ring transverse to the axis of the tube. The relative phases of the helical motions resulting from the deflection coil and decelerating ring can be adjusted for cancellation by sliding the coil along the axis of the tube. In practice, once a design of the tube and coil has been decided upon, this can be fixed.

What is of particular interest in this problem is the delicacy of adjustment necessary for good performance. A 100-volt beam must be generated, deflected, and corrected in such manner that it approaches all points on the target with not more than a tenth of a volt energy "squandered" in helical motion.

VIII. Electron Multiplier

In spite of the variety of electron multipliers offered by the literature, it was thought desirable to add still another to the list — one which was more nearly suited to the requirements of the image orthicon. A brief consideration of the diffuse spray of secondary electrons emerging from the first multiplier stage (defining-aperture disk) suggests immediately the difficulties of getting all of them to enter the relatively narrow mouth of the more conventional electron multipliers. This is particularly true because it was desirable, for other reasons, to retain the axial symmetry of the electric field in front of the first stage. To focus the secondary electrons into a narrow-mouth multiplier might very well require objectionably strong assymetric electric fields. Once committed to the symmetry of fields, one is also committed to a relatively large entrance opening for the second stage of the multiplier because the secondary electrons spray out symmetrically or "fountain-wise" from the first stage.

It was found to be relatively easy to arrange for substantially all of the secondary electrons from the first stage to strike the large annular-disk second stage shown in Figure 2. The arrangement consisted of surrounding the first stage with electrodes all at lower potential than the first stage, with the one exception of the second stage. In this way the electrons were offered two alternatives: to return to their place of origin, the first stage, or to land on the second stage.[10] Energetically the electrons could return to the first stage, since they were emitted from it with a few volts of spare energy. But to return to the first stage, the electrons must approach it at nearly normal incidence or, more accurately, with all but their emission energy directed normal to the surface. The brief excursion of the electrons into the strong dispersing field provided by the more positive second stage makes the probability of such return small. The secondary electrons from the first stage accordingly quickly find their way to the second stage.

Here the problem is to multiply the electrons again and send them on to a third stage, and so on through a number of stages to the final collector. The use of a series of parallel-screen multipliers is well suited geometrically to the problem, but the efficiency of the screen-type multiplier is low. That is, for a secondary-emission ratio of four, the gain per stage is only about two. The "pinwheel" type of multiplier shown schematically in Figure 2, on the other hand, has an efficiency of 80 to 90 per cent. By inspection it is evident that the electrons incident on a "pinwheel" see an almost opaque surface. There are no holes, as there are in the screen-type multiplier, through which electrons are lost. The secondary electrons, however, readily pass through the blades toward the succeeding stage. They are helped in their path by the coarse-mesh guard screen which shields them from the suppressing action of the negative potential of the preceding stage. Succeeding stages have their blades opposed to accentuate their opacity. The operation of the multiplier was found to be uncritical to electrical adjustment and mechanical alignment. Both these features are highly desirable to simplify the construction and operation of an otherwise complex tube.

Total gains of 200 to 500 are readily obtained for the five-stage multiplier. These gains are usually more than sufficient to exhaust the sensitivity possibilities of electron multiplication. The "useful" gain obtainable with electron multiplication is discussed in the following section.

IX. Sensitivity and Signal-to-Noise Ratio

It was pointed out in the introduction that the image orthicon

[10] The third possibility, that of retaining their freedom in space, is usually of negligibly short duration.

derives its increased sensitivity over the iconoscope and orthicon from (1) the higher photosensitivity of a conducting photocathode relative to that of an insulating mosaic; (2) the multiplication by secondary emission of the electron image at the target; and (3) the use of an electron multiplier for the signal current. The gain from (1) and (2) is about a factor of five. It must be remembered that this factor reflects more the state of the art of making photosensitive surfaces than any intrinsic limitations. The gain from (3) is a function of the signal-to-noise ratio in the transmitted picture. The term "noise" as used here refers to the more or less fundamental current fluctuations associated with amplifiers or generated in the pickup tube. These fluctuations give rise to a masking effect, often referred to as "snow", in the transmitted picture. The video signal current must exceed the noise current before a picture can be seen. The noise currents, therefore, set the threshold scene brightness that a pickup tube can transmit; they also define the scene brightness required for the transmission of good pictures, that is, pictures with high signal-to-noise ratios.

The performance of the iconoscope and orthicon is limited by the noise currents in the first tube of the television preamplifier. The performance of the image orthicon is limited by the much smaller noise in the scanning beam. The multiplier, accordingly, provides a useful gain in sensitivity up to the point at which the shot noise in the scanning beam is made equal to, or slightly greater than, the noise current in the preamplifier. The usual preamplifier noise current[11] is 2×10^{-9} ampere for a 5-megacycle bandwidth. The shot noise in the scanning beam is $(2eI\Delta f)^{1/2} = I^{1/2} \times 10^{-6}$ ampere for the same bandwidth, where I is the scanning-beam current in amperes. The "useful" multiplier gain is, therefore,

$$\frac{2 \times 10^{-9}}{I^{1/2} \times 10^{-6}} = \frac{2 \times 10^{-3}}{I^{1/2}} .$$

A more convenient way of expressing this gain is to make use of the relation between the scanning-beam current and the maximum signal-to-noise ratio that can be obtained when the beam is fully modulated. Under these conditions, the maximum signal is the beam current itself; the noise associated with this signal is the shot noise in the beam; and the signal-to-noise ratio R is given by

[11] H. B. DeVore and H. A. Iams, "Some Factors Affecting the Choice of Lenses for Television Cameras," *Proc. I.R.E.*, vol. 28, pp. 369-374; August, 1940.

$$R = \frac{I}{I^{1/2} \times 10^{-6}} = I^{1/2} \times 10^{6}.$$

With this relation, the useful gain of the multiplier may be written as $2000/R$. Some comments and caution are needed in the application of this gain expression.

The useful gain was computed for 100 per cent modulation of the scanning beam. In practice, for medium- and high-light pictures, modulations in the neighborhood of 50 per cent are realized. The lowered modulation results, for the most part, from the fact that all of the electrons that strike the target do not stick — some are reflected or scattered back. Further, for low-light pictures, near threshold, the modulation is still lower because the potential swing of the target is smaller than the emission velocities of the electrons in the scanning beam — only the higher-velocity electrons can land. Whatever the source of lower modulation, the useful gain is reduced in proportion to the modulation.

With the above limitations, the useful gain of the multiplier is of the order of 20 for a high-light picture and of the order of 200 for a low-light picture. The combined gain of the electron-image section and the multiplier make the image orthicon from 100 to 1000 times as sensitive as the iconoscope or orthicon.

The sensitivity of the image orthicon is high enough to make comparisons with the performance of the eye both significant and interesting. The image orthicon has approximately the same intrinsic sensitivity[5] as the eye. This means that, for scene brightnesses *near the threshold for the tube*, both tube and eye can transmit the same pictures. On the other hand, the greater flexibility of the eye relative to a television system enables it still to "see" scenes whose brightness is as little as one thousandth of the threshold scene brightness for the pickup tube. The eye attains this low threshold by sacrificing resolution for operating sensitivity.

X. Signal Versus Light Characteristics

A representative curve for the video signal as a function of light is shown in Figure 5. Three equivalent abscissa scales are shown for convenience in referring to scene brightness, image brightness, or photocathode current. Also, the video signal is given in microamperes of modulated signal at the target. It is this current which determines the signal-to-noise ratio. The final output signal is the product of the video signal at the target and the gain of the electron multiplier, usually several hundred. The multiplier is an almost noiseless device.

The curve is divided, for purposes of discussion, into four parts by the letters *A*, *B*, *C*, *D*, and *E*. These will be considered in order, starting from the left.

The low-light range *A-B* is particularly simple. Here the signal out is proportional to the light in, just as it is for the orthicon. At the lowest point on the curve, the video signal is equal to the shot noise in the scanning beam. The beam current is adjusted in this range just to discharge the picture. As point *B* is approached, higher signals and signal-to-noise ratios are obtained. At *B*, the light is just sufficient to cause the target to be fully charged (i.e., to the potential of

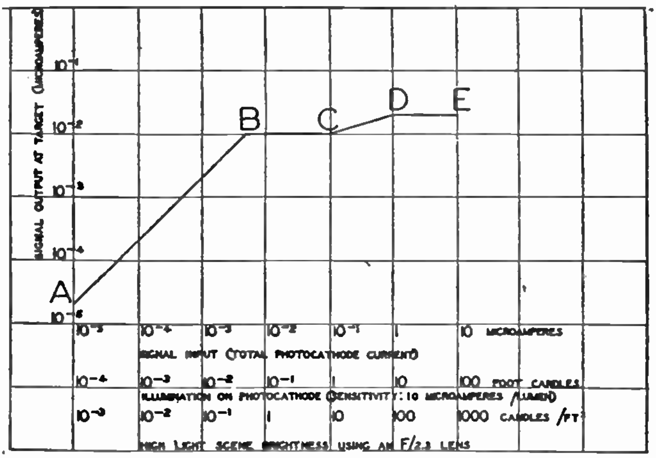
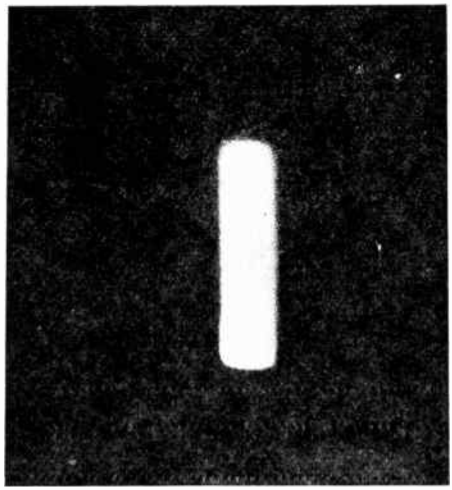


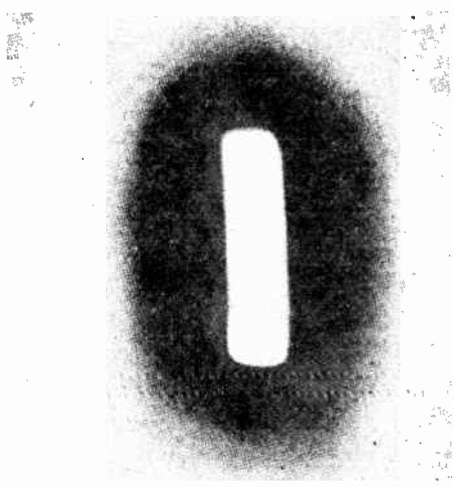
Fig. 5—Signal versus light characteristic

the fine-mesh screen) in a frame time of 1/30 of a second. One would ordinarily expect that increasing the light level beyond *B* would tend to saturate the transmitted picture. The high lights would remain constant in amplitude in this range; the low lights would continue to increase and tend to make the entire picture white. This is what one ordinarily would interpret from Figure 3. Actually, pictures transmitted by the image orthicon in the range *B-C* have, except for large black areas, the same or improved contrast. The explanation follows.

Figure 6(a) shows the transmitted picture of a single spot of light whose brightness is located at *B*. The picture is normal. Figure 6(b) shows the transmitted picture of the same spot illuminated to ten times the previous brightness. One sees in this figure that the signal output did not change for a tenfold increase in original picture



(a)



(b)

Fig. 6—Transmitter picture of light spot at low and at high spot brightness.

brightness, that the contrast of the spot is maintained in the immediate neighborhood of its boundaries, and that the rest of the background, supposedly black in the original, has begun to lighten up. The black halo surrounding the light spot in Figure 6(b) is the key to the preservation of good picture contrast in the *B-C* range. This halo is formed by low-velocity secondary electrons originating in the light spot and scattered into the immediate neighborhood of the light spot. Where they land, they tend to keep the target charged negatively and to counteract the effect of stray light, tending to wash out the picture. In brief, the brighter areas in the *B-C* range tend to maintain their potential higher than neighboring less-bright areas by spraying the less-bright areas with more low-velocity secondary electrons than they get in return. While the "halo" effect is unnatural in Figure 6(b), it is not visible, as such, in the usual fine-detail half-tone picture (see Figure 8), and serves only to maintain picture contrast.

The "halo" has another useful function. If the spot of light in Figure 6(b) is moved rapidly across the field of view, the transmitted picture is not a continuous white streak as one would expect from an orthicon or from an image orthicon in the low-light range *A-B*. The transmitted picture is a series of relatively sharp tilted images of the spot separated by 1/30-second intervals. In effect, the sharp tilted image is not unlike what one obtains from a focal-plane shutter in a photographic camera. The mechanism for generating the effect is the discharging action of the halo electrons. When the spot of light is displaced from an initial position, the halo electrons erase, by discharging, the initial charge pattern. The brighter the light, the more rapid the erasing action and the more sharply resolved are pictures in motion.

The second rise in video signal, namely, the range *C-D*, has an interesting origin. An outline of the argument for its existence will be given here. The signals in both the ranges *B-C* and *C-D* are determined by the charge accumulated on a picture element just prior to being scanned by the electron beam. In the range *B-C*, this charge is equal to the total charge that the *entire target*, considered as a parallel-plate capacitor, can accumulate divided by the number of picture elements. In the range *C-D*, the picture-element charge is the total charge that an *element* can accumulate as determined by the capacitance of that element, alone, to the signal plate. If the spacing between target glass and fine-mesh screen is small compared with the diameter of a picture element, these two charges are equal and there is no "second rise" in the *C-D* range. As the spacing between glass and screen is increased, the capacitance of the target as a whole decreases linearly with the reciprocal spacing, while the capacitance of a picture element alone levels off to a constant value, independent of spacing and equal to the capacitance of a disk, the size of a picture element, in free space. The usual spacing is such that the capacitance of a picture element alone is two or three times the capacitance that would be computed for the picture element by dividing the number of picture elements into the total target capacitance.

Thus far, a basis has been established for the separate picture elements having more capacitance and being able to store more charge than is possible when these picture elements act together as a complete target. It turns out, however, that the additional storage capacity does not become effective until the light is sufficiently intense to charge the target as a whole in a small fraction of a frame time. Hence, the flat plateau *B-C* before the "second rise" *C-D* sets in. The end of the second rise, point *D*, should and does occur when the light is sufficiently intense to charge the target as a whole in a line time.

Beyond *D* the signal output curve again levels off and the transmitted picture does not change with changes in scene brightness.

To summarize: in the low light range, the image orthicon acts like an orthicon; in the high light range, the transmitted picture is substantially independent of scene brightness, the contrast and half-tone scale being maintained by redistributed secondary electrons on the picture side of the target. These redistributed electrons have also the property of tending to keep moving images in sharp focus.

XI. Resolution

Starting at one end of the tube with a well-focused image on the photocathode, the picture undergoes three transformations before emerging from the multiplier at the other end in the form of a modulated signal current. The transformations are, in order: optical image to electron image, electron image to charge pattern on the target, charge pattern to modulated stream of electrons in the scanning beam. Each transformation has been capable separately of resolving over 1000 lines per inch; the combination has resolved well over 500 lines per inch.

The resolution of the electron image is limited by the emission velocities of the photoelectrons. The resolution of the charge pattern on the target is limited, at high lights, in part by the fine-mesh screen, and at low lights, in part by the leakage along the glass target. The ability of the scanning beam to resolve the charge pattern is controlled by a number of factors, among which are defining aperture diameter, thermionic-emission velocities, angle of approach to the target, and magnitude of the potential differences in the charge pattern. The magnetic field strength, once adjusted for focus, has no first-order effect on the resolution of either the scanning beam or the electron image.

On the other hand, the resolution of both the scanning beam and the electron image improves with increasing electric field strength on the scanned side of the target and in front of the photocathode, respectively.

An expression has been derived[12] for the limiting current density that may be focused by an electron gun into a spot on a target. This current density is proportional to the target potential and to the $\sin^2$ of the angle of convergence of the electrons approaching the target. Experience with oscilloscopes and kinescopes has led to high anode potentials, kilovolts and tens of kilovolts, for the purpose of getting small spots. It may, accordingly, appear surprising to find even smaller spot sizes attained in the image orthicon at a target potential of approximately zero volts. The smaller beam-current densities used in the pickup tube are only part of the explanation. The larger part is the difference in the convergence angles of the electrons approaching the pickup tube target and kinescope screen. For the orthicon type of pickup tube the $\sin^2$ of this angle is near unity, while for the kinescope it is usually 10^{-3} to 10^{-4}. Thus the low-velocity scanning beam makes up for its low velocity by its large convergence angle.

XII. Performance

Representative pictures transmitted by the image orthicon are shown in Figures 7, 8, and 10. Figures 7 and 8 are the transmitted pictures of slides projected on the photo cathode. Figure 10 shows the results of a test in which a direct comparison was made between the operating sensitivity of an image orthicon and of a 35-millimeter camera using Super-XX film. The experimental setup for the comparison is shown in Figure 9. The original subject was illuminated with an ordinary 40-watt bulb attenuated with neutral filters. The television camera was focused on the subject alone and its picture was reproduced on a receiver located alongside the subject. The 35-millimeter camera photographed simultaneously the original and reproduced pictures. Both cameras used *f*/2 lenses and an exposure time of 1/30 second. It will be seen from Figure 10 that only in the first exposure, at 2-foot-lamberts brightness of the subject, do both original and reproduced pictures appear. At 0.2 foot-lambert only the picture reproduced by the television camera is present. And, in fact, the television camera continues to transmit a picture even at 0.02 foot-lambert, which is the brightness of a white surface in full moonlight.

[12] D. B. Langmuir, "Theoretical Limitations of Cathode-Ray Tubes," *Proc. I.R.E.*, vol. 25, pp. 977-991; August, 1937.

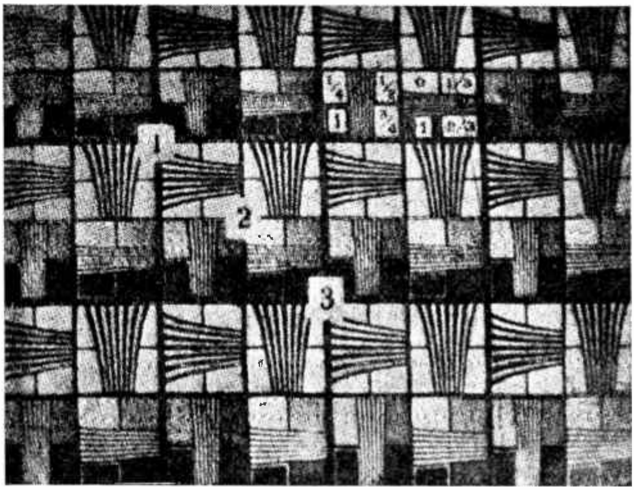


Fig. 7—Test pattern transmitted by image orthicon.

ACKNOWLEDGMENT

The work on the image orthicon has had an extended course. Throughout, it has profited from the experience and helpful criticism of many of the writers' associates both in these Laboratories and in other divisions of The Radio Corporation of America. Much of the work was made possible by an immediate background of pickup-tube research, largely as yet unpublished, and contributed by a number of

Fig. 8—Half tone transmitted by image orthicon.

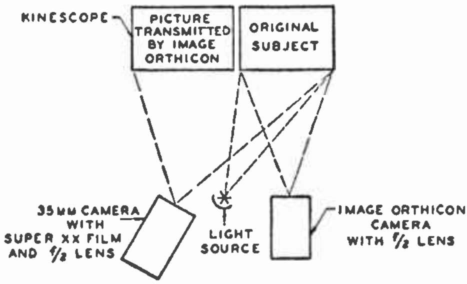


Fig. 9—Setup for comparing the sensitivities of image orthicon and photographic film.

individuals. Among these are H. B. DeVore,[13] L. E. Flory,[14] R. B. Janes,[15] H. A. Iams,[14] G. L. Krieger,[11] G. A. Morton,[14] P. A. Richards,[15] J. E. Ruedy,[14] and O. H. Schade.[16] The writers would particularly like to acknowledge the encouraging direction of B. J. Thompson (now deceased) and V. K. Zworykin, and the valuable contributions of S. V. Forgue,[14] J. Gallup,[16] and R. R. Goodrich.[14]

The groundwork for the image orthicon had already been laid prior to the war. Early in the war, effort was directed under an Office of Scientific Research and Development contract toward developing the image orthicon in a form suitable for military purposes.

(a) 2-foot lamberts (b) 0.2 foot-lambert (c) 0.07 foot-lambert (d) 0.02 foot-lambert

Fig. 10—Comparison of sensitivities of image orthicon and 35-millimeter Super XX film. (Incandescent light source.)

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[13] Formerly, RCA Laboratories.
[14] RCA Laboratories Division.
[15] RCA Victor Division, Lancaster, Pa.
[16] RCA Victor Division, Harrison, N. J.

A METHOD OF MEASURING THE DEGREE OF MODULATION OF A TELEVISION SIGNAL*†

By

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Summary—A method of measuring the degree of modulation on a standard television signal is described. The double sideband output of the transmitter energizes a linear diode monitor, the output of which contains a direct current component in addition to the visual signal. Means are provided to interrupt this composite signal periodically by short-circuiting the diode output load impedance for a brief interval, thus establishing a reference zero signal. The resultant modified signal, including the zero reference level, may be observed by means of a cathode ray oscilloscope capable of handling only alternating current signals. The trace on the face of the oscilloscope will contain all of the information required to measure the degree of modulation attained.

Introduction

THE need for determining the degree of modulation which was attained on the signal radiated by a television transmitter was apparent very soon after experiments were begun with television transmission. Most of the modulation monitoring methods which were developed for sound broadcasting were not applicable to television broadcasting. The method of measuring the degree of modulation by observing the carrier frequency envelope on a cathode-ray oscilloscope was applicable to television provided that the information given by the trace was properly interpreted. The current television standards require that the carrier envelope achieve maximum amplitude at the peak of the synchronizing signal and that this maximum amplitude shall be independent of light and shade in the picture signal. As a consequence of this method of operation, the peak carrier envelope amplitude becomes a constant, whereas the average carrier envelope amplitude becomes a variable dependent upon the content of the picture signal. Therefore, modulation measurements under existing standards for television transmission must be made in terms of the peak carrier envelope amplitude, in contrast to sound broadcasting practice, wherein such measurements would be referred to the constant which in that case would be average carrier envelope amplitude.

When the radio frequency envelope of the visual transmitter was

* Decimal Classification: R254.1 × R583.

† Reprinted from *RCA REVIEW*, June, 1946.

monitored on a cathode-ray oscilloscope, the operators were in a position to assert with confidence that the signals being radiated were in accordance with the current standards. This method was reasonably satisfactory, but the location of the cathode-ray oscilloscope was determined by the probable accuracy of results rather than by operating convenience. The cathode-ray oscilloscope, a relatively expensive piece of equipment, was made unavailable for other purposes when frequent monitoring of the radio frequency envelope was considered necessary. A more expedient method of obtaining the information offered by the cathode-ray oscilloscope envelope monitoring method had been sought for some time.

An article by A. W. Russell[1] suggested the use of a vibrating switch to "preserve the direct current level in oscillograph amplifiers." While the usefulness of this method in studying the operating characteristics of many vacuum tube circuits was immediately evident, its application to the measurement of modulation was not conceived until several months had elapsed. During the course of the experimenting which followed, the switching mechanism which was used became identified as the "Vibroswitch."

A diode rectifier, which derived its signal from the coaxial radio frequency transmission line between the transmitter and the vestigial sideband filter, has been used for many years as a radio monitor. The quality of the picture was observed on a kinescope while the wave form and amplitude of the composite signal were observed on a cathode ray oscilloscope as a regular operating procedure. The "Vibroswitch" was applied to the diode monitoring system.

Theory

The circuit diagram of the diode rectifier, "Vibroswitch," and cathode ray oscilloscope arrangement is shown in Figure 1. When the circuit constants have been properly chosen, the instantaneous potential difference developed across the diode load impedance Z_c is substantially proportional to the instantaneous carrier envelope amplitude. In a constant peak carrier amplitude system of modulation (direct current transmission), which is currently standard for television, the peak carrier amplitude is attained during the synchronizing pulse interval. The minimum carrier amplitude occurs when a maximum white signal is present. If the modulation were complete during a given maximum white interval, the concurrent instantaneous carrier envelope amplitude would be zero, and as a result the concurrent instantaneous potential

[1] A. W. Russell, "Preserving the D. C. Level in Oscillograph Amplifiers". *Electronic Eng.*, Vol. XV, No. 175, page 173, Sept., 1942.

difference across Z_c would also be zero. It, therefore, appears that if we periodically short-circuit Z_c, we will artificially create the conditions which would obtain during complete modulation. If the rate at which the short-circuiting occurs is sufficiently rapid, the resultant revised signal will be passed by the cathode-ray oscilloscope amplifiers, and the amplitude of the resultant trace should be proportional to the instantaneous potential drop across Z_c and, therefore, within certain limitations, proportional to the instantaneous carrier envelope amplitude. One limitation is imposed by the degree of linearity possible between the voltage applied to the diode circuit and the resultant current. Another limitation is imposed by the effective diode circuit time constant. These circuits must be so designed as to permit the rate of

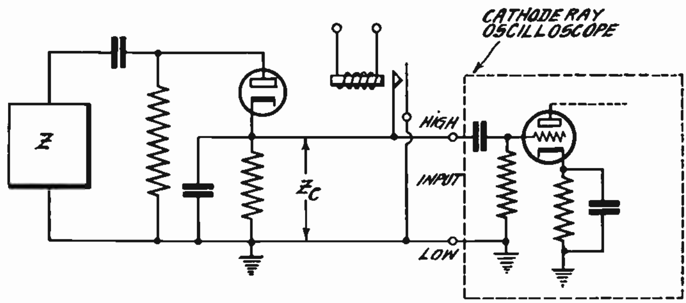


Fig. 1—Diode rectifier, "Vibroswitch," and cathode ray oscilloscope circuit arrangement.

change of potential difference across Z_c to follow the rate of change of carrier envelope amplitude required to transmit the desired intelligence. Further, the information being transmitted during the short-circuiting interval cannot be recorded by the cathode-ray oscilloscope. The interpretation of the results must be made in the light of these limitations.

THE "VIBROSWITCH"

The original "Vibroswitch" was a standard vibrator such as is used in automobile receiver power supply units, but revised for 60 cycle alternating current operation. However, the contact spring tension varied with use to a degree that rendered this instrument too unreliable for regular use under operating conditions. Experimentation then proceeded through the use of a motor driven segmented disc, a motor driven cam, a loudspeaker element equipped with contacts and, more recently, a specially constructed switch using the coil and magnet from a Baldwin headset. The mechanical schematic diagram of this unit is shown in Figure 2. The physical appearance is evident in

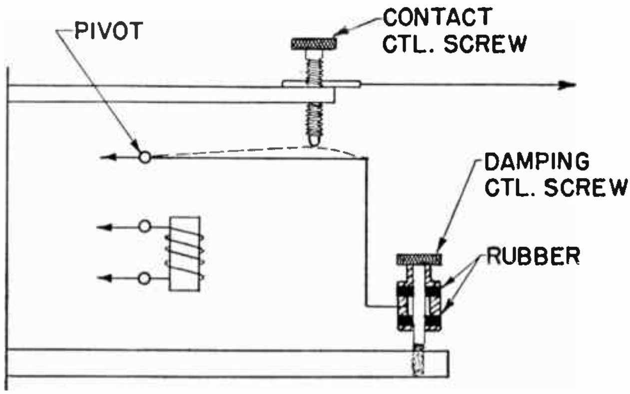


Fig. 2—Mechanical schematic diagram of the "Vibroswitch."

Figure 3. The fundamental problem insofar as the "Vibroswitch" is concerned is to obtain a short closed contact period with clean make

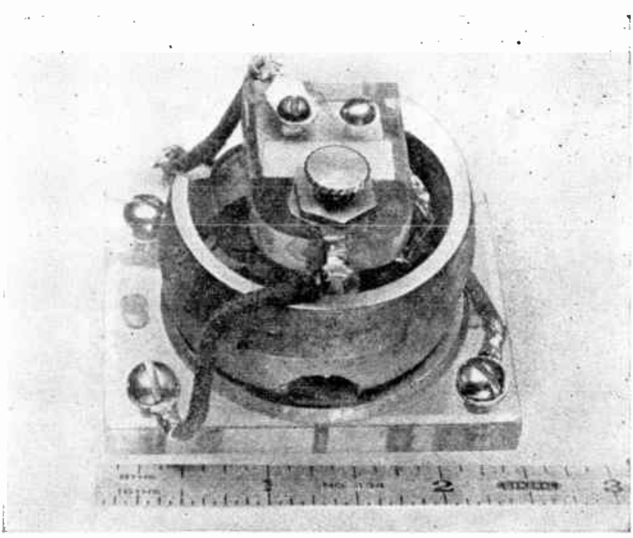
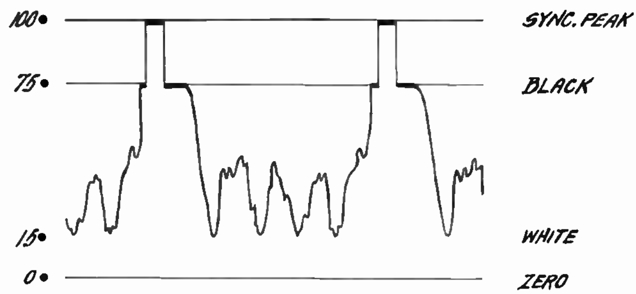


Fig. 3—A recent physical form of the "Vibroswitch."

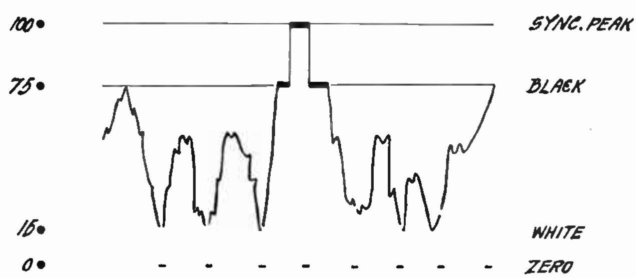
and break. Most of the earlier models suffered from mechanical oscillation of the swinger, causing variation in contact resistance at the instant that the contact was closed. This led to a confused trace on the oscilloscope.

INTERPRETATION OF THE OSCILLOGRAMS

Figure 4 gives the expected oscilloscope traces. The actual appearance of the trace on an oscilloscope is shown in the photographs



(a) Horizontal deflection rate approximately one half the field repetition rate.



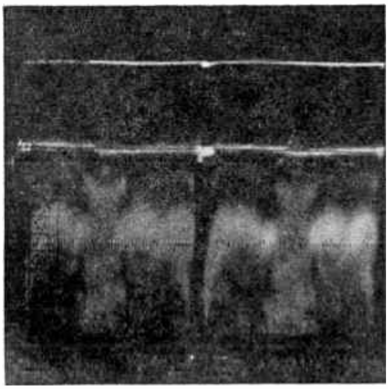
(b) Horizontal deflection rate approximately one half the line repetition rate.

Fig. 4—Representation of the expected Oscilloscope Trace:

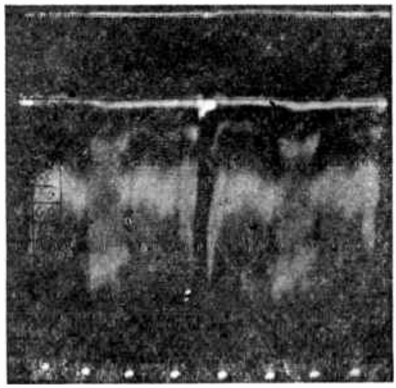
included in Figure 5. If the vertical deflection circuit of the monitoring oscilloscope operates with the direct current component of the signal re-inserted, it is possible to set up a scale reading 0 to 100 on the face of the oscilloscope and using the zero carrier level indication provided by the short-circuiting interval of the "Vibroswitch" cycle, set the gain of the oscilloscope amplifier so that the peak of sync falls at 100 and the zero carrier dot or line falls at zero The amplitude of the white signal and black level can then be read directly in per cent of peak carrier envelope amplitude. Similarly, variation of black level or peak carrier as a function of average brightness can be observed and read in per cent of peak carrier envelope amplitude.

General Comments

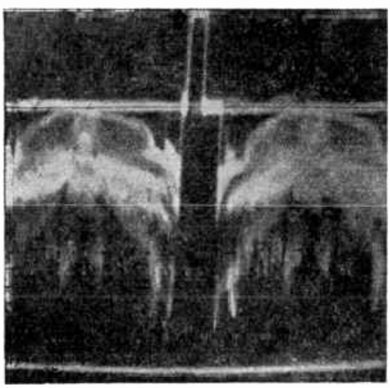
The optimum repetition rate of switching would probably vary with



(a) Horizontal deflection rate one half the field repetition rate—"Vibroswitch" not operating.



(b) Same horizontal deflection rate—"Vibroswitch" operating.



(c) Horizontal deflection rate one half the line repetition rate—"Vibroswitch" operating.

Fig. 5—Photographs of oscilloscope traces:

each application. Experience with monitoring standard television transmissions indicates that a repetition rate in the order of 800 to 1000 cycles per second is acceptable. The switching rate should be nearly, but not exactly, in synchronism with the signal being observed.

The mark or short-circuiting interval should be short, perhaps on the order of 10 per cent, but long enough so that there can be no doubt that the circuit has been fully discharged and that a positive mark is evident at the zero carrier level. The cathode ray oscilloscope amplifiers must be linear over a sufficient swing to pass the composite signal without compression.

Measurements of black level in per cent of peak carrier envelope amplitude, white signal in per cent of peak carrier envelope amplitude, and variation of black level as a function of average brightness using the "Vibroswitch" technique have been checked against the envelope cathode ray oscilloscope method. The results of the two methods were found to be in substantial agreement.

This device permits measurements on low power equipment which would not provide sufficient voltage to deflect the plates of an envelope cathode ray oscilloscope directly.

Acknowledgment

The actual device described herein is the result of the work of many engineers who have been associated with the author. Their contributions have been directly responsible for the processing of an "idea" into a practical and useful tool.

INTERLOCKED SCANNING FOR NETWORK TELEVISION*†

By

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Summary—The benefits of operating the scanning systems of two or more independent television broadcasting plants in locked coincidence are discussed. The problem of producing locked coincidence is explored, and methods of achieving the desired result are indicated. Some of the possible benefits of using stable (high inertia) frequency sources for scanning systems are noted.

BENEFITS OF INTERLOCKED SCANNING

THE desirability of maintaining continuity of synchronizing information on the television broadcast signal was recognized by the designers of early television plants, and the plants were so arranged that no interruption of the synchronizing signal occurred while switching between cameras or between local studios. However, no means for maintaining this continuity of synchronizing signal has evolved as yet for operational use when switching between local studios and remote pickup points or network programs where independent synchronizing signal generators are involved. To render less objectionable the attendant loss of synchronizing signal at receivers when such switches are made, it is usual practice to go to a dark screen before the switch is made and to fade up from a dark screen after the switch is completed. This allows time for synchronizing circuits in the receivers to lock in on the synchronizing signal transmitted after the switch before a picture reappears on the receiver, and hence the resultant disturbance of the image due to the momentary loss of the synchronizing signal is less noticeable. As an additional aid in bridging this gap in synchronizing signal continuity, it is good practice to have the 60-cycle components of the two synchronizing signals involved phased for approximate coincidence.

If complete time coincidence between the two signals (local and remote) could be maintained, it would be possible to preserve continuity of the transmitted synchronizing signal and thereby eliminate the necessity of going to a dark screen during switching, or, of checking and maintaining the vertical phasing of the two signals. From

* Decimal Classification: R583.13.
† Reprinted from *RCA Review*, December, 1947.

the standpoint of smooth program presentation, this is of importance because it is frequently necessary to switch between the local and remote pickups several times in the course of one program, particularly when presenting commercial announcements. In addition, if the local synchronizing signal is transmitted continuously on either the local or remote picture signal (this requires processing the remote signal to remove its synchronizing signal and adding the local synchronizing signal) the receivers are fed a relatively more noise-free synchronizing signal. Also made available between the local and remote signals are lap dissolves, super-impositions and all other processes normally available between local cameras or studios. As network television broadcasting grows, the foregoing aids to smooth presentation of programs assume increasing importance. Therefore, an analysis of what is involved in producing locked coincidence between a local and a remote signal is timely.

Requirements for Producing Interlocked Scanning

Complete coincidence is required between the local and remote signal to achieve the above benefits, and coincidence must therefore be on a line, field, and frame basis. This means that even fields must coincide with even fields, etc., and in the final analysis that each line of the 525 lines in a frame of one signal must coincide with its counterpart in the other signal. This complete matching is required before the local synchronizing signal can be used on the incoming signal and comply with FCC standards of transmission. If even and odd fields are matched and line blanking in the two signals coincide, the local synchronizing signal would align with the incoming signal as shown in Figure 1, where one line frequency synchronizing signal pulse (local) rests on the vertical blanking (remote), or one equalizing (2 times line frequency) pulse (local) rests on the last horizontal blanking (remote), or is lost in the video of the last line of alternate fields. When the required complete coincidence is obtained, the lock applied to maintain the coincidence must be quite rigid. Any hunting permitted by the lock would: first, render impossible the use of local synchronizing signal on the remote picture signal and hence, continuity of synchronizing signal transmission; and second, render ineffective the use of superimpositions and lap dissolves. The only gain in using a loosely-locked coincidence between the local and the remote signal lies in the fact that the vertical components of the two signals will remain approximately in phase and therefore the tendency for vertical scanning at receivers to lose synchronization momentarily following a switch will

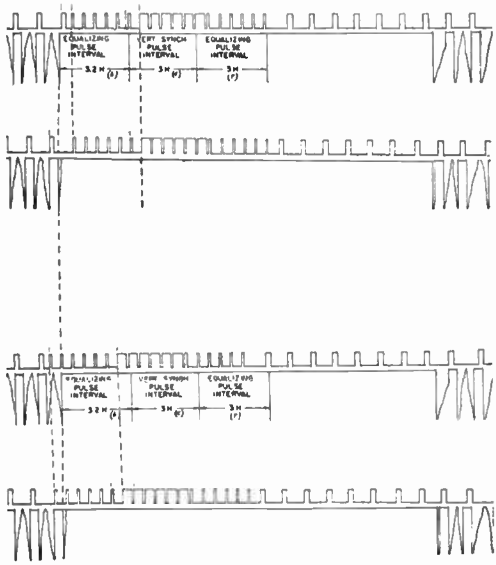

Fig. 1—Local synchronizing signal superimposed on a remote video signal showing the two possible conditions, both mis-matches, which can obtain when even vs. odd fields are locked in coincidence.

be reduced. Figure 2 serves to illustrate the coincidence required to achieve the desired results.

Methods of Achieving Interlocked Scanning

The block diagram of Figure 3 is an arrangement which has been used in "On the air" demonstrations of the operation of two independent television plants under conditions of "locked coincidence." While admittedly an experimental arrangement, it served to confirm the

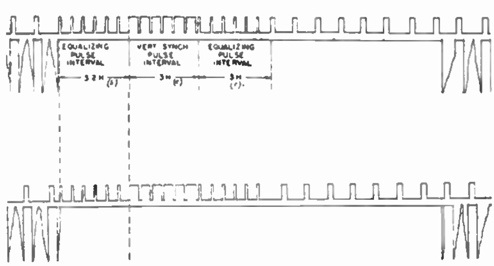

Fig. 2—Local synchronizing signal superimposed on a remote video signal showing the desired result which obtains when even vs. even fields are locked in coincidence.

possibility of realizing the benefits mentioned heretofore, and also to demonstrate effectively the program possibilities during the hours when election returns were coming in for a fall election, 1941. The regularly scheduled program on that occasion was wrestling from a sports arena via the Telemobile Unit. In the Radio City studios a camera was focused on a black-board upon which election returns were recorded. By using only the top of the black-board and by keeping the top of the picture from the Telemobile Unit relatively clear of action (normally the case) a superimposition of the election results upon the incoming sports picture provided the latest election returns without interrupting the sports event.

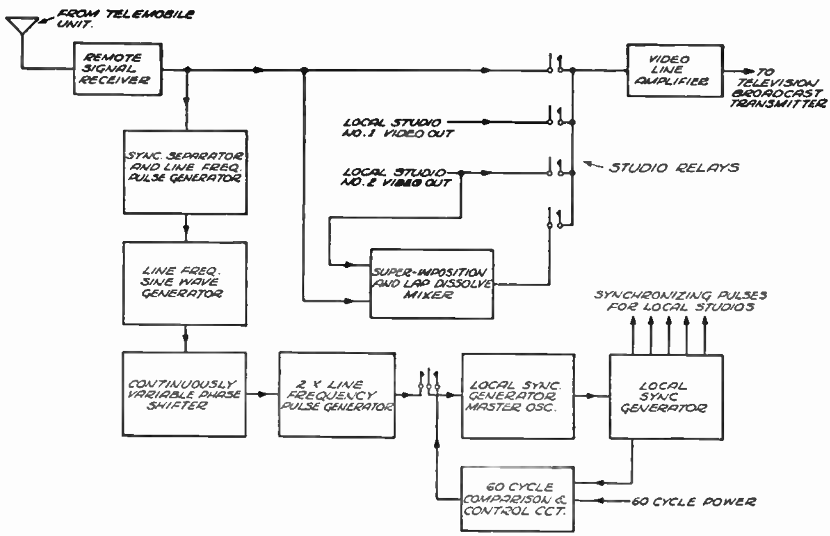

Fig. 3—Block diagram of equipment used to lock the local synchronizing signal generator with the incoming remote video signal.

Figure 4 is a schematic of the block diagram of Figure 3. The synchronizing signal separator and line frequency pulse generator are omitted, the former consisting merely of a conventional synchronizing signal separator, which drives a line frequency blocking oscillator. The line frequency sine wave generator has already been described.[1] The continuously-variable phase shifter made use of a rotating magnetic field and a pickup coil whose physical position could be advanced or retarded without limit in the rotating field. At the time of the demonstration the most convenient equipment for accomplishing this result

[1] R. A. Monfort and F. J. Somers, "Measurement of the Slope and Duration of Television Synchronizing Impulses," *RCA REVIEW*, Vol. VI, No. 3, pp. 370-389, January, 1942

was a small Selsyn unit. The circuit for deriving the three-phase excitation for the Selsyn from the single phase output of the sine wave generator is shown in Figure 4. The resistance-capacitance components on the first grid of two of the legs are chosen to get the desired shift as indicated by the vectors. Following the continuously variable phase shifter the sine wave is processed to provide the locking information to the master oscillator of the local synchronizing generator. As stated before, the lock must be quite rigid before the desired benefits can be realized. In the demonstration referred to previously, the local synchronizing generator was a standard commercial unit.[2] The master

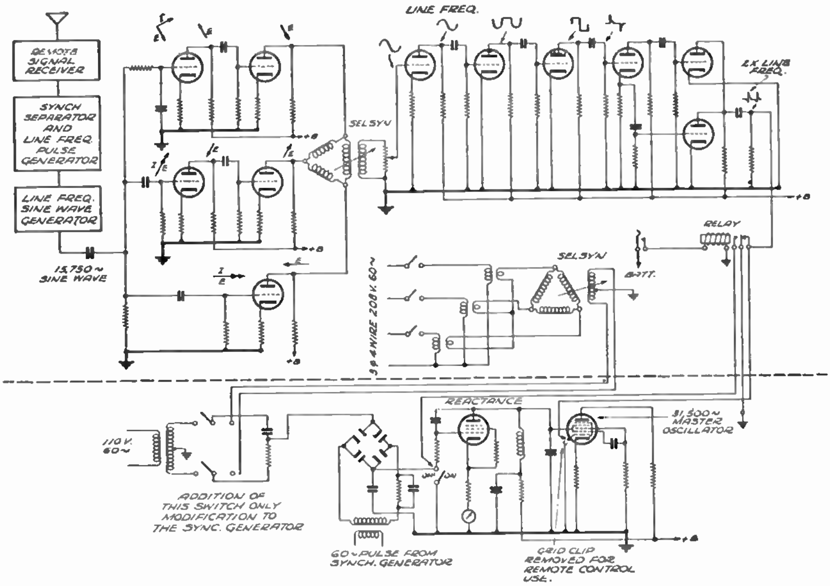

Fig. 4—Schematic diagram of equipment used to lock the local synchronizing ing signal generator against the incoming remote video signal.

oscillator in this unit is of the negative transconductance type and operates at twice line frequency. By trial and observation it was determined that control or lock of the desired degree of rigidity could be obtained by injecting a pulse of twice line frequency into the first grid of this master oscillator. Therefore, for the demonstration the processing required was the conversion of the line frequency sine wave output of the continuously variable phase shifter into twice line frequency pulses. This was accomplished by the symmetrical clipping of the sine wave to produce a symmetrical square wave which was in turn differentiated. A push-pull input into grid current biased grids

[2] A. V. Bedford and J. P. Smith, "A Precision Television Synchronizing Signal Generator", *RCA REVIEW*, Vol. V, No. 1, pp. 51-68, July, 1940.

of two tubes, the plates of which were in parallel, provided the desired result. An alternative and equally effective method would be the use of an unfiltered output from a full wave rectifier driven by the continuously-variable phase shifter.

The method of coupling this double frequency pulse into the master oscillator and of transferring the control of the local master oscillator from the 60-cycle power frequency to the incoming video signal is also indicated in Figure 4. No major modification of the local synchronizing generator was required, three clip leads from a relay clipped on at the proper points and the lifting of the grounded grid cap to the first grid of the master oscillator sufficing. However, it will be noted that the 60-cycle power into the comparison circuit which normally controls the master oscillator of the local generator was modified by the insertion of a continuously variable phase shifter. The same style Selsyn was used for both the 60-cycle and 15750-cycle phase shifters. The reason for this modification to the local generator is apparent when the task of securing coincidence between the two signals by the use of the line-frequency phase shifter alone is considered. A maximum of approximately 262½ revolutions of the line frequency phase shifter may be required if the control should be transferred to the remote signal when coincidence between an even field of one signal existed with an odd field of the other signal. It is much faster to use the 60-cycle phase shifter for rough setting of coincidence, transfer control, and finish the exact alignment of coincidence between the two signals by means of the line-frequency phase shifter.

For checking coincidence, the pulse cross unit, which was described in the paper on the sine wave generator,[1] provides an effective indicator. The local and incoming signals are mixed and applied to the pulse cross monitor and the phase shifters (60 and 15750 cycles) are adjusted to secure coincidence of the two signals on the pulse cross. Figure 5 shows blanking from one signal generator and synchronizing signal from another. There is lack of coincidence in terms of both line and field. Figure 6 shows coincidence for line but not for field. Figure 7 shows odd vs. even field coincidence. Note that one line frequency synchronizing signal pulse is sitting on the field frequency blanking. Figure 8 illustrates the same condition except the line frequency phase shifter has been rotated one revolution from the condition of Figure 7. Here an equalizing pulse is resting on the last line frequency blanking preceding field blanking. Figure 9 shows even field vs. even field coincidence.

Figure 9 is the same as Figures 7 and 8 except one signal is shifted through 262½ lines. Any hunting between signals is readily discerned

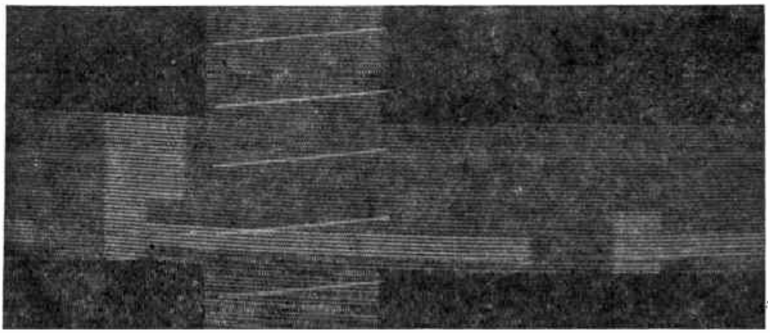

Fig. 5—Pulse cross pattern photograph showing complete lack of coincidence.

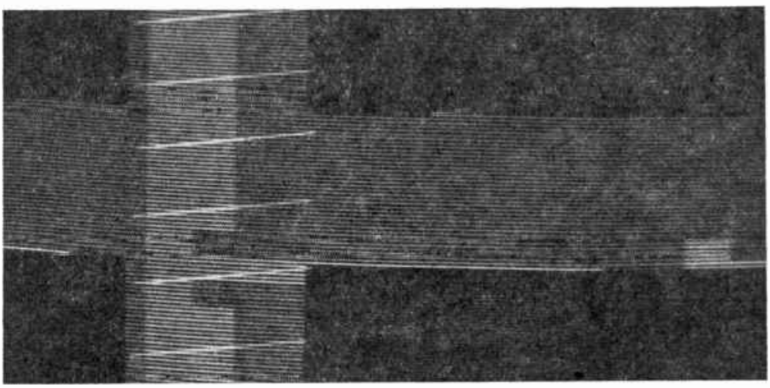

Fig. 6—Pulse cross pattern photograph showing coincidence at line frequency but not at field frequency.

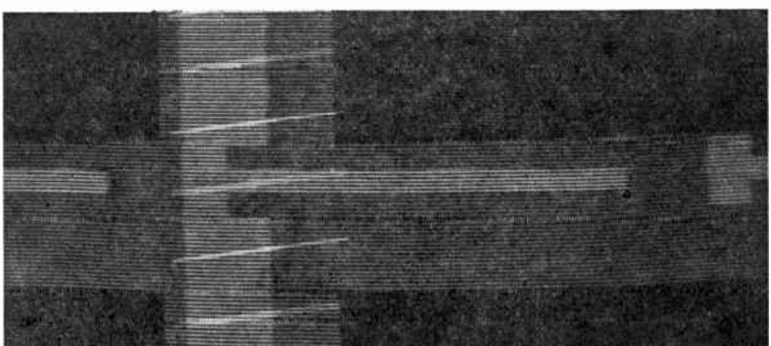

Fig. 7—Pulse cross pattern photograph showing odd vs. even field matching (the last line frequency synchronizing signal pulse is superimposed on the first line of the field frequency blanking.)

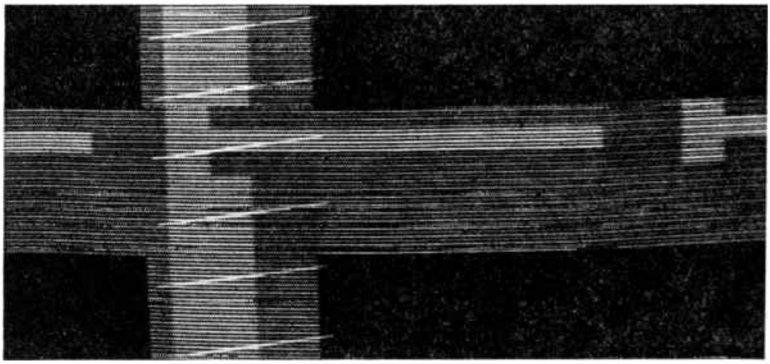

Fig. 8—Pulse cross pattern photograph showing odd vs. even field matching (the first equalizing pulse is superimposed on the last line frequency blanking pulse.)

on the pulse cross. Variations in the width of the "front porch," or delay of line-frequency synchronizing signal with respect to line-frequency blanking, are easily noted as shown in Figure 10.

The method and equipment outlined function satisfactorily and do not require excessive time in aligning the two signals for coincidence provided the continuity and stability of the received signal are good — i.e., provided one alignment will suffice for the transmission. For regular operational use, a means of quickly reverting to independent systems would be mandatory to cover the contingency of momentary interruption in the incoming signal. A means of automatically establishing coincidence and lock between the two signals as well as reverting to independent operation would of course be a highly desirable feature.

The results obtainable from interlocked scanning will be enhanced by the use of synchronizing generators that are not locked to local

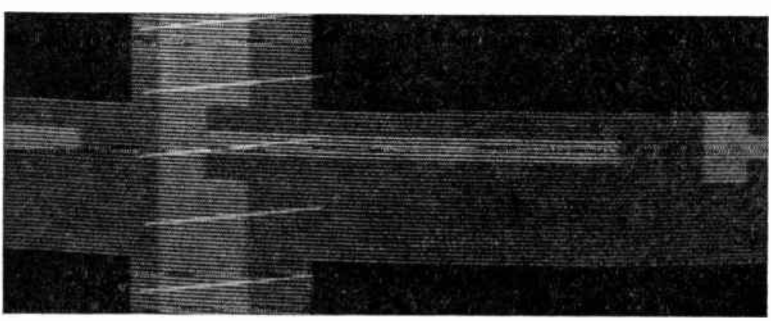

Fig. 9—Pulse cross pattern photograph showing even vs. even field matching

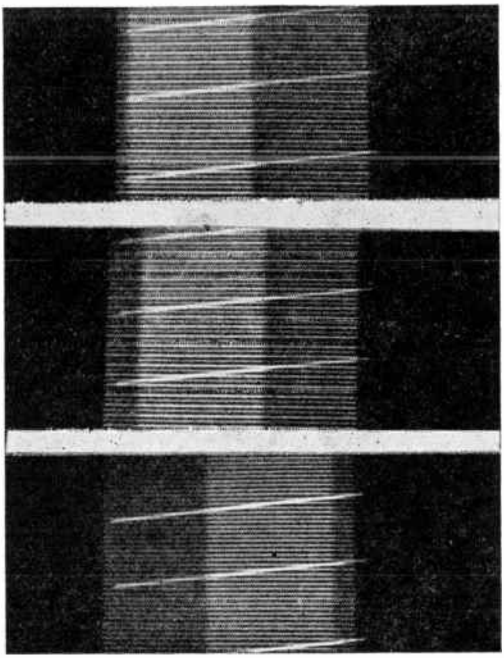

Fig. 10—Portions of pulse cross patterns showing variations in "front porch" or in the delay of line-frequency synchronizing signal with rospect to the line-frequency blanking signal.

60-cycle power supplies, but which are controlled by high-inertia frequency sources. The most obvious benefit from the use of high-inertia systems lies in the fact that a momentary loss of the incoming signal would not ordinarily produce the same discontinuity of control of interlocked scanning that is inevitable with the system described. In fact, it appears possible to provide high-inertia controls for individual synchronizing generators of such excellence that appreciable discontinuity in incoming signal can be tolerated before, from an operational point of view, it would be necessary to sever the interlocked scanning tie-in.

A disadvantage in using the high-inertia frequency control system for synchronizing generator control lies in the fact that projector motors in film studios could no longer be synchronously driven from the local 60-cycle power.

Conclusions

The general direction of work toward one solution to the problem of interlocked scanning systems has been indicated. The work done has served more to show the nature and magnitude of the problems involved than to provide a complete answer. The use of the incoming signal for control is indicated by the economics of the problem.

DEVELOPMENTAL TELEVISION TRANSMITTER FOR 500-900 MEGACYCLES*†

By

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Princeton, N. J.

Summary—*Tubes and circuits have been developed which make possible a one-kilowatt-peak-power wide-band television transmitter for the 500-900-Mc band. A detailed account of these developments is presented.*

The ultra-high-frequency tube employed in this transmitter is a modified form of the 600-Mc oxide-coated-cathode pulse-triode developed during the war[1]. In a push-pull amplifier a pair of these tubes gives a continuous-wave output of more than one Kw at 800 Mc, and in wide-band service such as would accommodate a color television picture of the highest quality, a peak power of more than one Kw is readily obtained. A novel feature of the new tube is the tungsten-wire grid which leads to an unusually rugged tube. Because of excellent grid cooling there is no trace of grid emission; this fact undoubtedly contributes to the stability of the system.

Wide-band operation imposes severe requirements on circuit design. In the rf amplifier, precaution is taken to keep the stored energy low. In the video amplifier, operation at high power level is accomplished by employing the above mentioned pulse triode to cathode-modulate the rf amplifier. This combination of a large-area-cathode "Class-A" modulator tube and an intermediate-area-cathode "Class-B" rf tube is advantageous for wide-band service.

Introduction

DURING the course of the development of a 600-Mc oxide-coated-cathode pulse triode during the war[1], it became apparent that a modified version of this tube would be advantageous for continuous-wave applications. Consideration of the fundamental electronics of ultra-high-frequency transmitting tubes[2-7] indicated that

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* Decimal Classification: R583.4 × R583.6

† Reprinted from *RCA Review*, December, 1948.

[1] R. R. Law, D. G. Burnside, R. P. Stone, W. B. Whalley, "Development of Pulse Triodes and Circuit to Give One Megawatt at 600 Megacycles" *RCA Review*, Vol. VII, No. 2, pp. 253-264, June, 1946.

[2] D. C. Prince, "Vacuum Tubes as Power Oscillators", *Proc. I.R.E.*, Vol. II, p. 275, June; p. 405, August; and p. 527, October, 1923.

[3] W. G. Wagener, "The Developmental Problems and Operating Characteristics of Two New Ultra-High-Frequency Triodes", *Proc. I.R.E.*, Vol. 26, pp. 401-414, April, 1938.

[4] A. V. Haeff, "Effect of Electron Transit Time on Efficiency of a Power Amplifier", *RCA Review*, Vol. IV, No. 1, pp. 114-122, July, 1939.

[5] C. C. Wang, "Large-Signal High-Frequency Electronics of Thermionic Vacuum Tubes", *Proc. I.R.E.*, Vol. 29, pp. 200-214, April, 1941.

[6,7] See following page.

such a tube might perform satisfactorily in the 500- to 900-Mc band, particularly if the interelectrode spacing could be kept small.

On the basis of this theory and as a result of preliminary tests it appeared that a push-pull rf power amplifier employing a pair of these tubes should give one-Kw continuous-wave at 800 Mc. Furthermore, from an estimate of the ratio of stored-to-active energy, it seemed that such an amplifier should have a bandwidth of between 15 and 20 Mc. In the light of the thinking at that time, development of wide-band transmitters to explore the possibilities of television at higher frequencies was very desirable. Also, there were a number of basic questions to be answered. Are oxide-coated-cathodes suitable for use in moderate power ultra-high-frequency transmitting tubes? What is the best way to modulate the triode? And finally, is neutralization necessary in grounded-grid circuits? This paper outlines the results of a study to answer these and other questions.

TUBE DEVELOPMENT

The essential features of the tube developed for these tests may be seen in Figure 1. Starting at the lower left and proceeding clockwise around the photograph may be seen: 1) the stem assembly; 2) the cathode-support assembly; 3) the heater; 4) the cathode; 5) the cathode-heater assembly; 6) the cathode-stem assembly; 7) the grid assembly; 8) the grid-stem assembly; 9) the final assembly; and in conclusion, in the center of the photograph, the complete tube including air-cooled radiator.

The cathode is approximately one and one-half inches in diameter and has a coated area of about 7.5 square centimeters. The heater, a single helical coil of 0.010 inch diameter insulated tungsten wire, requires about 3 amperes at 18 volts to bring the cathode to operating temperature in the absence of back bombardment.

The grid consists of 180 pieces of 0.007 inch diameter tungsten wire silver-soldered to the oxygen-free-high-conductivity copper cap and support cone. The ends of the tungsten wires are lightly nickel-plated to facilitate "wetting" by the solder. In order that none of the wires shall inadvertently bow inward, they are sprung outward by a stainless-steel jig which holds them as they are soldered in a hydrogen-atmosphere furnace. The slots in the base of the grid cone provide flexibility to mitigate ill effects from differential expansion between

[6] G. J. Lehmann and A. R. Vallarino, "Study of Ultra-High-Frequency Tubes by Dimensional Analysis", *Proc. I.R.E.*, Vol. 33, No. 10, p. 663, October, 1945.

[7] R. R. Law, "Electronics of Ultra-High-Frequency Triodes", accepted for publication in *Proc. I.R.E.*

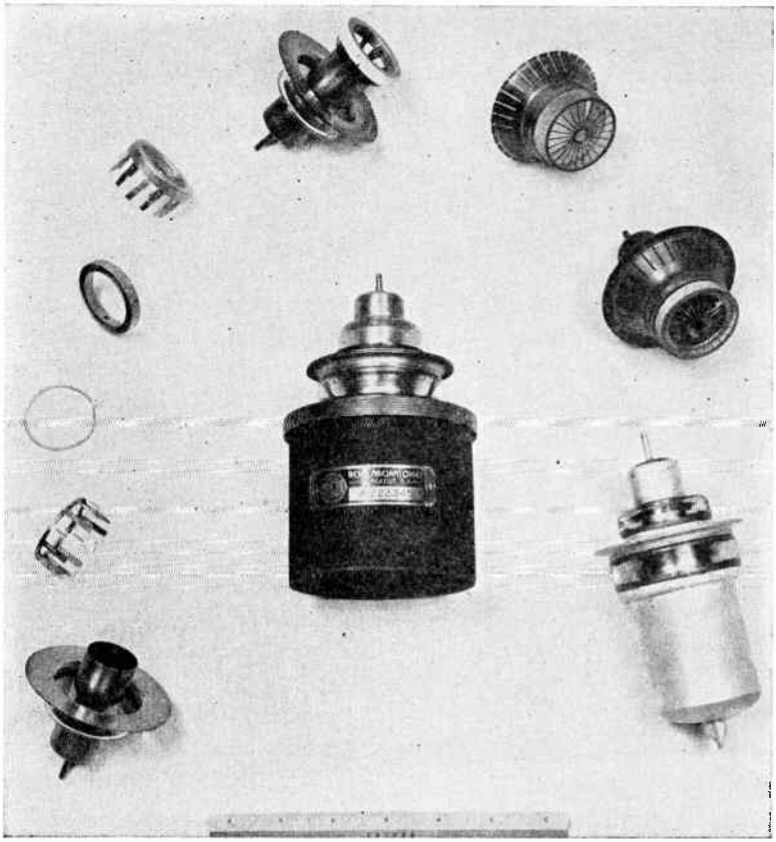

Fig. 1—Details of ultra-high-frequency continuous-wave triode.

the grid cone and the Kovar grid flange. The grid cone is spot welded directly to the grid flange. An electroplated layer of nickel on the base of the cone facilitates making this weld. Because of the excellent grid cooling there is no trace of grid emission. This construction makes possible an unusually rugged tube, and the wires are so refractory that the grid is almost indestructible.

One of the perennial problems in making vacuum tubes is how to seal in the mount or assembled tube without damaging critical parts. Although use of a nitrogen atmosphere to prevent oxidation[8] will suffice for thoriated-tungsten filament tubes, oxide-coated-cathode tubes are more susceptible to damage by overheating and contamination. "Cold-sealing" has been accomplished by radio-frequency-heating tech-

[8] S. Frankel, J. J. Glauber, J. P. Wallenstein, "Medium-Power Triode for 600-Megacycles", *Proc. I.R.E.*, Vol. 34, pp. 986-991, December, 1946.

niques[9], but for experimental tubes the equipment required for this process cannot be justified. In view of the attractiveness of the gold-diffusion seal[10], development of a technique for joining Kovar-to-Kovar by this method was undertaken. It was found that such a seal could be made by the following procedure: 1) copper-plate the Kovar parts, approximately 0.001 inch thick; 2) fire in hydrogen atmosphere for 20 minutes at 1030°C to bond the copper to the Kovar; 3) perform glassing operation in the conventional manner; 4) remove oxide and again copper-plate, the copper layer so applied adheres firmly to the Kovar and provides a vacuum-tight copper-Kovar bond; and finally, 5) a gold ring compressed between the two copper-plated Kovar parts serves to unite the parts during baking by the familiar gold-diffusion process.

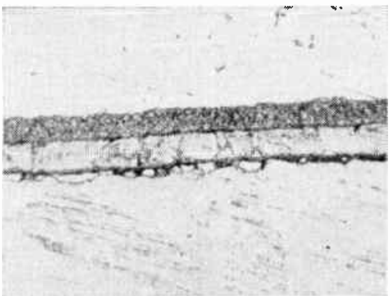

Fig. 2—Diffusion seal for bonding Kovar-to-Kovar.

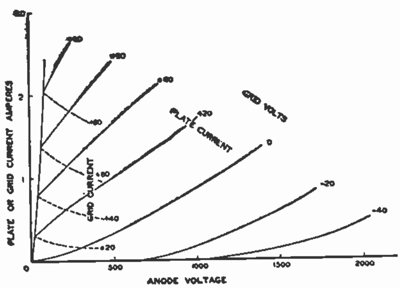

Fig. 3—Static characteristics of ultra-high-frequency continuous-wave triode.

Figure 2 is a photomicrograph of one side of such a composite seal. The Kovar is below, the gold above. The lighter layer between the gold and Kovar is the copper-plate that was fired at 1030°C. The dark layer is unfired copper. There is no visible evidence that the gold has penetrated the copper during the 450°C bake, but when the parts are forcibly pulled apart a layer of gold adheres to the copper. This sealing technique permits assembly of the tube parts without contamination and in the present case does away with the all too familiar problem of poisoning of the oxide-coated-cathodes. Furthermore, the better vacuum obtainable with cleaner parts undoubtedly improves life.

The static characteristics of this tube are shown in Figure 3. For purposes of comparison, its continuous-wave performance as an oscil-

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[9] W. P. Bennett, E. A. Eshbach, C. E. Haller, W. R. Keye, "A New 100-Watt Triode for 1000 Megacycles", *Proc. I.R.E.*, Vol. 36, No. 10, pp. 1296-1302, October, 1948.

[10] J. B. Fiske, H. D. Hagstrum, L. P. Hartman, "The Magnetron as a Generator of Centimeter Waves", *The Bell System Technical Journal*, Vol. XXV, No. 2, April, 1946.

lator at 800 Mc is shown in Figure 4. On the basis of this performance the empirical relation for efficiency[7] indicates that its equivalent cathode-grid spacing is 0.015 inch. This checks with the dimensions of the parts when allowance is made for the fact that the hot spacing is approximately 60 per cent of the cold spacing.

Preliminary life tests were made at power levels corresponding to 300, 400, and 500 watts continuous-wave output in rf amplifier service. With the limited data available there is no correlation between life and power level. Inasmuch as the 300-watt-continuous-wave level with a single tube corresponds to transmitting one-kw-peak-power with a "black" picture in the two-tube television transmitter, the remainder of the tests were run at the 300-watt-continuous-wave level.

The primary factor affecting life is cathode temperature. Because of back bombardment resulting from transit-time effects, the heater input must be materially reduced. In practice, it is found that after the tube is up to normal power, the heater power should be reduced about 50 per cent. An additional point of interest came out of the life test data in view of the fact that early samples employed a cathode coating of less than 3 milligrams per square centimeter, whereas 7 to 12 milligrams per square centimeter is recommended. This compromise was introduced in the early tests to minimize cathode peeling during manufacture which frequently occurs during humid weather. These samples gave 400 to 600 hours life. Later samples employing the recommended coating gave more than 1,000 hours life. Evidently life is proportional to cathode thickness.

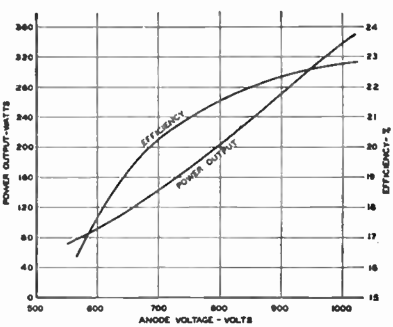

Fig. 4—Performance of continuous-wave triode as oscillator at 800 Mc.

Circuit Development

From the static characteristics of Figure 3 and the experimentally determined optimum operating point it may be deduced that the rf anode voltage swing is about 1,000 volts peak-to-peak under conditions corresponding to one kilowatt power output from the two-tube transmitter. In view of the 0.080 inch grid-anode spacing, the energy stored at the tube electrodes is about 1.0 micro-joule or less than 0.06 micro-joule per square centimeter. The active energy, or energy supplied by the tubes on the other hand is 1.2 micro-joules or approximately 0.08 micro-joule per cycle per square centimeter at 800 Mc. Inasmuch as

$$Q = (2\pi)\ (\text{Stored Energy})/(\text{Active Energy per Cycle})$$

and bandwidth or frequency separation between half-power points Δf is[11]

$$\Delta f = (\text{Operating Frequency})/(Q \text{ of System}),$$

the bandwidth resulting from the capacitance of the tube elements themselves would be greater than 180 Mc. This bandwidth is so large as to be of little practical significance other than to emphasize the fact that the circuit will be the determining factor.

In the case of the cathode-modulated rf amplifier the energy stored in the cathode-grid circuit has no adverse effect on bandwidth. In fact, it may be desirable to store energy in this circuit to stabilize the driver during peak power pulses. In contrast, energy stored in the anode circuit is very detrimental. Unfortunately the present tube cannot be operated in fundamental-mode circuits at these frequencies, and harmonic-mode operation materially increases the stored energy. For push-pull operation the anode line must be three-halves wavelength long. Connections for dc are made at the midpoint. Furthermore, if the circuit is to be tunable over an appreciable frequency range, the electrical position of the discontinuity between tube and circuit will shift. In view of this, it is desirable to keep the characteristic impedance of the circuit so low that special means of transformation will not be required. This is possible in the present case. The four-inch-diameter-internal-conductor one-quarter-inch-radial-separation line even with the most unfavorable match gives rise to a maximum voltage of less than 600 volts peak-to-peak. On this basis the maximum energy stored in the circuit at 800 Mc would be about 5 micro-joules.

To complete the estimate of bandwidth, allowance must be made for the energy stored in the unmatched portion of the output system. Because it is difficult to adjust the load coupling loop size, matching stubs are desirable. With a voltage-standing-wave ratio of $\sqrt{2}$, the energy stored in the stubs and unmatched portion of the output system is about 3.0 micro-joules. To sum up, at 800 Mc the energy distribution would be:

Energy stored in tubes	1.0 micro-joules
Energy stored in circuit	5.0 micro-joules
Energy stored in matching stubs	3.0 micro-joules
Total energy stored	9.0 micro-joules
Active energy per cycle	1.2 micro-joules

[11] F. E. Terman, RADIO ENGINEERING HANDBOOK, McGraw-Hill Book Co., Inc., New York, N. Y., 1943, pp. 429-430.

whereupon the estimated bandwidth is

$$f = \frac{(800)\,(1.2 \times 10^{-6})}{2\pi\,(9.0 \times 10^{-6})} = 17 \text{ megacycles.}$$

In practice the bandwidth may be somewhat greater than this if the load-coupling-loop match or the tube-to-circuit match are more favorable.

The physical layout of the circuits developed for this transmitter may be seen in Figures 5 and 6. In Figure 5, the unit on the left is the driver and the unit on the right is the cathode-modulated rf amplifier. The coaxial line serving to transmit power from the anode cavity of the driver to the cathode cavity of the amplifier with its

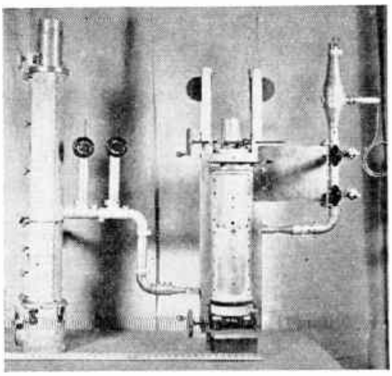

Fig. 5—500-900 Mc driver and rf power amplifier.

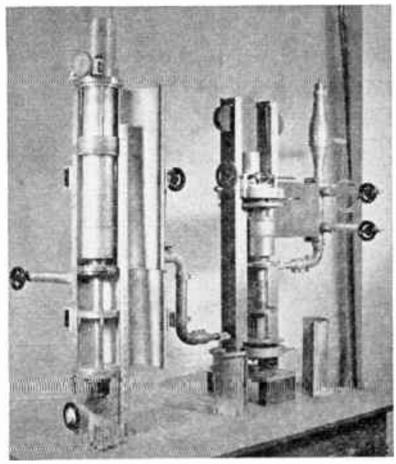

Fig. 6—Rf circuit details.

matching stubs may be seen in the center of the picture. The coaxial line serving to transmit power from the amplifier to the water-cooled resistance load with its matching stubs may be seen in the right of the picture. In Figure 6, the covers of the units have been swung open or partially removed to reveal circuit details.

To take advantage of the bandwidth capabilities of this amplifier, the wide-band modulator shown in Figure 7 was developed. This modulator employs a pair of the aforementioned pulse triodes[1] to cathode-modulate the rf amplifier. These modulator tubes make possible a current change of more than two amperes at an equivalent transconductance of 0.15 amperes per volt. Such a combination of a large-area-cathode "Class-A" modulator tube and an intermediate-area-cathode "Class-B" radio frequency tube is advantageous in wide-band service. The cathode-coupled stage ahead of the pulse triodes serves to lower the effective input capacitance presented to the video amplifier.

Bias to the cathodes of the amplifier and final modulator stage is supplied through the special choke "L" which is split into sections of progressively increasing size to provide relatively high impedance over the video band. So long as this impedance is large compared to $R_p/(1+\mu)$, the gain is substantially constant and equal[11] to $\mu/(1+\mu)$. In operation, the dc current to the anodes of the amplifier and modulator is substantially constant; modulation at video frequencies serves to switch the current from the amplifier to the modulator and vice versa.

PERFORMANCE OF SYSTEM

The overall performance of the rf power amplifier and composite system is indicated in Figures 8, 9, and 10.

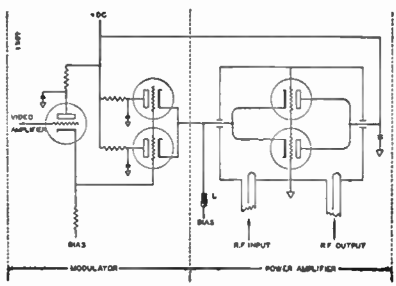

Fig. 7—Wide-band modulator circuit.

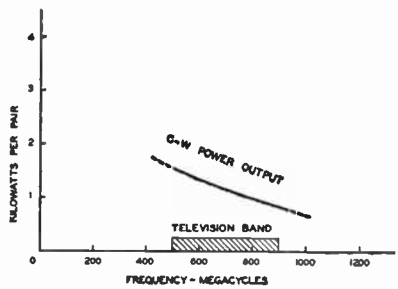

Fig. 8 — Variation of rf power amplifier output with frequency.

Figure 8 shows the variation of continuous-wave power output with frequency at a fixed anode voltage of 900 volts. It will be observed that the anode efficiency of the amplifier is appreciably higher than that shown for the oscillator in Figure 4. The difference arises from the fact that the oscillator must supply its own exciting power whereas the amplifier does not. Inasmuch as the power gain of this tube as an amplifier under typical operating conditions is approximately three, the amplifier output and efficiency may be expected to be about 50 per cent greater than that of the oscillator

Figure 9 indicates the broadband potentialities of the amplifier. Here is plotted the relative voltage in the sidebands as a function of sideband frequency relative to the 800 Mc carrier. The circled points joined by the solid line were taken by direct measurement of sideband voltage in a calibrated spectrum analyzer. Conjugate points were measured with the same circuit adjustment; in each case adjustments were made so that both sidebands would have equal amplitude. The ease with which these adjustments could be accomplished suggests that the response is nearly symmetrical. Due to the finite bandwidth of the spectrum analyzer it was not possible to measure the response immediately adjacent to the carrier. The intermediate region was checked by direct observation of percentage modulation on the oscilloscope[12]. To increase the accuracy of readings the data were taken at high percentage modulation. Thus, the amplitude of the modulating voltage was adjusted to give 90 per cent modulation at low frequency, whereupon its amplitude was held constant as its frequency was varied. Because of this choice, the ordinate scale of Figure 9 corresponds to percentage modulation as well as relative sideband voltage.

As may be seen from this response curve, the —3-db points are approximately 19 Mc apart. If vestigial sideband operation were employed, such an amplifier would accommodate a color television picture of the highest quality.

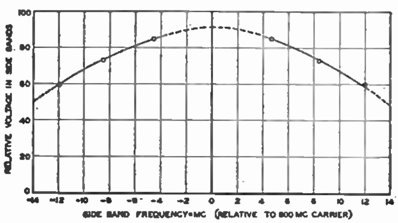

Fig. 9—Wide-band response of rf power amplifier.

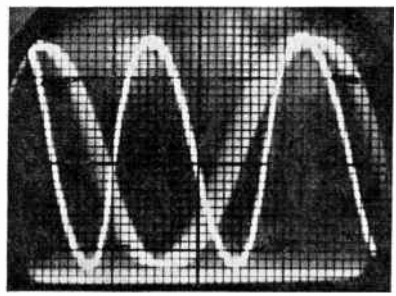

Fig. 10—Rectified rf output with sine wave modulation.

The linearity of modulation is indicated in Figure 10. This photograph shows the rf load-diode-voltage as a function of time with 1.6 Mc sine wave modulation. Zero power level as established by the vibroswitch[12] is indicated by the horizontal line at the bottom of the picture. Under the conditions portrayed the peaks correspond to 1.4 Kw and the modulation is very nearly 100 per cent.

Contrary to the experience of previous workers[8,13,14] there has been no indication of a need for neutralization. This difference arises from the low impedance of the present tubes as compared to those previously available. Although it is possible to adjust the present amplifier so that it will oscillate, this adjustment is far from the normal operating

[12] T. J. Buzalski, "A Method of Measuring the Degree of Modulation of a Television Signal", *RCA Review*, Vol. VII, No. 2, pp. 265-271, June, 1946.

[13] E. Labin, "Design of the Output Stage of a High Power Television Transmitter", *Electrical Communications*, Vol. 20, No. 3, p. 193, 1942.

[14] C. E. Strong, "The Inverted Amplifier", *Electronics*, Vol. 13, pp. 14-16, July, 1940.

point. Because of the relatively high mu, the cathode-anode capacitance is low and the undesired cathode excitation is small in comparison to the driver excitation required for the low impedance system throughout the 500-900 Mc frequency range. Oscillation is possible only when the amplifier is unloaded and when the anode and cathode cavities are detuned to give the appropriate grid-anode-voltage phase relationship[15].

Conclusions

Tests in a developmental 500-900 Mc television transmitter indicate that oxide-coated cathode triodes may be used for moderate power ultra-high-frequency transmitter applications. Although modulation of the triode presents a serious problem, for wide-band service this difficulty is in part overcome by employing a large-area-cathode "Class-A" modulator tube to cathode-modulate an intermediate-area-cathode "Class-B" rf tube. With tubes and circuits of proper design operating under wide-band conditions there is no need for neutralization in this frequency range.

Acknowledgments

The writers wish to express their appreciation for the stimulating discussions and practical assistance given by many members of the RCA Laboratories Technical Staff and Service Groups.

[15] E. E. Spitzer, "Grounded-Grid Power Amplifiers", *Electronics*, Vol. 19, pp. 136-141, April, 1946.

MOTION PICTURE PHOTOGRAPHY OF TELEVISION IMAGES*†

BY

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Summary—The permanent recording of television programs, for documentary, historical, legal, or critical purposes and as an aid to networking may be accomplished by motion picture photography of the television image, making use of the practical arts developed by motion picture engineers for recording the sight and sound of a television broadcast. This paper describes the apparatus and methods developed for the photographing of the television cathode-ray image. Development is traced from the first attempts in 1938 through the experimental cameras to the commercial camera system now in use. Subsequent sections deal in some detail with 16- and 35-millimeter equipment considerations, kinescope phosphors and film spectral characteristics, resolving power of films, exposure of film, processing and printing of kinescope film, the photographic monitor, and sound recording.

INTRODUCTION

A METHOD of permanently recording the otherwise transient video signal is desirable in the advancement of television art. Motion picture photography of the television image is a method of doing this. Use is made of the practical arts developed by the motion picture engineers in recording the sight and sound of a television broadcast. The reasons for recording television programs are manifold and are similar to the reasons for recording the sound of a standard broadcast program. Television recordings may be made for documentary purposes, to preserve an historical event, for legal purposes, and for critical purposes. These recordings may be used for a delayed or a repeat broadcast or for syndication to other television stations unable to obtain network programs because of the lack of coaxial cable or microwave relay connections with network sources. Television recordings of auditions are useful in the marketing of programs or talent.

It is the purpose of this paper to describe the apparatus and methods developed for the motion picture photography of the television cathode-ray image.

In 1938, the first attempts were made to photograph the television image on motion picture film. Kinescopes or cathode-ray tubes at that time used low efficiency phosphors and operated on relatively low

* Decimal Classification: R583 × R582.

† Reprinted from *RCA Review*, June, 1948.

second-anode voltages compared with present day practices. The amount of light obtainable from these cathode-ray tubes was not enough to produce a full exposure on the fastest films then obtainable with an exposure of one thirtieth of a second in a sixteen frame-per-second camera. By photographing the cathode-ray tube at eight frames per second with an exposure time of one fifteenth of a second, recognizable images were obtained on the motion picture film. Of course, these films when projected on a twenty-four frame projector show an unnatural rapidity of motion and are considered to have nothing more than historical interest.

The cameras used in these early experiments were spring-motor driven and the shutter rate was therefore nonsynchronous with the frame rate of the television system. This gave rise to phenomena termed "shutter bar", or banding — a black or a white bar which in a nonsynchronous system moves across the film image when projected, at a rate dependent on the difference in frequency between that of the television system and the frame rate of the motion picture camera. The width of the bar depends on the shutter angle of the camera. If the shutter angle and the frame rate of the camera combine to give an exposure time of less than one thirtieth of a second, less than a full television frame is photographed and a black or under-exposed section of the image results. If the exposure time is greater than one thirtieth of a second, there is an overlapping of the television image frame, i.e., a full frame plus part of the succeeding frame are photographed. This results in a white or over-exposed section on the film frame. It is apparent from this that the shutter speed of any camera used to photograph the television image should be precisely one thirtieth of a second or a multiple of one thirtieth such as one fifteenth or one tenth of a second if "shutter bar" is to be avoided. This rule applies to still cameras as well as to motion picture cameras. The degree of contrast between the under-exposed or over-exposed portion or banded section of the image and the correctly exposed portion decreases with the multiple increase in exposure in units of one thirtieth of a second. If the exposure is less than one thirtieth of a second, the error in exposure will be fifty per cent; for an exposure just under a fifteenth of a second, the error will be twenty-five per cent; and for an approximate tenth of a second exposure, the error will have decreased to twelve and one-half per cent. Since it is impossible to find still cameras with a shutter accuracy of the degree necessary to photograph the television image at a thirtieth of a second, better results can be obtained at the slower shutter speeds in respect to uniformity of the photograph. The problem of photographing rapid motion on the television screen with a still camera is serious because of the distortion of motion, but no more so than in the case of direct photography.

Motion picture photography of television images was undertaken during the war to record television transmissions from cameras installed in aircraft and in guided missiles. Due to the conditions under which the television images generated by the Block[1] and Ring[2] equipment used in these tests were recorded, nonsynchronous cameras operated by batteries or spring motors at approximately sixteen or eight frames per second were used. "Shutter banding" was noticeable in these films which did not destroy their value in the studies then underway.

Some further work was done with an Eastman Cine-Special driven with a synchronous motor at fifteen frames per second. The shutter on this camera is open for 170 degrees which results in an exposure just under one thirtieth of a second. By phasing the motor drive so that the shutter opened and closed during the vertical blanking period of the television image, acceptable results without banding were obtained. When these fifteen frame-per-second recordings were projected through a standard silent projector at sixteen frames per second, no undesirable results due to change of speed were noticeable.

For a complete recording of a television program it is necessary to record the sound portion. Present day motion picture practice is to record sound at a twenty-four frame per second rate. It is desirable that recordings of television programs be capable of being played back on standard motion picture sound projectors. This necessitates the adoption of the standard twenty-four frame rate to record television programs.

Experimental Camera

A method has been devised of recording twenty-four frames of the standard thirty frame television signal.[3] The equipment now in use uses this method. A shutter driven by a sixty-cycle synchronous motor at twenty-four cycles per second is utilized. This shutter has a closing angle of 72 degrees and an opening of 288 degrees. At the twenty-four cycle per second rate these angles represent a closing time of 1/120 of a second and an opening time of 4/120 or 1/30 of a second, which is the time for one full television frame. Figure 1 shows the time sequence of such a shutter in relation to the television scanning cycle. The camera is driven by a synchronous motor from the same source of 60

[1] M. A. Trainer and W. J. Poch, "Television Equipment for Aircraft", *RCA REVIEW*, Vol. VI, No. 4, pp. 469-502, December, 1946.

[2] R. E. Shelby, F. J. Somers and L. R. Moffett, "Naval Airborne Television Reconnaissance System", *RCA REVIEW*, Vol. VI, No. 3, pp. 303-337, September, 1946.

[3] D. W. Epstein—U. S. Patent No. 2,251,786.

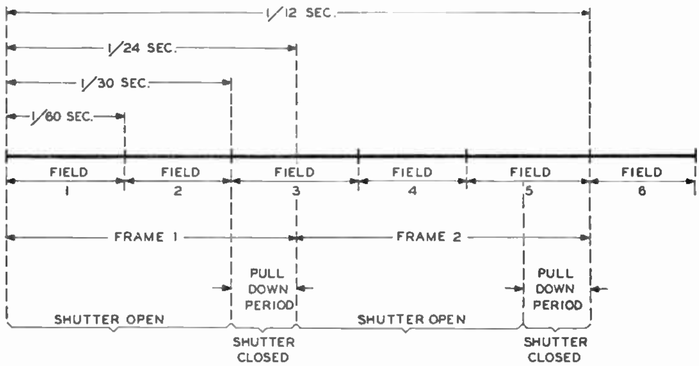
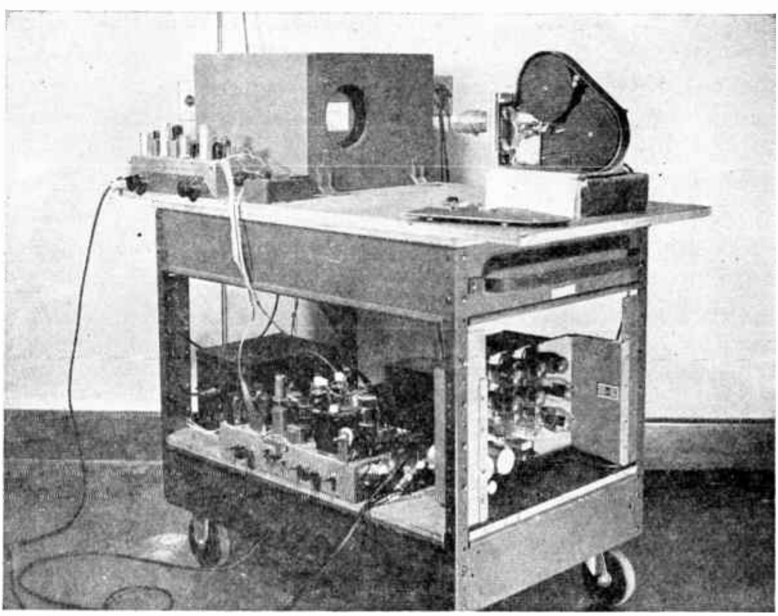


Fig. 1—Time sequence of exposure and pulldown timing of the camera in relation to the field rate of the television image.

cycle current as is used for the television synchronizing generator. If the phasing of the motor is such that the camera shutter opens on the beginning of one field, it remains open for that and the succeeding field, then closes. The shutter remains closed for 1/120 of a second or half the third field, while the film is advanced, then opens. It remains open for this last half of the third field, for the full succeeding fourth field, and for the top or first half of the fifth field. The shutter then closes at the same point that it opened in relation to the scan of

Fig. 2—"Breadboard" camera and photographic monitor.



the image. A half field later the shutter opens completing the cycle. Pull-down of the film occurs during this time that the shutter is closed.

The one hundred twentieth of a second time allowed for pulldown is considerably less than is found in standard motion picture cameras. The normal pulldown time of a standard camera is between a fortieth and a fiftieth of a second. To achieve this high rate of pulldown in the television recording camera without undue strain on the film and to maintain registration of the film image is quite an undertaking. That it has been solved satisfactorily is to the credit of the motion picture engineers.

The Eastman Kodak Company constructed a "Breadboard" camera using the foregoing principles. This camera is shown in Figure 2. The camera was capable of a 200-foot load of 16-millimeter film, allowing the recording of 5½ minutes of program time. The satisfactory operation of this camera proved the 288-degree shutter to be practical in the photography of the thirty-frame television image at a twenty-four frame rate.

Commercial Camera

The Eastman Kodak Company in cooperation with the National Broadcasting Company, encouraged by the successful operation of the breadboard camera, began the design of a commercial recording camera capable of recording a half hour of program with a 1200-foot load of 16-millimeter film. The design of this camera was quite complicated by a number of factors. It had been determined through tests with the breadboard camera that the shutter has to rotate with a low flutter rate. A slight change in angular speed of the shutter results in banding of the film image. In severe cases this banding alternates from black to white on alternate film frames. It is therefore necessary to design the shutter drive to have the utmost constancy of angular speed. This is accomplished by using an 1800 revolutions-per-minute synchronous motor to drive the shutter at the necessary 1440 revolutions-per-minute rate through a set of precision gears. Another synchronous motor of larger capacity is employed to drive the film transport mechanism and the Geneva intermittent. The two motors are kept in step during the starting and stopping periods by a phase coupling device which allows the stronger of the two motors to assist the weaker until they both reach synchronous speed. The coupling then floats so that there is no physical connection between the motors.

The shutter motor then drives the shutter independently of any varying or intermittent change of load in the camera. The armature of this motor acts as a flywheel to damp out any tendency to flutter in the dynamically balanced shutter blade.

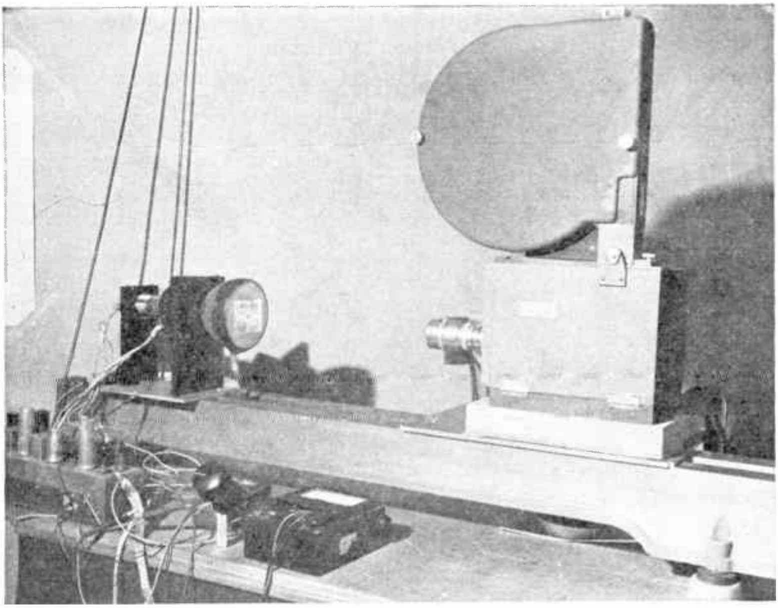


Fig. 3—The Eastman Kodak television recording camera.

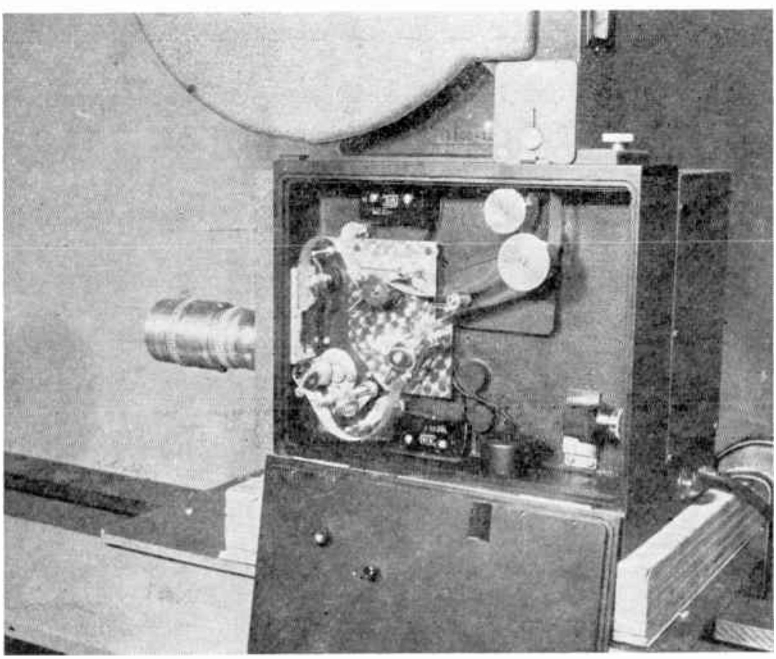


Fig. 4—Eastman Kodak camera interior showing film threading.

An eight tooth sprocket pulldown actuated by an accelerated Geneva star is employed for film pulldown. The pulldown angle of approximately 57 degrees is obtained by means of a spline-and-slot type of accelerator interposed between the constant speed shaft and the geneva driver.

The Eastman Kodak television recording camera is shown in Figures 3 and 4. Nylon is used in the film gate and pressure pad to minimize emulsion pile up, a potent source of trouble in motion picture cameras.

All friction points in the takeup side of the 1200-foot magazine are equipped with ball bearings so that take-up of film progresses smoothly from the two-inch core diameter to the full ten-inch diameter of the full 1200-foot roll. Loop loss indicators are provided and actuate microswitches in the event of loop loss, lighting a warning lamp mounted on the base of the camera.

Focussing and framing of the picture frame is done by means of a right-angle view finder equipped with a magnifying lens. This unit is snapped under the pad spring in place of the pressure pad. Visual focusing is done by means of this finder and checked by exposing film at several different settings about this visual optimum. The processed film is examined under a microscope to determine actual best focus.

The lens equipment of the camera is a 2-inch Eastman Anastigmat f1.6. Apertures of f2.0 to f2.8 are normally used to avoid sharpening the shadow of the shutter in the film plane during the opening and closing time, thereby reducing the possibility that a slight timing error may result in banding. It has been determined that at apertures of f5.6 and above, the cutoff of the shutter in the image field becomes abrupt and causes banding on the order of two or three television lines.

16- and 35-Millimeter Equipment Considerations

There are several reasons for the choice of 16-millimeter film for Kinescope recording rather than 35-millimeter. The main reason is that the cost of 35-millimeter is somewhat more than three times the cost of 16-millimeter for the same period of recording. The current quality of television images, which will undoubtedly undergo gradual refinement, is considered to be roughly equivalent to 16-millimeter home movies, although actually somewhat better with reference to contrast and detail. No marked improvement, however, is to be had by recording on 35-millimeter rather than 16-millimeter at the present time. With the use of fine grain high resolution 16-millimeter film emulsions, no loss of resolution in recording the television image is noticeable.

Fire regulations covering the use of 35-millimeter film, which apply regardless of whether the 35-millimeter film is acetate safety base or the combustible nitrate base, are rigorous. The cost of providing space that meets these regulations for the use of 35-millimeter film is extremely high, and the changes needed in existing space are difficult to accomplish. 16-millimeter films are available only in acetate safety base which is classified by the Underwriter Laboratories as having a safety factor slightly higher than that of newsprint. The use of 16-millimeter films, therefore, are not restricted by fire regulations. It should be noted that in New York City these restrictions apply to space in which equipment capable of operating with 35-millimeter film is installed, so in order to forestall trouble, all equipment should be single purpose 16-millimeter equipment rather than dual purpose 35- or 16-millimeter equipment.

Another factor in the choice of 16-millimeter film is the high cost of 35-millimeter projection equipment. Most television stations are providing projection facilities for 16-millimeter film only for this reason. In order to service these stations with syndicated programs photographed from the kinescope, 16-millimeter prints will be needed.

Kinescope Phosphors and Film Spectral Characteristics

The spectral sensitivity of the film emulsion can be matched to the phosphor spectral characteristic for the greatest actinic efficiency. There are three general classifications of film emulsions in terms of their spectral characteristics:

1. *Panchromatic*, sensitive from the ultraviolet (4000 Angstrom Units (A°)) through the red (7000 A°). The spectral response of these emulsions correspond approximately to that of the eye and so are generally used for direct photography;

2. *Orthochromatic*, sensitive from the ultraviolet through green, (5700 A°) is used in direct photography where it is desirable to reduce the red sensitivity;

3. *"Ordinary"* or non-color sensitive emulsions, nonsensitized, responding to the ultraviolet and blue portions of the light spectrum. This type of emulsion is used in coating films and papers generally employed in making positive prints from negatives. In 16-millimeter form it is economical in comparison to the panchromatic and orthochromatic types. Another advantage is the ease of handling as relatively bright safelights may be used.

To match these film characteristics, kinescope phosphors are available with light output ranging from the ultraviolet through the entire visible spectrum. Three types of phosphors in common use in television techniques are as follows:

1. *P1, green fluorescence*, commonly used in oscillographic work.

It is the most efficient visually, but has poor actinic efficiency.

2. *P4, white fluorescence*, used for black and white reproduction of television images in most home receivers. This phosphor has a high output in the blue and green portions of the spectrum, but is down in the red. It has the advantage in kinescope photography that picture quality is most readily judged visually.

3. *P5 and P11*, these two phosphors are blue with high ultraviolet output. Photographically, they are very efficient. There is the difficulty in using a blue phosphor in judging the quality of image visually, due to the fact that the human eye has a low response in the blue region and cannot evaluate the quality of the ultraviolet component of the image light output at all.

Tests have been made on the P11, zinc sulphide, phosphor as to the relative actinic efficiency to the panchromatic, orthochromatic, and "ordinary" non-color sensitive emulsions. A Weston exposure meter was used to determine the light output of the aluminized P11 phosphor kinescope. A series of exposures were made of the image on the tube and the correct exposure as judged by visual inspection of the negative was chosen. The Weston meter was then set with this exposure data to find the Weston rating of the type of film for the P11 light output. This rating was then compared with the Weston rating for daylight as given by the manufacture of the film used. It was found that the exposure required for the P11 phosphor image for panchromatic film was one sixth that required for white light of equal intensity, for orthochromatic it was one twelfth, and for the ordinary or non-color sensitive stock, the exposure ranged from 1/16 to 1/32 of the exposure needed for white light.

With these facts it is apparent that for recording of television images a zinc sulphide, blue-fluorescing screen is desirable since it makes possible the use of high resolution, low cost, positive type of film stocks.

Resolving Power Of Films

Present day television systems operating on a 30-frame, 525-line standard, are capable of resolving 483 lines in the vertical direction, and inside the studio plant, before being transmitted on the 4.5 megacycle channel of the radio frequency transmitter, of resolving over 600 television lines horizontally.

Manufacturers of photographic film rate the resolution of their products in lines resolved per millimeter. By dividing the 7.2-millimeter height of the 16-millimeter frame, into 483 lines and dividing the result by two to convert from television lines to photographic lines per millimeter, it is found that in order to resolve the television scanning lines, a resolving power of better than 33 lines per millimeter at contrast ranges below 1 to 10 is required. To resolve 600 television lines on 16-millimeter film, the emulsion must have a resolving power of 42 lines per millimeter.

The subject contrast of the test charts used to determine film resolution is of the order of 1 to several hundred times. Super X, a panchromatic emulsion, is rated by the manufacturer at 55 lines per millimeter. In television terms there would be resolved 792 television lines in the 16-millimeter frame. It might be thought that such a film would be suitable for photography of the television image. This is not so, for this film, used in photography of the television image, does not fully resolve the scanning lines.

The reason for this discrepancy lies in the different method used by the manufacturer to rate the film. As pointed out above, the resolution is determined by photographing a chart that has a contrast ratio between the black lines and the white spaces of several hundred times. Resolution is then determined by the point at which these lines are barely resolved on the film, or at a point where the contrast ratio is slightly greater than unity.

The resolving power of film, of television pickup tubes, and of image reproducing tubes, falls off with decrease in the subject contrast of the test target. A film rated at 55 lines per millimeter at a subject contrast of several hundred times may have only a resolving power of twenty lines per millimeter when the contrast is in the order of 1 to 10.

It has been pointed out in the literature[4,5] that the television image may have contrast ratios of fifty times in large areas, falling off to contrast ranges less than 1 to 10 in fine detail. A film emulsion rated at 90 lines per millimeter under normal test conditions has the necessary resolving power at the lower contrast ranges to resolve the required 42 lines per millimeter.

Suitable emulsions with resolving powers in excess of 90 lines per millimeter are to be found in the fine grain sound recording and print stocks. Both low and high gamma emulsions are available. For recording a positive image on the kinescope, the low gamma variable density type of emulsion is used. When recording from a reversed negative image on the cathode-ray tube, a high-gamma variable area or print type of emulsion is used.

Some improvement in the quality of resolution of the photographic

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[4] A. Rose, "A Unified Approach to the Performance of Photographic Film, Television Pickup Tubes, and the Human Eye", *Jour. Soc. Mot. Pic. Eng.*, Vol. 48, No. 10, October, 1946.

[5] O. H. Schade, "Electro-Optical Characteristics of Television Systems: Introduction; Part I—Characteristics of Vision and Visual Systems", *RCA REVIEW*, Vol. IX, No. 1, pp. 5-37, March, 1948.

image is made by the use in the video feed of a phase and amplitude equalizer.* The increase in amplitude or contrast of fine detail that can be obtained by boosting the high frequencies in the kinescope image compensates somewhat for the normal falling off of fine detail contrast in the film image. Phase correction is used to reduce the transient white that follows black. The amount of high frequency peaking that can be used is limited by the noise component of the video signal.

Exposure Of Film

Present day aluminized kinescopes operating at high second-anode voltages in the order of 27 kilovolts are capable of brightness ratios of several hundred times in large areas. Most films used in normal photography can handle a range of this order. However, to make full use of such a latitude requires a very accurate exposure. Generally the object brightness range under controlled lighting arrangements never exceeds a ratio of 1 to 30 in direct photography. It is, therefore, necessary that the contrast of the photographic kinescope be maintained within the limits set by the latitude of the particular film used and that the brightness range be set to duplicate the 1 to 30 ratio used in direct photography so as to duplicate the printing contrast of a normal negative.

This is most conveniently achieved by the following method. A plain raster is used on the kinescope such as would be obtained by the use of the blanking signal or pedestal without picture modulation. The brightness of this raster is varied by means of the video gain control or kinescope grid bias control. The beam current of the kinescope is measured by means of a microammeter. Since the light output of the tube is dependent on the watts input to the screen, the measure of beam current affords a measure of the brightness of the tube. Film is exposed to this raster at beam currents varied by steps. The density of the film processed as a normal negative is measured and plotted against the logarithm of the beam current. A normal negative developed to a gamma of 0.65 which has been exposed to an object with a brightness range of 1 to 30 or in logarithmic units, a range of 1.5, should have a density range from 0.25 in the shadows to approximately 1.4 in the highlights. The change in beam current necessary to produce such a range on the kinescope can be read from the plot of the log beam current and film density. The average brightness of the kinescope with picture then would be set by using a beam current that produces a density in the middle of the above range. The video signal is adjusted

* Designed by E. D. Goodale, Television Development, NBC. It is planned that a paper on the equalizer will be published in the September 1948 issue of *RCA REVIEW*.

to a level that will put the blanking level of the composite signal just at visual cutoff of the kinescope. A picture signal judged to have an a-c axis of 50 per cent should be used for this adjustment. This method is largely empirical, but, with experience on the part of the operator, can be made to give consistent results.

Processing and Printing of Kinescope Film

A number of tests have been made in cooperation with the film manufacturers on the processing and printing of films exposed to the kinescope. Both reversal processing and negative processing of the original film were tried. Results show that standard processing methods result in optimum picture quality. Negatives exposed to television images originating in iconoscope cameras are developed to a gamma of 0.7 as determined by a standard IIB sensitrometric test. Film of orthicon pickups gives best results when processed to approximately 0.6 to 0.65. These are interim values as tests on the processing of films have not been completed.

Printing is done according to standard motion picture laboratory practice. Step printing in which the print stock and negative are exposed to the printing light a frame at a time is preferred over continuous printing, where the negative and print stock run past an illuminated slit at a continuous speed. There is a sufficient amount of slippage between the negative and print stock in the continuous printing process to degrade the resolution of the television image. Contrary to the opinion held by many workers, the fact that the film image of a television image is poorer in resolution than in the case of direct photography does not mean that less care in the handling of the film in printing and in projection can be used. The fact is that the utmost care must be taken to maintain the original quality inherent in the film negative throughout the printing process and in the projection of the resulting print.

Films of iconoscope programs can be printed at one printer light setting: i.e., the densities and contrast range of the film resulting from the recording of the outputs of a number of iconoscope cameras does not change sufficiently to warrant changes in the intensity of the printing light.

In film recordings of programs picked up by orthicon cameras the picture negative must be timed for printing. There is considerable difference in the contrast range between different orthicon cameras. Light changes in the order of 100 per cent are sometimes required when a switch between cameras occurs.

This is an undesirable condition as the timing of negatives is an expensive, time consuming, procedure. The cure for this situation is in better control of the output levels of the various orthicon cameras in the studio. Much of this change can be charged to the fact that the spectral characteristics of the orthicon vary from tube to tube. An orthicon with excessive infrared response has a different tonal graduation as compared to an orthicon with no or little response in this region. If the spectral characteristics of the orthicon can be standardized within closer limits than is now done, much of this timing difficulty in the film recordings may disappear.

In kinescope recordings meant for retransmission through the television system a print gamma of 2.2 and a maximum density of 2.4 are recommended. Further tests may show the desirability of changing these recommendations, but to date the best results in the televising of release prints have been obtained under such conditions.

Emulsion position in the final print is of great importance in television because films may be spliced with other films for special purposes. The use of a non-standard emulsion position requires a change of focus in the film projector when interspliced with films using standard emulsion position. This would require the constant attention of the projectionist to maintain optimum focus throughout the spliced film, therefore it is advantageous to insist upon a standard emulsion position for all film to be used in television. The Society of Motion Picture Engineers' standard for 16-millimeter film is emulsion "toward the screen."

In the recording of television images there are several methods of obtaining the final print:

A. the use of reversible film stock in photographing a positive cathode-ray-tube image (A dupe negative is made of this material and prints are made from this negative. The final prints then have standard emulsion position.);

B. the photography of the cathode-ray-tube image using high contrast positive stock and a negative kinescope image resulting in a positive print from which dupe negatives are made for prints having standard position; and

C. the use of a positive image, photographing with a negative type of film from which final prints are made, resulting in a non-standard emulsion position (However, by reversing the direction of horizontal scanning the original negative may be made to have the same emulsion position as that of a dupe negative. Prints made from this negative, then have a standard emulsion position.)

Other factors must be considered in determining the method of recording. Where it is expected that a great number of prints will be

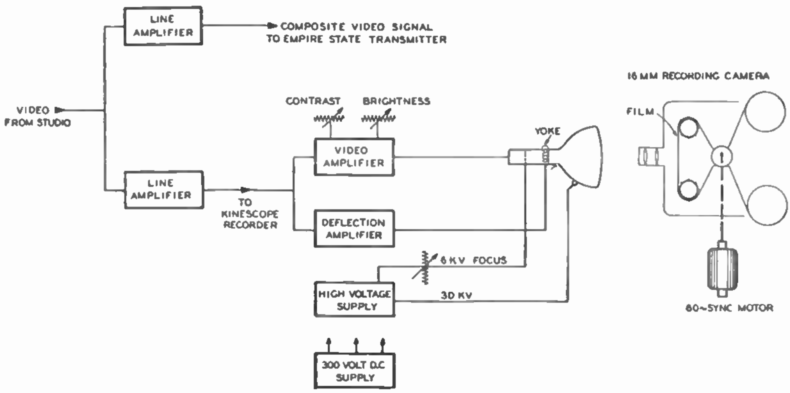

Fig. 5—Kinescope photographic monitor block diagram.

required, methods A or B would be desirable because of the protection of the original material.

Photographic Monitor

The various sizes of cathode-ray tubes may be used for television recordings. In order to insure adequate exposure on the fine grain positive type emulsions, the voltages used with these tubes should be on the order of 20,000 to 30,000 volts. All other factors being equal, such as relative spot size, uniformity of focus over the picture area, brightness and contrast, the smaller tubes offer advantages in the size of the photographic setup. A five-inch kinescope with a flat screen, aluminized P11 phosphor, and the same general type of electron gun

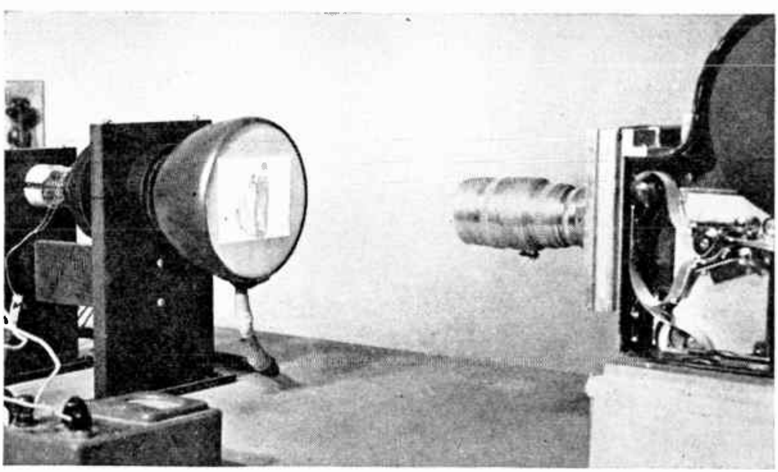

Fig. 6—Close-up of the kinescope with cover removed. (On the tube face is an actual television image of a baseball pickup.)

construction as is used in the 5TP4, gives adequate resolution, contrast and brightness.

A block diagram of the experimental photographic monitor setup is shown in Figure 5 and a photograph of the unit in Figure 6. The video amplifier is flat to eight megacycles, and down ten per cent at ten megacycles. A radio-frequency high voltage power supply delivers 29 kilovolts to the second anode of the kinescope. It should be mentioned that these tubes with the P11 phosphor have a high X-ray output at these voltages as compared with the 5TP4 at the same voltages, and that it is necessary to use more shielding than is evident in the photograph to safeguard personnel.

The deflection circuits follow conventional design. Provision is made to switch the direction of horizontal sweep in order to obtain a "dupe" negative emulsion position.

The camera is mounted on one end of a five-foot lathe bed isolated from the table with shock absorbers to absorb camera vibrations. The five-inch kinescope is mounted at the other end of bed. Isolation of the tube from vibrations transmitted through the lathe bed is accomplished by means of rubber shims around the deflection yoke and around the neck of the tube. The deflection chassis, the radio-frequency power supply, and the 285-volt regulated power supply are mounted on the lower shelf.

Sound Recording

Recording of the sound portion of a television program is done with standard 16-millimeter sound-on-film recording equipment operating with a synchronous drive at 24 frames per second. A switch operates both the motion picture cameras and the sound recorders simultaneously. Synchronizing marks are momentarily injected into both the sound and video channels by means of a remote switch. The marks are produced by a 420-cycle tone generated by an oscillator. The tone in the video channel produces bars in the kinescope image and an easily identified modulation of the sound track. The picture bars and the track tone are lined up for synchronization of the picture negative with the sound track negative.

Conclusion

A practical method of recording television programs, both sight and sound, has been developed. The recordings can be retransmitted through the television system with acceptable results. As the quality of the television image improves, the quality of recording in kinescope photography will be improved to a degree where the average viewer will be unable to tell if the program he is seeing is "live" or "canned." Certainly, these recordings made from wide-band channels available within the studio plant, unlimited by the restricted band width of the radio-frequency transmitter, will compare favorably with live programs as reproduced on the home television screen.

General References

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DEVELOPMENT AND PERFORMANCE OF TELEVISION CAMERA TUBES*

BY

R. B. JANES, R. E. JOHNSON AND R. S. MOORE

Summary—Three new television camera tubes have resulted from an intensive development program extending over several years. These are (1) the well-known 2P23 image orthicon which is especially suited for remote pickups where a wide range of illumination is encountered and versatility is of the greatest importance; (2) the 5655 image orthicon which is capable of producing pictures of studio quality when the illumination can be controlled; (3) the 5769 image orthicon which may be used for either remote or studio pickups. The construction and operation of these tubes are described in detail. The development of image orthicons is traced by an examination of their limitations and the improvements which have resulted from changes in their construction.

INTRODUCTION

DURING the past fifteen years a number of television camera tubes have been developed. The first to be considered here is the iconoscope[1]. This tube is still used for the transmission of motion picture films and has been extensively used in studio work. When carefully used with the needed complicated correcting circuits, bias and frame lighting, it is capable of producing a high-quality picture. Its resolution is satisfactory and its half-tone response is good. It is also completely stable at all light levels. However, in order to obtain satisfactory pictures, incident light levels of 800 to 1,200 foot-candles are needed on the subject. Even under these conditions "dark spot" and "flare" can be troublesome, particularly for rapid changes of illumination and for scenes that contain dark areas near the bottom of the picture. Although the signal-to-noise ratio may be satisfactory at lower light levels, shading becomes nearly impossible to correct unless the scene is evenly lighted. Lowering the beam current to decrease dark spot is of little help because the signal output drops nearly as rapidly as the dark spot and the tube becomes unusable because of low signal-to-noise ratio.

The type of iconoscope presently available is the 1850-A, which has a diameter of 6¾ inches and a mosaic area of 17 square inches.

* Decimal Classification: R583.6.

[1] V. K. Zworykin and G. A. Morton, TELEVISION, John Wiley and Sons, Inc., New York, N. Y., 1940.

The signal output in microamperes for a typical tube is shown in Figure 1 plotted against illumination on the mosaic in foot-candles for two values of beam current. These two values of 0.15 and 0.2 microampere are, in general, the values used in operation. As the operating personnel become more conscious of dark spot and flare, they tend to use the lower value of beam current. However, some may prefer the greater signal and signal-to-noise ratio obtained from the higher value. With a beam current of 0.2 microampere and a mosaic illumination of 6 foot-candles, a signal-to-noise ratio (assuming 3×10^{-3} microamperes of amplifier noise) of 60 can be obtained. Because this is "peaked-channel noise", i.e., noise of high frequency which consequently appears to the eye as fine grain, it is not objectionable; it is equivalent to a flat-channel noise ratio[2] of 180 to 1. The use of "high peaking" to improve resolution may reduce this ratio to about 100 to 1. It should be noted that these signal-to-noise ratios are expressed as the ratio of highlight signal to root-mean-square noise.

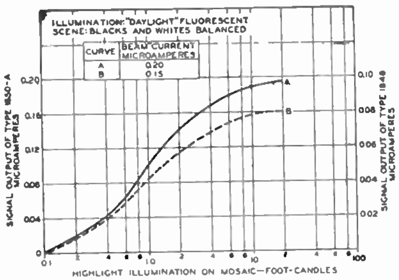

Fig. 1 — Typical signal output characteristics for iconoscopes type 1850-A and type 1848.

Another type of iconoscope which has been manufactured is the 1848. It has a 4½ inch diameter and a mosaic area of a little over 6 square inches. The signal output in microamperes is shown in Figure 1 plotted against illumination on the mosaic in foot-candles for 0.15 and 0.2 microampere beam current. At 6 foot-candles and with the same amplifier noise as for the 1850-A, the value of signal to peaked-channel noise is about 30 to 1. To many users this performance has not been acceptable when compared with that of the 1850-A. The 1848, however, has certain advantages over the 1850-A. When the amount of mosaic illumination is the same, the depth of focus of the 1848 is better because only about ⅓ as much total light is needed to obtain the same mosaic illumination. Not only is the shading of the 1848 usually somewhat easier to handle, but its smaller size makes the final equipment less bulky. Because of its size the 1848 has been used to a certain extent in portable outdoor equipment but here its low sensitivity has not made it popular. In motion picture applications where design factors of size and depth of focus are not important, the

[2] For a comparison of peaked- and flat-channel noise, see O. H. Shade, "Electro-Optical Characteristics of Television Systems", Part I, *RCA Review*, Vol. IX, No. 1, pp. 32-34, March, 1948.

1850-A, because of its greater signal-to-noise ratio, appears to be the logical choice. For studio work, the choice between the 1848 and 1850-A is more difficult.

There have been many proposals for increasing the sensitivity of iconoscopes to make them more useful for studio and perhaps outdoor pickup. One involves the use of an image stage of multiplication.[3] This proposal permits the use of a continuous photocathode instead of a photosensitive mosaic with a possible increase in photosensitivity of 2 or 3. However, in order to keep the tube to a practical size, the photocathode area has to be small, usually about three square inches in size. This size requires the use of a small diameter lens. Although the depth of focus improves, there is little, if any, reduction in the light level needed, because the signal from a pickup tube depends on the total amount of light striking the photosensitive surface rather than upon the illumination per unit area. There is, however, a gain of four or five because of the secondary-emission gain at the mosaic or target. Focusing the electron image from the photocathode to the mosaic is tricky even with magnetic focusing with the result that a loss of resolution and distortion of the picture is likely. Also, care must be taken to prevent interaction between the image-focusing coil and the beam-deflecting coil. Because the gain in sensitivity of such a tube is small unless very high photosensitivities can be obtained, it has not been considered as desirable as the tubes described later in this article.

Another possibility for increasing sensitivity of the iconoscope is the use of signal multiplication. This method involves collecting the secondary emission from the iconoscope mosaic and putting it through a multiplier. This procedure is very difficult with the iconoscope because the large area from which electrons must be collected adds spurious signals. Furthermore, neither image multiplication nor signal multiplication offers any hope of eliminating an inherent fault of the iconoscope—the dark spot.

The next pickup tube to be developed was of the orthicon type[4], the now obsolete 1840. In design and operation this tube was a tremendous departure from iconoscope tradition. Instead of the use of a high-velocity electrostatically focused beam to discharge the mosaic, a low-velocity beam focused by a long magnetic field was used. The vertical deflection is magnetic but in order to avoid the need for high

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[3] H. A. Iams, G. A. Morton, and V. K. Zworykin, "The Image Iconoscope", *Proc. I.R.E.*, Vol. 27, No. 9, pp. 541-547, September, 1939.

[4] A. Rose and H. A. Iams, "The Orthicon", *RCA Review*, Vol. 4, No. 2, pp. 189-199, October, 1939.

power the horizontal deflection is electrostatic. Because the beam strikes the mosaic at a velocity of only one or two volts in the lighted areas and not at all in the dark areas, there is no secondary-electron redistribution and, consequently, no dark spot. Also, the sensitivity of the tube is greater because of more efficient collection of the photo-electrons emitted during storage. A curve of signal output is given in Figure 2. As the curve shows, the signal-to-noise ratio is the same at 1 to 2 foot-candles on the mosaic as at 6 for the iconoscope. In addition, even lower light levels can be used because of the absence of dark spot. Part of this sensitivity is usually used to provide greater depth of focus because of the smaller mosaic size (4 square inches). The tube was very useful in picking up scenes where the incident illumination was only 100 to 200 foot-candles.

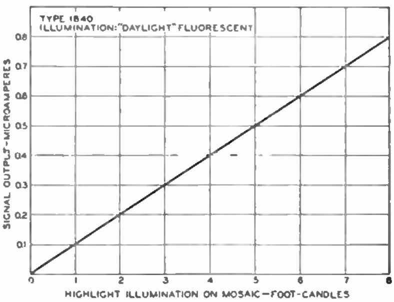


Fig. 2—Typical signal output characteristics for orthicon type 1840.

Although the 1840 has two advantages over the iconoscope: freedom from dark spot and greater sensitivity, it has many disadvantages. The iconoscope is stable at all light levels, but the orthicon will tend to charge up in areas of bright illumination because the mosaic potential is not limited to small values. The "sharpness" of the picture in still scenes approaches that transmitted by the iconoscope, but in moving scenes the tube loses resolution much more readily because, in part, of its much longer storage period. Scenes reproduced by the orthicon, moreover, show a smaller range of intermediate grays than the same scenes reproduced by the iconoscope. The lack of any detail in the low lights is particularly noticeable and is the result of the linear signal output characteristic of the orthicon as compared to the non-linear output characteristic of the iconoscope which saturates at high light levels.

In order for the orthicon to handle the very large signals from local highlights, the grays are pushed down into the noise. It is nearly impossible to transmit any information in dark areas of the picture when other parts are bright. This limitation is a severe disadvantage at baseball or football games when shadows begin to fall across the field. Although the orthicon is useful because it can transmit scenes which the iconoscope cannot, its versatility is severely limited.

Further development work has been done to overcome some of these limitations. Best results were obtained from a tube of the orthicon type which uses all-magnetic scanning and a 5-stage signal multiplier. With all-magnetic scanning any difficulty with "sharpness" of the picture on stationary scenes disappeared, although on moving scenes the same loss of resolution was still apparent. The tube had greater sensitivity than the 1840 because of the use of the signal multiplier, but such factors as instability and the inability to transmit the darker areas in scenes were not improved. The greatest advantage of the tube is that its resolving power for still scenes exceeds that of any other tube. The signal-to-noise ratio was also adequate for light levels of 100 foot-candles but not all of the most severe limitations of orthicons were overcome in this developmental tube.

General Description of the Image Orthicon

The most important development in camera tubes for the past several years has been the now-well-known image orthicon. This tube,

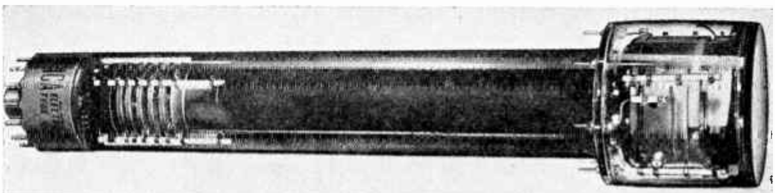


Fig. 3—Typical image orthicon.

on which fundamental work was done by Rose, Law, and Weimer[5], appears to offer the most promise, both for outdoor and studio pickup. The development has led to three commercial types, the 2P23 for poorly lighted, remote pickups, the 5655 for studio work, and the 5769 for general use.

The image orthicon, pictured in Figure 3, combines the features of several of its predecessors. It includes in one envelope an image section, a target or mosaic assembly, low-velocity scanning of the orthicon type, and a 5-stage signal multiplier. Before its performance is described, a brief summary of its construction and operation will be given.

The tube itself consists of a three-inch-diameter image section and a two-inch-diameter scanning and multiplier section. The over-all length is 15½ inches. This size has proven to be a good compromise between camera performance and portability. Although a smaller size

[5] A. Rose, P. K. Weimer and H. B. Law, "The Image Orthicon — A Sensitive Television Pickup Tube", *Proc. I.R.E.*, Vol. 34, No. 7, pp. 424-432, July, 1946.

would make a lighter-weight camera possible, the loss in performance would be objectionable. For convenience, the image orthicon may be described in three parts — one covering the image section, one the scanning section, and one the multiplier section. Figure 4 is a schematic drawing of the tube.

The image section contains a semi-transparent photocathode on the inside of the face plate, a grid (grid No. 6) to provide an electrostatic accelerating field, and a target which consists of a thin glass disc with a fine mesh screen very closely spaced to it on the photocathode side. Focusing is accomplished by means of a magnetic field produced by an external coil, and by varying the photocathode voltage. Light from a scene being televised is picked up by an optical lens system and focused on the photocathode which emits electrons from each illuminated area in proportion to the intensity of the light striking the area. The

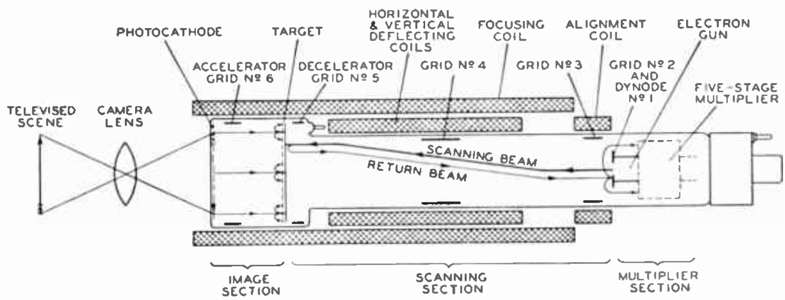


Fig. 4—Schematic arrangement of image orthicon.

streams of electrons are focused on the target by the magnetic and electrostatic fields.

On striking the target, the photoelectrons cause secondary electrons to be emitted from the glass. The secondaries thus emitted are collected by the adjacent mesh screen which is held at a definite potential of 1.5 to 2.5 volts above that of the scanned side of the glass target. The potential of the glass disc, therefore, is limited for all values of light and stable operation is achieved. Emission of the secondary electrons leaves on the photocathode side of the glass a pattern of positive charges which corresponds with the pattern of light from the scene being televised. Because the target is a very thin sheet of partially conducting glass, the charge image is also seen on the scanned side of the target by the scanning beam.

The electrons are emitted from the gun through a small defining aperture and are focused into a fine beam by means of the magnetic field of an external focusing coil and the electrostatic field of grid No. 4. Magnetic deflection is used to scan the target. Grid No. 5 serves to adjust the shape of the decelerating field between grid No. 4 and the target in order to obtain uniform landing of the scanning beam over the entire target area. The electrons stop their forward motion at the surface of the glass and are turned back, except when they approach the positively charged portions of the pattern and are deposited on the glass. This deposition leaves the glass with a negative charge on the scanned side and a positive charge on the photocathode side. These charges will neutralize each other by conductivity through the glass in less than the time of one frame.

The electrons turned back at the target form the return beam which has been amplitude modulated by absorption of electrons at the target in accordance with the charge pattern. The returning modulated beam strikes the first dynode which as a result, emits secondary electrons. These secondaries in turn are drawn down through a series of multipliers of high secondary emission which increase the signal a 1000 fold. The increased signal is finally collected at the anode of the multiplier and fed to a pre-amplifier through a resistance of the order of 10,000 to 30,000 ohms.

As this brief summary shows, the image orthicon has all the advantages of the orthicon together with the added advantages of greater sensitivity due to the image section, and of greater stability due to the mesh screen near the target. Its sensitivity is about 100 times greater than that of the iconoscope and it is stable over a light range of several hundred to one. These features make it very versatile and especially useful for outdoor scenes. Without any change in adjustment, the tube can handle a high light scene and then be used for a scene in deep shadow. It is also far superior to the orthicon in reproducing scenes containing both high lights and shadows.

This gain in sensitivity, particularly for the earlier image orthicons, was achieved only with a loss of signal-to-noise ratio and useful resolution. Also, the half-tone response differs from that of the orthicon. Why the tube possesses these properties can probably be best explained by a detailed examination of its construction and a description of how each section of the tube contributes to these properties. Image orthicons now available include the 2P23, 5655 and 5769. Because the 2P23 is the oldest and most widely used, it will be described first.

Construction and Operation of the 2P23

Photocathode

The photocathode of the 2P23 consists of a semi-transparent layer of the cesium-silver-oxide type. As in the 1840 orthicon, the layer must be semi-transparent because the light strikes it from one side and photoelectrons are emitted from the other. However, the 2P23 has the advantage over the 1840 in that this layer can be continuous. Higher sensitivities are possible with such a layer, the sensitivity to incandescent light being of the order of 10 to 20 microamperes per lumen compared to about 3 for the 1840. (This figure for the 1840 includes the loss in the semi-transparent signal plate needed in that tube.) It has proven difficult, however, to obtain a reproducible color response from tube to tube in the 2P23. Figure 5 shows the spectral response characteristics for typical low and high sensitivity tubes. Individual tubes may have spectral response characteristics anywhere between these two extremes. The high infrared sensitivity of some 2P23 tubes leads to very good sensitivity when low-temperature incandescent lighting is used but also leads to peculiar renditions of certain colors or objects. In outdoor use, for example, green grass appears to be white or very light gray. A fairly good color response with a two-fold or more loss of sensitivity can be obtained by the use of fluorescent lighting or by proper filters with incandescent light. Care must also be taken in the processing of the tube to keep the conductivity of the photosurface fairly high. Otherwise, for high light scenes, the picture will be distorted geometrically because of the voltage drop in the photosurface.

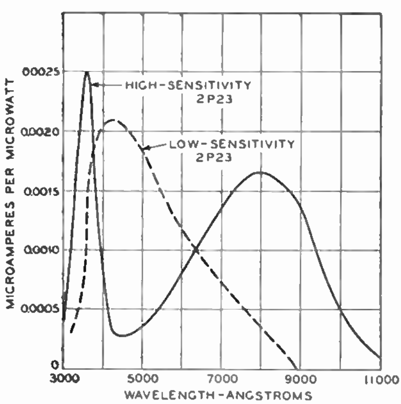


Fig. 5—Approximate spectral sensitivity characteristics of image orthicon type 2P23.

The size of the target limits the size of the picture on the photocathode to a rectangle with a diagonal of about 1.6 inches. This size is much smaller than that of any of the other pickup tubes so far discussed and means that available short-focal-length lenses having a rather small diameter may be used. Part of the increased photosensitivity, therefore, will have to be used to obtain greater depth of focus. Because of the high sensitivity of the tube the exchange of depth of focus for sensitivity is not a disadvantage. On the other hand, the use of small lenses makes it much easier to use a revolving turret containing three or four lenses of different focal length. The use of such a turret has become universal in the latest camera design.

Focusing of electron image

The use of a magnetic field to focus the emitted photoelectrons onto the target gives uniform resolution with little distortion. In general, this resolution is much higher than that of other sections of the tube. When a magnetic field of about 75 gauss and a photocathode voltage of about —400 volts are used, the electrons will make one "loop" in going from the photocathode to the target. Only a small improvement in resolution is possible at higher fields and voltages, but a serious deterioration of the resolution and signal occurs with a photocathode voltage below —200 volts. The voltage on grid No. 6, a cylindrical-type grid nearest to the photocathode, is adjusted to minimize picture distortion (in particular, the so-called "S" distortion) and to improve corner resolution. Best results are obtained when the voltage of this grid is about 80 per cent of the photocathode voltage. A lower grid No. 6 voltage produces "S" distortion in one direction and a higher voltage, "S" distortion in the opposite direction. With the focusing coil usually used with the 2P23 there is a small reduction in the image size at the target. Because the image section is near the end of the focus coil where the magnetic field is flaring, the image size at the target has a diagonal of about 1.4 inches for a 1.6 inch diagonal on the photocathode.

Image section crosstalk

Although the image section is not a serious limit to the resolution of the 2P23 directly, the resolution in this section can be seriously impaired by leakage of the strong magnetic deflection fields from the scanning yoke into the image section. The leakage fields cause a vibration of the image electrons from the photocathode around their normal path during the 1/30-second storage time on the target and, as a result, blur the charge image. Because the storage time decreases with increased illumination, this "crosstalk" effect on resolution is more serious at low light levels on the photocathode. Since the tube is generally used so that the high lights are just out of the storage range, "crosstalk" control is a serious problem. Although the effect may be reduced within limits by going to high magnetic fields and photocathode voltages, more than 75 gauss is not practical in portable equipment because of the added scanning power needed. Other approaches to the problem offer better practical solutions.

Magnetic shielding to reduce crosstalk

A number of methods will reduce "crosstalk" difficulty, all of which, in one way or another, involve the shielding of the image section from the stray magnetic fields of the deflecting coils. One method which has worked out quite well is to wrap the outside of the external focusing coil with some magnetic shielding material such as silicon steel or Mu metal. In order to avoid high absorption of scanning power, the shielding material should be thin strips and wound in several layers separated by insulating material. For the same reason, the strips making up each layer should not be too wide. The use of such a winding tends to pull the stray flux lines away from the photocathode into the shielding material. Resolution gains of as much as 200 lines have been achieved by this method.

Loss of resolution due to initial velocities of emission

There is another cause that can impair the resolution in the image section: the initial emission velocity of the photoelectrons from the photocathode. This problem has been analyzed in a paper by H. B. DeVore[6]. The loss of resolution is most noticeable when blue light is used to illuminate the scene being televised and the photocathode of the camera tube has red and infrared response. Under such conditions the initial emission velocities of the photoelectrons are appreciable and can visibly limit the resolution. Because many 2P23's have high red and infrared response this limitation applies chiefly to this tube. The photocathodes used in the 5655 and the 5769 on the other hand, have little red and no infrared response, so the loss of resolution is not appreciable. In scenes where there is considerable "blue" light such as in skylight, the resolution of the 2P23 will be inferior.

Target mesh structures

When light causes photoelectrons to be emitted from the photocathode, the image section focuses these photoelectrons onto the target mesh assembly. This assembly is truly the heart of the tube and is the main reason for its amazing performance. This type of target differs from any which have been used before in commercial pickup tubes in that the signal is impressed on one side and taken off the other. Such a structure is called a two-sided target or mosaic.

There have been many attempts in the past to fabricate a successful two-sided target. The patent literature is evidence of many types, most of which have proven to be too difficult to manufacture. The use of two-sided mosaics was first attempted in connection with iconoscopes. An image section was also used and the mosaic or target consisted of an insulated (generally enameled) wire mesh in which

[6] Henry B. DeVore, "Limiting Resolution in an Image Orthicon Type Tube", *Proc. I.R.E.*, Vol. 36, No. 3, pp. 335-346, March, 1948.

the openings were closed with metal plugs. Such tubes actually operated, but the difficulty of making a target free of blemishes such as pinholes and surface irregularities proved almost insurmountable. If low-velocity scanning is used, such a target would also have too high a capacitance and lead to bad lag effects under low lighting conditions. The two-sided problem was solved through an entirely different approach*. This consisted of a very thin glass membrane of controlled resistance together with a fine mesh screen mounted close to it on the image or photocathode side. Such a glass membrane can readily be made free of pinholes and bad surface imperfections. Many problems, however, had to be solved in order to obtain a practical assembly that was free enough from other imperfections to be used in commercial television.

Target resistivity

The operation of the tube makes several demands on the glass target. The photoelectrons striking the image side of the glass through the mesh openings give rise to secondary electrons that are collected by the mesh. At low light levels the potential of the glass target does not ordinarily reach that of the mesh screen during the 1/30-second storage time. In order to obtain good resolution for such a storage time, the lateral leakage or leakage between elements is kept as low as possible, by the use of a high-resistance glass and by the use of a target as thin as possible. When the scanning beam approaches the scanned side of an element that has been charged, the beam "sees" the same potential as that of the charged side because of the thinness of the target. Electrons are deposited from the beam on the target until the potential returns to nearly the equilibrium potential under the beam when no light is present. After the beam leaves the element, the positive charge on the image side and the negative charge on the scanned side must combine in less than a frame time or 1/30th second. Otherwise, a "sticking picture" which will be of opposite polarity from the original picture will be seen if the picture is moved. If the picture is stationary the signal output from lighted areas will decrease. If the glass resistivity is very high the signal output for a stationary picture will, after a few scans, fall nearly to zero. If the resistivity is only slightly too high, the sticking picture will disappear as the tube warms up in the camera because the resistivity of glass falls about 2 or 3 to 1 for each 10-degree-centigrade rise in temperature. A satisfactory upper limit for the glass resistivity has been found to be about 9×10^{11} ohms per centimeter measured at 20 degrees

* See reference (5).

centigrade temperature. For the 2P23 a temperature of about 35 degrees centigrade at the target is needed to eliminate picture sticking entirely, when the glass has this value of resistivity.

The "sticking picture" puts a top limit to the resistivity of the glass. Mechanical handling puts a lower limit of about 0.0001 inch on the thinness of the target. A lower limit for the glass resistivity, necessary to keep lateral leakage low, is approximately 3×10^{11} ohms per centimeter measured at 20 degrees centigrade. With such a resistivity the tube can be operated at a temperature of about 65 degrees centigrade before serious loss of resolution sets in. At higher temperatures the resistivity will fall below 10^{10} ohms per centimeter, and a serious loss of resolution will occur particularly at low lights.

Target surface effects

Besides the mechanical difficulties of making a glass target of sufficient thinness and of the necessary strength, many other problems arise during the processing of the tube. In order to obtain sensitivity, the image side should have as high a ratio of secondary-electron emission as possible. Enough cesium reaches the image side of the target during processing of the photocathodes to give a ratio of 4 or 5 to 1. At times, however, too much cesium gets on the image side, the lateral leakage falls, and the target becomes "leaky", i.e., the resolution of the picture is poor. Such leakage, of course, shows up first at low light levels, where storage is complete over a frame time. During operation or shelf life, it sometimes happens that enough cesium will migrate to the target to cause "target leakage". This trouble can be minimized by operating the tube at the lowest possible temperature. (A minimum lower limit would, of course, be 35 degrees centigrade because of the target resistivity problem.)

Target contact-potential effects

There are other changes which occur in the glass that are largely the same in all tubes. During operation the scanned area on the beam side changes slowly in contact potential with respect to the unscanned area. With uniform light on the photocathode, overscanning will show the previously scanned area to be, in general, darker than its surroundings because the contact potential of the scanned area changes with respect to the thermionic cathode. When the tube is placed in operation the maximum area of the target which makes the picture magnification smallest should be used, because later it will be impossible to increase the area used because of the white edges that will show in the picture.

In addition to the "sticking picture" that can occur when the tube target is too cold, another type of sticking can develop. If the camera is stationary and the tube "looks at" a strongly lighted area for a period of time (5 minutes or more), a sticking picture of the same or most often of opposite polarity can develop. This trouble is more likely to occur when the tube is cold, but it will develop in all tubes if the picture is left stationary too long, regardless of temperature. Unlike resistivity sticking, the sticking or "burning-in" disappears slowly. This type of sticking can be avoided only by care in handling the camera. In order to remove a sticking picture once it has been developed, point the camera at a flat lighted scene and operate it for several hours.

Target secondary emission

One last point should be mentioned before leaving the rather complicated glass target. Whereas the image side should have high secondary emission to produce as much signal as possible, the scanned side should have low secondary emission. Although the scanning beam approaches the target at nearly zero velocity and is turned back in the dark areas, it strikes in the lighted areas with a velocity corresponding to one or two electron volts. Even at such potentials, some secondary emission will occur. Any such emission increases the number of electrons returning to the multiplier from the lighted areas and thus increases the noise, because more beam current is needed to discharge lighted areas. The secondary emission of the scanned side is reduced by evaporating on it a very light layer of some metal such as silver which has low secondary emission. The reduction in the amount of noise is readily noted after such an evaporation. During operation there is generally a slow increase in the secondary emission so that after several hundred hours of operation the picture becomes more noisy. A re-evaporation of silver at this time by the manufacturer will once more reduce the secondary emission.

Loss of scanning

It is important to note what damage can be done to the target by a stationary beam. With light on the photocathode and a sharply defined beam, a hole can actually be started in the target. If the beam is defocused the bombarded area may become either darker or lighter than the rest of the target. Removing the multiplier voltage will not, of course, be of any help because the beam still will strike the target. Removal of the photocathode voltage is only a partial solution because light can pass through the photocathode and strike the target, which being nearly always somewhat photosensitive, will charge up and allow the beam electrons to land. The only positive method of preventing damage in case of scanning failure is to bias off the beam or target. This precaution should always be taken when the equipment is not being monitored.

Target-to-mesh spacing — close spacing

For proper functioning of the glass target, a fine mesh screen the potential of which can be varied is needed on the image side. This mesh serves a two-fold purpose—as an element to increase the capacitance of the target, and as a limiter to prevent the target from charging to high potentials such as occurs in the 1840 orthicon. The maximum amount of charge that can be deposited on an element of the target depends on the capacitance of the element and the potential of the mesh above the potential of the scanned side of the target. This

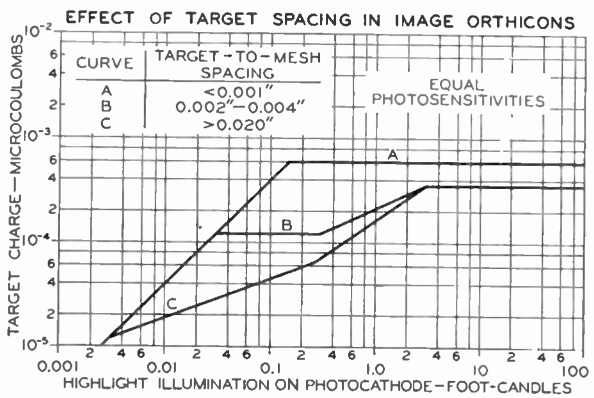


Fig. 6 — Effect of target-mesh spacing in image orthicons.

maximum charge also determines the maximum signal-to-noise ratio of the tube. At low light levels the target element never reaches mesh potential so that the charge rises linearly with light. In this region the tube has the same characteristic as a regular orthicon, namely, a loss of resolution for scenes in motion. For a spacing of the mesh to the target much less than the diameter of a picture element (for a 500-line picture this diameter is somewhat less than 0.002 inch), the charge rises linearly with light until it becomes limited by the capacitance and mesh voltage. When this limit is reached the charge caused by the high lights cannot increase, although that due to the low lights continues to rise. Curve A, Figure 6 shows an extremely simplified curve of the charge developed on the target during the time for a complete picture frame plotted against the illumination on the photocathode for a target-to-mesh spacing of less than 0.001 inch. Ordinarily, it would be assumed that the picture contrast would decrease rapidly after the high light signal becomes constant. Redistribution of secondary electrons from the bright areas onto the lowlight areas, however, tends to preserve the contrast even when the high lights are well above the knee. If an intense small area of light, such as a direct reflection of the sun from the windshield of a car, is present in the picture, this redistribution gives rise to a disturbing black border around the light area. The image orthicon cannot faithfully transmit such a scene. Also, in pictures with large black and white areas the contrast is preserved only at the area boundaries when the high lights are well above the knee. In such pictures the blacks appear gray although the resolution and "snap" may appear to improve because of the outlining of the edges.

The redistribution of secondary electrons makes a valuable contribution to the resolution of moving objects. As the picture moves, the border around the high lights discharges the high light signal at its former position in less than a frame time so that only the latest image is seen when the picture is scanned. To obtain the best picture contrast and the most natural-appearing picture (that is, with blacks "black" instead of gray) the high lights should be run just at the knee of the curve. To take advantage of the better resolution in motion some loss of contrast is usually taken by operation somewhat above the knee.

Target-to-mesh spacing — wide spacing

If the spacing between the mesh and the target is much greater than the diameter of an element, the charge curve becomes more complicated. In the case of the close-spaced target, the capacitance of an element to the mesh greatly exceeds the "free space" capacitance of the element itself. For a wide-spaced target the "free space" capacitance is larger than its capacitance to the mesh. For low lights the charge will rise linearly with light until the point is reached where it is limited by the product of the capacitance of the element to the mesh and the mesh voltage. This point, as curve C of Figure 6 shows, is lower than the equivalent point for a close-space target because of the much smaller mesh-to-target capacitance of each element. However, as the light is increased above this point, the charge of the wide space target can increase because of the free-space capacitance of each element. This increase is at a slower rate because the discharge of the first part of each lighted section also partially discharges sections beyond it. Because the "free-space" capacitance is largely between neighboring elements, the edges of a lighted section both horizontally and vertically will produce a greater signal because of the partial discharge of nearby elements. As the light is increased, these partially discharged elements are more and more recharged before the beam reaches them. Finally, a point is reached where the free-space capacitance of an element and its mesh voltage limits further increase. As curve C of Figure 6 shows, this point occurs at a light level a hundred times above that needed to reach the first knee of the curve. The effect in the transmitted picture of the free space or interelement capacitance is that the blacks are outlined with a white edge at the transition to white, occurring at the right end and the bottom of the blacks when the beam is scanning from black to white or gray.

Experience has shown that the proper target-to-mesh spacing depends on what the tube is expected to do. Very wide spacings, where the interelement capacitance is the only one that needs to be considered, has the advantage of taking the mesh completely out of focus. However, because the signal-to-noise ratio is too small to be useful except at very high light levels and because the resulting white edges are very annoying, the useful range is limited to intermediate and close spacings. The best spacing is determined by actual tests.

Tests of target-to-mesh spacing

Tubes with a wide variety of target-to-mesh spacings have been tested. For very wide spacings in the order of 0.020 inch to 0.080 inch the signal is far too low and the white edges are annoying. Spacings of the order of 0.004 inch to 0.008 inch produce useful tubes, but even in this range the signal-to-noise ratio is borderline and the white edges are still troublesome. A range of 0.002 inch to 0.004 inch was finally selected as the best for the 2P23 and the 5769. The white edges have largely disappeared and the signal-to-noise ratio is improved. Such tubes are of "intermediate" spacing, since the spacing is only slightly larger than an element diameter. Nearly all the properties are those of close-spaced targets as curve B, Figure 6 shows; only at high lights is there an increase in charge due to "wide-spaced" characteristics. In general applications this increase is not evident and can largely be overlooked. The properties of tubes with "close spacings" will be discussed in connection with the 5655.

Target potential

With regard to the mesh, two further points are important. As we have seen, one of the items limiting the target charge is the capacitance of an element, which is determined by the spacing of the target to the mesh. For the 2P23 and 5769 the spacing is in the range of 0.002 inch to 0.004 inch. The target charge is also determined by the externally applied mesh potential. Experience has shown that the optimum potential depends to a great extent on the picture content. For scenes with flat, even lighting, especially when the scene illumination remains constant, the mesh potential can be set in the range of 2 to 2.5 volts above the point where the picture is cut off. (In actual experience, this cutoff point is at —1.0 to —2.0 volts with respect to the thermionic cathode.) Above 2.5 volts the picture will be stable but may have a peculiar "differentiated" appearance. When the potential difference between the target and the mesh is more than 2.5 or 3 volts, the beam will be bent by the more positive areas of the high lights and a premature discharge of these areas will occur. This action is known as beam bending and gives a picture reproduction inconsistent with the charge distribution on the target. When the tube has to pick up a scene with high lights and then one that is largely in the shadows, a mesh potential range of 1.5 to 2.0 volts is preferable. In general, the mesh potential should be kept as high as possible, without endangering picture fidelity in order to obtain the best signal-to-noise ratio.

Development of a suitable mesh

The problem of a satisfactory mesh has been the subject of a long development program. The first experimental image orthicon tubes were made with woven screens. Attempts to obtain woven screens that had uniformly spaced wires proved unsuccessful. In addition, the finest weave that could be obtained with a usable opening was only 325 to the inch which severely limited the resolution because the size of the picture on the target has a diagonal of only 1.4 inches. Various methods of electroplating mesh have been tried in the past, but, although they are capable of producing mesh of up to 400 lines per inch, the open area is small, generally not over 25 per cent for the finer meshes. A new method of producing an electroplated mesh, was originated at RCA Laboratories[7], and has been developed to a degree that meshes of 500 openings to the inch with 50 to 65 per cent open area can be produced. Special methods of mounting and tightening the mesh, which is only a few tenths of a mil thick, were worked out. Because the mesh is so close to the target, it is nearly at the point of focus of the image electrons. If a wide-band amplifier is used, this mesh can be seen by looking carefully at the kinescope picture. With the standard television bandwidth it is just noticeable. However, a problem of a "beat" pattern does arise. Because the picture height on

[7] H. B. Law, "A Technique for the Making and Mounting of Fine Mesh Screens", *Rev. Sci. Instr.*, Vol. 19, No. 12, December, 1948.

the target is approximately an inch, there will be about 500 wires of mesh to a picture height. These wires can beat with the 525-line scanning frequency to produce a low-frequency beat pattern. In the construction of the tube this possibility is reduced to a minimum by mounting the mesh at a 45-degree angle to the scanning beam. Even in this case some regions of the picture may show beat patterns, particularly in highly lighted areas. These patterns, also can be eliminated at a sacrifice in resolution, by slightly defocusing the beam. In operation the beat pattern can be minimized by scanning as large a part of the target as possible so as to keep the picture height at a maximum.

Electron gun

The gun which produces the beam consists of a thermionic cathode which is held at ground potential, a control grid (grid No. 1) and an accelerating grid (grid No. 2). Grid No. 2 contains a small aperture about 0.002 inch in diameter which serves to define the beam. After emerging from this aperture, the beam with a velocity corresponding to about 300 volts, passes through the grid No. 3 region which is also at 300 volts. It then emerges into the focusing section which consists of a uniform electric field of about 200 volts and a magnetic field in the direction of the beam of 75 gauss. Any component of the electron beam which has only forward velocity will go straight down the magnetic field. Other components which have radial velocities will form loops around the magnetic field and return to a disc of focus at the end of each loop. By means of the fields previously mentioned the beam is focused on the target at the end of the 5th loop. When the beam passes through the focusing section, it is also deflected in vertical and horizontal directions by means of magnetic fields at right angles to the focusing field.

Landing of beam at target[8]

As the beam approaches the target it is slowed down. If no light is on the target the electrons in the beam will continue to land until the target potential drops to a value determined by the initial velocities of the thermionic electrons and the contact-potential difference between the thermionic cathode and the target. When this potential is reached, all of the beam will be turned back unless the tube is illuminated. In the lighted areas all the electrons of an ideal beam would land on the

[8] P. K. Weimer and A. Rose, "The Motion of Electrons Subject to Forces Transverse to a Uniform Magnetic Field", *Proc. I.R.E.*, Vol. 35, No. 11, p. 1273, November, 1947.

target until the target is driven back to equilibrium potential. This ideal condition, however, is not reached for several reasons. After the beam emerges from the aperture, its direction, as a whole, may not be along the magnetic field. In this case the radial velocity of the beam may be so high that none of it can land at the target under normal conditions. Nothing will be visible in the kinescope picture except noise, unless the mesh voltage is raised to a very high value. This condition is corrected by the use of an alignment coil which produces a magnetic field at right angles to the beam direction near the aperture. This field can be rotated and varied in intensity until the direction of the beam is along the magnetic field. Even after best alignment, however, all of the electrons will not land because of variations in the initial emission velocities which range from 0 to about 0.5 volt. For low light levels only those electrons with the highest initial velocity will land and the percentage landing or the "beam modulation" will be poor. However, even for higher light levels only a portion of the available electrons will land because of the radial velocities introduced by the gun and because of the secondary emission which occurs at the target. Some preliminary data indicate that even for a target on which silver is evaporated the secondary-emission ratio may be as high as 0.5 in the high lights. All of these effects combine to lower the possible signal-to-noise ratio since a larger beam current is needed to discharge the target. The failure to land at low lights can give rise to another condition which is described by the term "picture lag". So few electrons land that a picture is not discharged by the beam in 1/30 second. If the picture is moved there will be a trail behind it of the same polarity. This condition is not serious in the 2P23 except at very low light levels because of the very small target capacitance. The condition is more serious, however, in the 5655.

Edge landing

Besides the lack of 100 per cent landing in the center of the target, a problem of poor landing at the edges arises because of the radial velocities introduced by the deflecting field. In a transmitted picture with poor edge landing the signal will be the highest in the center and drop off progressively towards the edges. The radial velocities introduced by the deflection are largely counterbalanced by the use of a decelerator grid (grid No. 5) which is in the form of a short cylinder. This grid, when operated at a positive voltage between that of the target and the focusing grid (grid No. 4) produces a radial electrostatic field which is zero at the center of the picture and increases toward the edges. This field gives the electron beam a radial velocity opposite to that produced by the deflecting field.

Landing can also be influenced by many other factors. In the design of the tube it has been found necessary to control accurately the shape of the glass near the decelerating region. Otherwise, the deflected beam tends to strike the glass in this region and not reach the target. Also, the deflecting coil, image socket, and focusing coil must be carefully designed. Best results have been obtained with a deflecting coil 5 inches long. With a longer or shorter coil, the flare fields are different and cannot be counterbalanced as readily with a simple grid. For the same reason the three-inch image section of the tube should be as close to the end of the deflecting coil as possible. This space is limited to 0.5 inch by the length of the image leads and the socket thickness. Slightly better results can be obtained with a spacing of 0.3 inch. The effect on landing of any shielding windings on either the deflecting or focusing coil must also be considered. It is general practice to wrap the deflecting coil with a layer of iron wire, to help prevent leakage of the deflecting field into the target lead which generally returns over the deflecting coil from the image section to the rear of the focusing coil. However, this winding has a slight deteriorating effect on the landing because of its modications of the flare field. When image focus was studied, it was found necessary to reduce the "crosstalk" from the deflecting field into the image section. As mentioned previously, "crosstalk" can be reduced by the use of an external shield over the focusing coil. Such shields, because they modify the flaring of the deflecting field, can also affect the landing. The shields, therefore, should be designed with this item in mind. In fact, proper arrangement of shields will actually lead to better landing than can be obtained with no shields.

In general, the beam imposes no severe limitation on the center resolution of the 2P23 in its present state of development if a field of approximately 75 gauss and a focusing voltage of about 200 volts are used. However, the corner resolution is somewhat deteriorated when the center is in best focus. This deterioration occurs chiefly in the beam section rather than the image section. Some improvement can be obtained by proper adjustment of grid No. 5 provided the landing is not seriously affected. The resolution limits of the beam will be considered more thoroughly in the discussion of the 5655.

Return beam

The portion of the beam that does not land on lighted sections of the target returns and strikes the accelerating grid (grid No. 2) which also serves as the first dynode of the signal multiplier. The amount of deflection received in going to the target is not quite balanced by that received in returning so that the beam scans a small area of the first dynode. The size of this scan is roughly ¼ inch. This ¼-inch scan poses quite a serious problem in keeping the dynode free from spots because it is magnified by 25 or more in the kinescope picture. Fortunately, the dynode is not quite in focus for best focus at the target. In operation, however, it is usually necessary to defocus the picture slightly in order to minimize the spots, especially for dark scenes. These spots are nearly always white, indicating a lower secondary emission. In general, for dark scenes these spots are the most severe limitation on resolution, while for well-lighted scenes the target mesh is the limiting factor. Several methods have been tried for reducing the dynode spots including the use of highly polished surfaces and uniformly roughened ones for the dynode both being coated finally with an evaporated film of a metal with a high secondary emission. No completely successful solution, however, has as yet been found. In any event, the spot due to the aperture opening is always present.

Signal multiplier

The first orthicons and image orthicons were generally made with only one stage of signal multiplication. The signal in the form of secondary emission from this stage was collected by a nearby electrode. In general, the gain from this single stage was found to be insufficient. The purpose of signal multiplication by secondary emission is to obtain a nearly noiseless multiplication of the small signal which modulates the return beam to a level well above the noise of the first stage of the video amplifier so that amplifier noise is no longer a limitation. As mentioned previously, the maximum charge that a target element can have is limited by the product of its capacitance and the mesh voltage swing. For the 2P23 with a mesh voltage of 2 volts, the total charge for the whole target, if it is entirely highlighted, is about 1.6×10^{-10} coulombs. Because this capacitance is discharged in 1/30 second, the calculated signal current at the target is only about 5.0×10^{-9} amperes from the highlights. If the first video amplifier is connected directly to the target, a root-mean-square noise current of 3×10^{-9} amperes may be assumed. It can be seen that the highlight signal-to-noise ratio is less than 2 to 1 with the amplifier the limiting item. The gain of the multiplier should at least be such that the beam noise is the limiting item. This noise which is, of course, due to shot effect of the temperature-limited thermionic emission, is given by the expression $(2ei\Delta f)^{1/2}$. In this expression, e is the charge of an electron (1.59×10^{-19} coulombs), i is the beam current in amperes, and Δf is the frequency bandwidth of the picture in cycles per second. For a bandwidth of 4.25 megacycles and if it is assumed that all of the beam is useful in discharging the picture, the beam noise is 0.08×10^{-9} amperes. A gain of at least 40 is needed to bring this value up to the level of the amplifier noise. If the tube is to be used to pick up lower light scenes with no highlights so that the beam current can be reduced, a higher gain is useful. This condition is unusual in the field. Another reason for higher gains, however, is to reduce the number of electron tube amplifier stages required. For this purpose, a gain of several hundred is useful.

In order to obtain a gain of several hundred several multiplier stages are needed because the average gain per stage is usually only about 4. Many multiplier designs have been suggested and tried. Because the second stage must collect all of the electrons from an appreciable area of the first stage (because of the scanning of this stage), it has been found advisable to use a symmetrical multiplier. Such a multiplier can be a series of screens set one below the other around the gun and first dynode. The principal problem has been to get all the secondaries from the first stage over to the second stage. The best solution has proven to be the use of an extra cylindrical grid (grid No. 3) above the first dynode. The voltage of this grid is generally set at or slightly below the first-dynode potential. The beam, which emerges from the aperture in the first dynode with a 300-volt velocity, passes through an aperture in the top of grid No. 3 into the grid No. 4 section without being affected. However, the slow-velocity secondaries emitted by the return beam find themselves in a region of uniform potential except for the second dynode and will be attracted to it. This second dynode is parallel to the first but generally slightly below it. The magnetic field should also be weak near the first dynode to prevent the secondaries from spiraling about it and eventually returning to the first dynode. Any failure to collect all the secondaries at the second dynode will lead to a picture that is darker in one section than another. As in the iconoscope, this condition shows up most clearly when there is no signal due to light so it also is referred to as "shading". This shading signal differs from iconoscope shading in that it is smaller than the picture signal and can usually be cancelled by the insertion of a simple horizontal-sawtooth component in the amplifier. It does not vary greatly with illumination as does the iconoscope shading but depends almost entirely on the beam current. Consequently, the shading control requires little readjustment once it has been set.

The second dynode is an efficient multiplier consisting of a 32-blade pinwheel with the blades set at an angle of about 30 degrees to the plane of the multiplier. Primary electrons which strike the blades emit secondaries which are drawn through the slots to the next stage. A high-transmission screen is mounted on top of the pinwheel to prevent the secondaries from being pushed back into the surface by the lower voltage of the preceeding stage. Such multipliers will give gains of about 4 per stage at a primary electron velocity of 300 volts. A 5-stage multiplier as used in the 2P23 will give gains of about 1000. The signal output which will be discussed later is about 10 microamperes. This value is considerably higher than the 0.018 microampere output of the 1850-A iconoscope. The gain of the external amplifier can thus be reduced by a value of about 500 permitting the elimination of at least two amplifier stages.

Generally, the recommended voltage per stage of the multiplier is between 200 and 300 volts except for the first dynode which is held at 300 volts. The over-all voltage required by the multiplier stages will then be in the range of 1100 to 1500 volts.

Operation of 2P23

Although the 2P23 has proved very successful for field use because of its high sensitivity, wide light range, and the relative freedom from shading, it has several limitations. Because of the low capacitance of the mosaic, the maximum charge any element can attain is small. This limitation, and the fact that all the beam that approaches a lighted area of the target does not land, limits the signal-to-noise ratio to a relatively low value so that the picture appears somewht "noisy". The type of photocathode used, although it has a high over-all response does not faithfully reproduce colors in black and white. This poor color response is particularly troublesome in studio work. Also, the "white edge" effect gives a somewhat unnatural looking picture which, although it is not too serious for outside pickup, is very noticeable on high-quality studio scenes.

The effect of beam modulation on the signal-to-noise ratio and its improvement by a change in gun design will be considered first because the results are applicable to both the 2P23 and 5655. In low lighted sections of the target it has been pointed out that the percentage of the beam that lands is low because of the spread in initial velocities at the thermionic cathode. When light is not present, the target is driven to a voltage corresponding to the highest initial velocities. For small illuminations there is only a small percentage of the total beam which can discharge the target. At high lights this limitation is no longer present and it would be expected that the initial velocities would not be a problem. However, with the original gun design of the 2P23 the percentage of the beam that landed in the highlights was only about 15 per cent. For a maximum signal at the target of 0.005 microampere the beam current needed is 0.033 microampere. The beam noise given by the expression $(2e\,i\,\Delta f)^{1/2}$ is 0.0002 microampere. The maximum signal-to-noise ratio is, therefore, about 25 to 1. The type of noise, because it is due to the beam, is not "peaked channel noise" but is spread over the entire bandwidth so that long noise pulses in the form of streaks come through. This ratio is less favorable than the 60-to-1 ratio of the 1850-A iconoscope with a "peaked channel noise" in which the noise pulses are all short and appear only as small dots.

Development of an improved gun

The original gun of the 2P23 consisted of a thermionic cathode and a control grid with a large aperture spaced at a rather large distance from the cathode. The spacing between the control grid and the accelerating grid (grid No. 2) was also high. This gun could readily be manufactured but the beam current was drawn from a rather large area of the thermionic cathode. Because only the center portion of this beam passes through the small aperture in grid No. 2, it was expected that only the part of the beam that is emitted by the center of the cathode would be used. Extensive study, however, has shown that the center of the beam is not all that is used. The electrons drawn from parts away from the center of the cathode and which would be expected to have a radial component of velocity enter the beam and raise its radial component.

A new gun design has been developed to draw electrons, as far as possible, only from the center of the cathode area. The grid No. 1 aperture was made smaller, the grid No. 1-to-cathode spacing was made as close as possible and an extra accelerating aperture which is also at grid No. 2 potential was added close to the grid No. 1 aperture. The result of these changes is that the cathode is much more heavily loaded; i.e., for a given beam current a much higher percentage of the electrons are drawn from the center of the cathode. Because these electrons have a smaller radial velocity, most of their energy is in the forward direction and they will be more likely to land at the target. When the new gun is used, the percentage of modulation rises to about 30; the signal-to-noise ratio is increased to about 35 to 1. Although this value is still on the low side, it is acceptable for outside pickups. It is interesting to note that the signal-to-noise ratio of good 35-millimeter film is also 35 to 1.

In addition to the improvement in signal-to-noise ratio, this new gun provides an improvement in resolution due probably to the smaller radial velocities which make the final spot smaller. (The presence of radial velocities cause a point source to be imaged as a spot of measurable magnitude at the target.) With the new gun, the beam itself is at present not a limitation on the usable resolution of the tube unless the magnetic field and grid No. 4 focusing voltages are too low. In general, a field of 75 gauss and a grid No. 4 voltage of 200 volts is a good compromise between good resolution and the need for excessive scanning power. With such operating conditions the beam is able to resolve better than 1000 lines in the roughly one-inch vertical height of the picture, or more than 1000 lines per inch.

Whether the 30 per cent modulation efficiency represents a limit to what can be done with gun design is not known. Initial velocities of electron emission at the thermionic cathode, of course, even at highlights present some limitation. Measurements at a variety of cathode temperatures have shown no definite improvement. The main limitation may be the secondary emission and electron reflection that comes at the target.

The new gun was originally developed in connection with the new 5655 studio image orthicon. However, it is equally useful in the 2P23 and has been adopted there. The 5655 now differs from the 2P23 only in the target to mesh spacing and in the photocathode surface. Why these changes give a tube (the 5655) which is superior for studio use but which is not so versatile for remote pickup will be discussed next.

Use of closer target-to-mesh spacings

In discussing target-to-mesh spacing for the 2P23 tube, the spacings were arbitrarily divided into two groups: wide-spaced where the spacing is much greater than the size of a target element; and close-spaced where it is much less. Wide spacing is not used because the low capacitance results in an unusable signal-to-noise ratio and annoying white edges around lighted areas. The 2P23 has a spacing of 0.003 inch which is in the intermediate range where the signal-to-noise ratio is usable and the white edges are not particularly annoying. The question of what occurs as the spacing becomes closer is an important one.

If the spacing is reduced to 0.001 inch, the capacitance rises by a factor of about 3. The signal output will also be three time as high and the signal-to-noise ratio is improved by a factor of $\sqrt{3}$ or about 1.7, provided the modulation for the higher beam currents required does not change. Numerous experimental tubes have been made with this reduced target-to-mesh spacing in which the signal-to-noise ratio is definitely improved but is still borderline for high-quality studio work. The white edges have disappeared and the half-tone response is improved. This improvement is due to the much longer straight-line portion of the curve of target charge vs. illumination which is shown in curve A of Figure 6. If these tubes are operated so that the highlights are at a point somewhat over the knee of the curve, the half tones will have a better signal-to-noise ratio than the 2P23. The improved signal-to-noise ratio and the lack of white edges make the picture look more "natural".

The close-spaced assembly, however, is not without its drawbacks. Because an inferior picture is obtained below the knee of the signal-output curve, it is nearly always preferable to operate these tubes at the knee or slightly above it. Although a better picture is obtained at this point with a close-spaced target, for the 0.001-inch spacing three times as much light is required. This requirement makes the sensitivity of the experimental tubes appear to be lower than that of the 2P23. The close spacing is also not satisfactory for low-light scenes because the higher capacitance causes picture "lag". The higher capacitance causes a smaller potential rise at the target for a given amount of light. At these small potentials, the beam modulation is very poor and the target is not completely discharged in one frame time. For outdoor scenes the greatest drawback is probably a lack of versatility in handling a wide range of lighting. This drawback shows up in two ways. The beam current needed in the 2P23 is very small and the multiplier shading is a minor item. If the beam is set for a high-light condition and the camera swung over to a low-lighted scene, the shading is not noticeable and the scene can readily be handled. In the close spaced tubes, the higher beam current causes disproportionally more shading. Thus, when the camera is used on a high-light scene a picture of better quality can be obtained; but when the camera is swung to a low-light scene the shading and beam noise are often troublesome. If the beam current is reduced this trouble disappears. In field use, however, this reduction is not always possible, especially when a scene contains both sunlight and shadow which often happens during sport events. Also, because of the long straight section of the target-charge curve, the highlights have more tendency to cause "blooming' at the kinescope. In a scene with a few high lights the picture is set so that the normally lighted parts appear bright at the kinescope. The signal in the highlights therefore, can be high enough to cause loss of resolution in the kinescope and consequent "blooming".

From a manufacturing standpoint the close spacing also puts a greater demand on the thermionic cathode because of the need for a higher beam current. If the center of the cathode is slightly low in emission, the beam will be insufficient to discharge the highlights completely and loss of resolution will occur. Because of this demand for higher beam currents, grid No. 2 of the close-spaced tubes is operated as near to 300 volts as possible and not near 200 volts which has been satisfactory in some cases for the 2P23.

Operation and Construction of the 5655

Target-to-mesh spacing

Because it has not been possible to make a completely universal tube, it was decided to design the close-spaced tube for studio use where the lighting can be controlled. Under such conditions the apparent decrease of sensitivity and inability to handle scenes of widely varying illumination is not important. Because a 0.001-inch spacing does not give too good a signal-to-noise ratio, the question arises as to what happens at closer spacings. For a 500-mesh screen the center-to-center spacing is 0.002 inch. For 60 per cent transmission the hole size is about 0.0016 inch, that is, the center of the target element under each hole is 0.0008 inch from the mesh. One would expect that the capacitance would increase slowly for values of spacing less than 0.001 inch. However, the increase is definitely apparent in tubes with still smaller spacings. The capacitance increases until the target and mesh touch. The gain in capacitance over a 2P23 target is about 4 to 6 to 1 and the gain in signal-to-noise ratio is about 2 to 2.5 to 1. The signal is, of course, about 4 to 6 times higher and the apparent sensitivity ¼ to ⅙ as high. This gain in signal-to-noise ratio is very worthwhile and makes the tubes acceptable for studio application. For such tubes, the signal-to-noise ratio has an average value of about 80 to 1.

Target-mesh structure

The actual manufacture of a target-mesh structure with such a close spacing (the spacings are held between contact and 0.0004 inch) presents many problems. A description of the manufacturing process follows. The thin glass target is sealed at its edges to a metal ring. Before sealing, the surface of this ring is coated with a binder. After the seal is made, there is a measurable thickness of this binder above the glass. Also, at the inner edge of the ring the glass target tends to seal around the edge so that it is slightly depressed below the metal. If the mesh is then placed against this structure, it will have about the correct spacing for the 2P23. To obtain closer spacing it is necessary to design the structure so higher points on the mesh are actually pushed into contact with the glass target. The tolerances are so small that careful fabrication is needed. Any minor deviation from flatness of either the mesh or the target will also show up much more clearly in the close-spaced target.

Picture sticking

Before leaving the close-spaced structure, it is interesting to note that its "picture sticking" characteristics due to glass resistivity are different from those of the 2P23. It has been mentioned previously that in a cycle of operation the charges, remaining on the target after the beam has scanned it, must be neutralized in a frame time by conduction through the glass. If charge neutralization does not take place, the signal will fade for a fixed scene and return to its full value only when the picture is moved. A picture of opposite polarity will then be left on the target and can be seen if this area is lighted. The point of interest here is that for a given glass thickness and conductivity, fading depends on the target-to-mesh spacing in the region of very close spacings because, with the close-spaced assembly the capacitance and, consequently, the amount of charge to be neutralized are greater. For a 2P23 with a target of average thickness and conductivity, the percentage of fading at 20 degrees centigrade is about 10 per cent. If the temperature is increased to 30 or 35 degrees centigrade, the lowered resistivity makes the fading nearly negligible. For a close-spaced tube, however, the value at 20 degrees is closer to 30 per cent. A higher temperature of operation is needed to reduce fading to a low value. Generally, a temperature in the range of 40 to 45 degrees is sufficient. This need for a higher operating temperature, of course, reduces the operating temperature range of the close-spaced tube over that of the 2P23 because the 5655 is limited on the high side by lateral leakage in the same way as the 2P23. During tube life the 5655 target also changes more noticeably than that of the 2P23 because of the greater amount of charge that has to be transported through the glass. The change in contact potential, which generally makes the scanned area appear darker, may amount to several volts and requires shifting the mesh potential to more positive values during the life of the tube. In addition during operation the resistivity of the target slowly increases and after several hundred hours a higher temperature of operation may be needed.

Beat pattern

Because the mesh of the 5655 is nearly in contact with the target, the mesh is in better focus and is more visible than in the 2P23. The beat patterns also show up more readily. Some improvement has been made by using higher-transmission mesh with a transmission of 60 per cent or better. This value is equivalent to a wire 0.0004 inch in diameter. At present, however, the mesh limits the resolution of the tube in the highlights. As the techniques for manufacturing meshes improve, it is expected that eventually a finer mesh will be available.

Photocathode

Because the close-spaced tube has been designed for studio work a high sensitivity, especially for incandescent light, is not of prime importance. The spectral response, however, is of importance. Because of the variation in the spectral response of the 2P23 from tube to tube and because of the high infrared sensitivity of many of the tubes, it is very difficult to light the scene properly and get reproducible results. For these reasons, a different photosurface has been developed for the close-spaced 5655 tube.

The well-known photosensitive surfaces are those of cesium, silver-oxide, and silver, and those of cesium-antimony. Both can be made in the form of semi-transparent surfaces. It has already been shown that the first is not satisfactory for the 5655. The cesium-antimony surface gives a high response to both incandescent and fluorescent sources which is quite reproducible from tube to tube. It has no infra-red response but, unfortunately, also very little red response. Even with incandescent illumination its red response is too low to be satisfactory. A photosurface that overcomes this objection, however, has been developed. It consists of a silver-antimony surface sensitized with cesium. The silver and antimony are made into an alloy in the proportions that give best results. This alloy is evaporated onto the face plate at exhaust and then sensitized. The surface as shown in the spectral response curve of Figure 7 has the high blue response of the cesium-antimony (S4) surface plus an added red response. The over-all sensitivity to incandescent light is only about one-third of that of the 2P23 surface. For fluorescent light and sunlight, the surface compares very favorably with the 2P23. Under certain conditions such as near sunset when considerable blue light is scattered, the 5655 surface is more sensitive.

Operation

Because of the greater capacitance and different photocathode

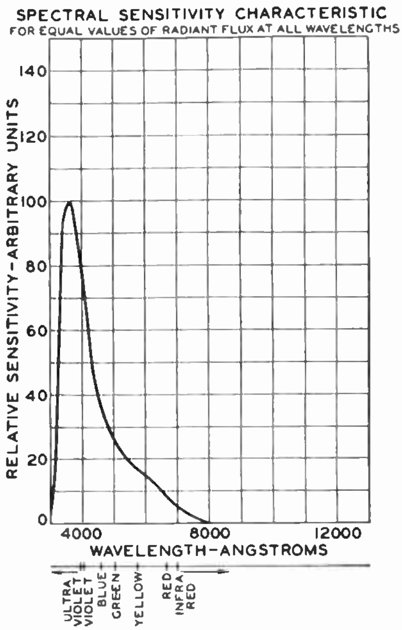
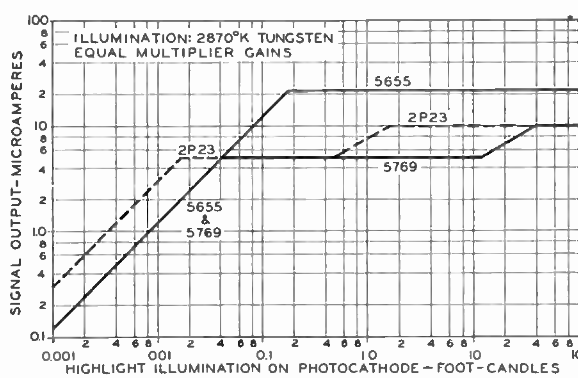


Fig. 7—Approximate spectral sensitivity characteristics of image orthicon types 5655 and 5769.

surface, the signal-output of the 5655 differs from the 2P23. The signal output curve is given in Figure 8. Because the maximum output is 4 to 6 times that of the 2P23, a two-fold or better increase in the signal-to-noise ratio results although more light is needed to obtain this higher value. As the curve indicates, the illumination is of the order of 0.2 foot-candle at the photocathode. In order to obtain good depth of focus, a minimum of 100 foot-candles should be used on a scene, with 200 to 300 a better choice.

The type of lighting needed to get a good spectral response is fairly well met with a mixture of fluorescent and incandescent sources. In general fluorescent lamps of 3500 or 4500 degrees Kelvin color temperature are most suitable for obtaining a good level of lighting. However, when only fluorescent lighting is used, the red response is somewhat lower than the yellow or blue. The use of incandescent flood lights will help this condition and at the same time permit high lighting of various scenes. Although the 5655 is able to handle a wide range of light it is more restricted than the 2P23. For this reason care should be taken to eliminate very brightly lighted

Fig. 8 — Simplified signal output characteristics for image orthicon types 2P23, 5655, and 5769.



areas such as can be caused by highly reflecting objects or by the use of intense spot lighting.

OPERATION AND CONSTRUCTION OF THE 5769

As has been discussed, the 2P23 has the advantages of fair signal-to-noise ratio, good ability to handle a wide range of scenes, and good sensitivity particularly for incandescent lighting. However, because of its high red and infrared response, it does not portray colors faithfully and resolution suffers particularly for the colors near the blue. The 5655, on the other hand, has a good color response. The 5769 is a new tube which is identical with the 2P23 except that it has the 5655 photocathode surface and color response. Its sensitivity for incandescent light is only about one third that of the 2P23 but for fluorescent and daylight it is nearly the same. For a large majority of outdoor pickups it is superior to the 2P23 because of its color response. In addition, operating experience has shown the 5769 to be more stable. Because the photocathode surface of the 2P23 demands an excess of cesium for best sensitivity, it is difficult to prevent migration of the cesium around the tube while it is on the shelf or in operation. If the excess cesium migrates to the target the resolution and signal output of the tube will be lowered, and the tube will fail because of "target leakage". The 5655 and 5769 have, in general, proved less susceptible to this condition and, consequently, are more stable. Because of its good color response, the 5769 can also be used for studio operation. In this use it is somewhat easier to handle because of its wide light range. The 5655, however, is superior in its signal-to-noise ratio and rendition of grays.

OVER-ALL OPERATING CHARACTERISTICS

The over-all operating characteristics of image orthicon tubes can best be considered in terms of the curves of signal output versus light. Figure 6 shows curves of the charge built up during a frame time for increasing illumination on the photocathode. If the charge is divided by the frame time of 1/30 second and multiplied by the multiplier gain, the signal output will be obtained. For a gain of 1000, the maximum signal for a 2P23 of average mesh spacing is about 5 microamperes. For a photocathode sensitivity of 14 microamperes per lumen and a secondary emission ratio of 2.5 at the target (the ratio of the cesiated glass is probably about 4 at 400 volts but the mesh transmission is 60 per cent), the high light illumination on the photocathode needed to obtain this maximum signal is about 0.02 foot-candle. With this information the signal output curve (Figure 8) for the 2P23 can be constructed. As in the charge curve, the signal-output curve rises linearly with light until a point is reached which corresponds to the maximum signal output. Above this point it does not rise except for very high values of light where interelement capacitance comes into effect. This curve is extremely simplified. It holds more closely for the case of a small area of light on a dark background, but even under this condition the break at the knee is rounded off because of the initial emission velocities of the secondaries from the target. These velocities range from 0 up to 5 volts with an average value of about 2 volts.

Should there be more than one bright spot, the curve becomes more complicated. Any lighted area tends to preserve its contrast by reducing the charge on the areas with a smaller amount of light by spraying them with secondary electrons. This spray, however, will also tend to discharge the poorly lighted areas which already have a rather low signal. Brightly lighted areas will, consequently, suppress the grays and make the gray parts of the scene appear flat and noisy. Gray suppression can be reduced by keeping the mesh potential high so that more of the electrons are collected and not redistributed. However, this expedient is of the most help when the highlights are just above the knee of the curve.[9]

In its present state the image orthicon can handle a scene containing very bright high lights and grays better than an orthicon. In outdoor work many such scenes will be encountered. Inferior results in the grays will always occur, however, so that where lighting is under control such as in the studio, every attempt should be made to avoid extreme highlights. Not only do the secondary electrons redistributed directly from the highlight parts of the target suppress the grays, but also secondary electrons from the mesh have the same effect. Internal reflection of the highlights inside the lens and tube add to this limitation. Very poor results are obtained if a direct light gets into the lens from the sun or other sources.

A simplified signal output curve is also shown in Figure 8 for the 5655. Because of its high capacitance, the maximum signal output of the 5655 is higher than that of either the 2P23 or 5769. This advantage leads to a higher possible signal-to-noise ratio and a better gray scale because of the longer straight part of the curve. For the same photosensitivity, however, more light is needed to reach these improved conditions. In addition, more beam current is necessary to discharge

[9] O. H. Schade, "Electro-Optical Characteristics of Television Systems", *RCA Review*, Vol. IX, No. 4, pp. 663-665, December, 1948.

the highlights. For any given scene, the 5655 can produce, in general, a picture as good or better than that produced by a wide-spaced tube. It also has about the same ability to hold down highlights. However, because of the greater beam current needed for a strongly illuminated picture, when the camera is turned to a low light scene, the high beam current produces more noise and shading than is produced by the 2P23 or 5769. If the beam current is reduced for the low-light scene, the picture is again good. Because a continual shifting of the beam current is not practical, the 5655, in general, is not recommended when high-light and low-light scenes must be picked up in quick succession.

CONCLUSION

This paper has discussed in some detail the limitations and operating problems as well as the advantages of the image orthicon. As a result of a continuous program of development, many of the limitations of this tube have been minimized and the image orthicon has evolved from a laboratory device to its present status of being the work horse of television. With continued development, the possibilities of which have by no means been exhausted, the image orthicon can be expected to show further steady improvement.

The operating problems show a measure of the complexity of the image orthicon. This complexity, however, makes it possible for this tube to do what no other camera tube can do. An outstanding advantage of the image orthicon is its exceptional sensitivity which enables the tube to pick up scenes illuminated at very low light levels—only a few foot-candles—and greatly extends the range of outdoor subjects which can be televised. In addition, the image orthicon can reproduce scenes having great depth of field—a valuable advantage which provides flexibility of operation. Both in outdoor and studio pickups, the use of image orthicons has resulted in steady improvement in picture quality. The tube has earned its place today as the best choice for a universal camera tube.

ACKNOWLEDGMENT

This development program has involved the continual aid of many groups. In particular, the authors wish to acknowledge the help of the following persons: A. Rose, P. Weimer and H. B. Law of RCA Laboratories Division, Princeton; O. H. Schade of the Tube Department at Harrison; H. N. Kozanowski, N. Bean and J. H. Roe of the Engineering Products Department at Camden; E. D. Goodale and the operating groups of NBC; D. Ulrey, L. B. Headrick, P. A. Richards, R. E. Barrett, R. Handel and L. Young of the Tube Department at Lancaster.

METHOD OF MULTIPLE OPERATION OF TRANSMITTER TUBES PARTICULARLY ADAPTED FOR TELEVISION TRANSMISSION IN THE ULTRA-HIGH-FREQUENCY BAND*

By

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Summary*—*A combining network has been developed which allows two transmitting tubes to be operated simultaneously into a common load without interaction between tubes and without reduction in band width. A number of variations of the combining network are discussed and a theoretical analysis is presented which shows that the necessary balancing adjustments are not critical.

A pair of tubes and a combining network may then be considered as a unit, with this unit combined with several other identical units to provide a transmitter with a large number of tubes in multiple operation. In this arrangement, each tube is free from interaction with other tubes.

Rather simple circuits which apply the principles set forth are described for operation at low radio frequencies. A complete television transmitter with a carrier frequency of 850 megacycles has been developed, using four tubes in multiple to demonstrate the principle of operation.

Introduction

THE ability to generate radio-frequency power has generally kept abreast of the demand for increases in power, particularly in the broadcast band of frequencies and in the medium-high frequencies. With large power tubes available and the techniques of multiple use of tubes in push-pull or parallel operation quite commonplace, transmitter design resolves itself into a problem of economics and good engineering practice. Estimates of power requirements for ultra-high-frequency television broadcasting, however, are far in excess of the power capabilities of any commercially available single tube or of any simple push-pull combination of these tubes.[1,2]

* Decimal Classification: R355.16 × 583.4.

[1] George H. Brown, J. Epstein, and D. W. Peterson, "Comparative Propagation Measurements; Television Transmitters at 67.25, 288, 510, and 910 Megacycles," *RCA Review*, Vol. IX, No. 2, pp. 177-201, June, 1948.

[2] George H. Brown, "Field Test of Ultra-High-Frequency Television in the Washington Area," *RCA Review*, Vol. IX, No. 4, pp. 565-584, December, 1948.

Several tubes may be used in essentially parallel operation by arranging the tubes in a circle on a common cavity.[3] In this method, the number of tubes is limited by practical considerations of high circulating currents in the tank circuit and criticalness of tuning, both effects due to the paralleling of the tube capacities. The authors have undertaken a study of circuit arrangements which alleviate these difficulties and have developed a simple bridge circuit which permits the multiple operation of transmitter tubes into a common load without interaction between tubes and with no limitation on band width other than that imposed by a single tube and tank circuit. In this method of operation, each final amplifier tube has its own associated tank circuit, feeds into a pure resistance load, and is entirely oblivious to the existence of the other amplifier tubes.

The Philosophy of the Use of a Bridge Circuit to Accomplish Multiple Operation of Amplifier Stages

The bridge circuit which forms the heart of this multiple operation method may be depicted for illustrative purposes by Figure 1. However, the reader should remember that this circuit *per se* is not readily applicable to the problem at hand. Circuits appropriate to particular frequency ranges will be described later in the paper.

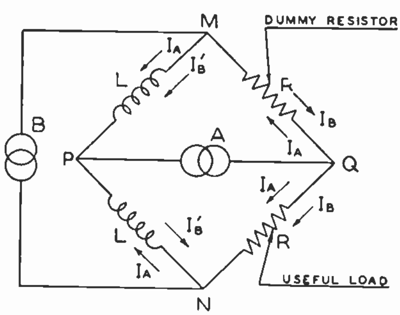

Fig. 1—A bridge circuit used to illustrate the principle of multiple operation.

The bridge of Figure 1 consists of two equal inductances and two equal resistances. This bridge is balanced, that is, the generator *A* produces no voltage across the terminals *M-N* and the generator *B* produces no voltage across the terminals *P-Q*. Thus, the two generators may operate simultaneously without interaction and the currents in the network produced by generator *B* simply superimpose upon the currents produced by generator *A*. The voltages generated by *A* and *B* are assumed to be sine waves of identical frequency.

Suppose for the moment that generator *B* is inoperative. Then generator *A* produces the two currents denoted as I_A in each of the resistors. One resistor in Figure 1 is called the useful load while the

[3] Donald H. Preist, "Annular Circuits for High-Power Multiple-Tube Generators at VHF and UHF," presented at the 1949 I.R.E. National Convention, New York City, March 9, 1949.

other is designated as a dummy resistor, for reasons which will soon become apparent. The power dissipated in the dummy resistor is $I_A^2 R$ and the power in the useful load is of exactly the same value. Hence the power delivered by generator A is

$$P_A = 2\, I_A^2\, R. \tag{1}$$

If generator A now becomes inoperative and generator B delivers power to the bridge, the currents in the two resistors will be $\bar{I}_B$ and the total power delivered by generator B is

$$P_B = 2\, I_B^2\, R \tag{2}$$

with this power divided equally between the two resistors.

Now let both generators become operative and assume that the voltage produced by generator B can be completely controlled with respect to both amplitude and phase. Control is exercised until the current I_B in the useful load is exactly equal to I_A in both amplitude and phase. Under this condition, the net current in the dummy resistor is zero and no power is dissipated in this dummy resistor. The power in the useful load is

$$P_U = (I_A + I_B)^2\, R = (2\, I_A)^2\, R = 4\, I_A^2\, R \tag{3}$$

and the total power of the two generators is delivered to the useful load. The two generators remain uncoupled one from the other, with the total power concentrated in the single useful load. It is interesting to note that when one generator ceases to operate, the current in the useful load is halved and the power in this load goes to one quarter of the full load power.

When the bridge circuit was first considered, the ability to maintain sufficiently accurate balance of amplitude and phase was immediately questioned. A subsequent analysis, given below, soon showed the rather remarkable insensitiveness and practicality of the circuit arrangement. To illustrate this point, assume that the currents $\bar{I}_A$ and $\bar{I}_B$ are no longer equal and in phase but are related as follows:

$$\bar{I}_A = K\bar{I}_B \angle\beta \tag{4}$$

where K is a simple numerical coefficient. For this condition

$$P_B = 2\, I_B^2\, R \quad (5) \qquad \text{and} \qquad P_A = 2\, I_A^2\, R = 2\, K^2\, I_B^2\, R \quad (6)$$

with the total power given by $$P_A + P_B = 2(1 + K^2)\, I_B^2\, R. \tag{7}$$

Then the total current in the useful load is

$$\bar{I}_U = \bar{I}_B (1 + K \cos\beta + jK \sin\beta) \tag{8}$$

and the power in the useful load is

$$P_U = (1 + K^2 + 2K \cos\beta)\, I_B^2\, R. \tag{9}$$

Likewise, the current in the dummy load is

$$\bar{I}_D = \bar{I}_B (1 - K \cos\beta - jK \sin\beta) \tag{10}$$

and the power lost in the dummy resistor is

$$P_D = (1 + K^2 - 2K \cos\beta)\, I_B^2\, R. \tag{11}$$

One may easily note that the sum of the Equations (9) and (11) is identical with Equation (7).

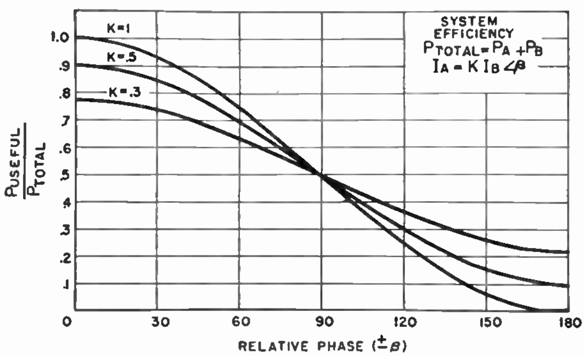

Fig. 2—Power in the useful load in terms of the total power as a function of the phase relationship between the currents produced by the two generators.

Dividing Equation (9) by Equation (7) gives

$$\frac{P_U}{P_A + P_B} = \frac{1 + K^2 + 2K \cos\beta}{2(1 + K^2)}. \tag{12}$$

Equation (12) is plotted in Figure 2 as a function of the phase angle, β, with the numerical value of K as a parameter on each curve. This diagram reveals the inherent insensitiveness of the circuit to correct phase adjustment. To be specific, suppose that generators A and B are each delivering 500 watts to the circuit. Then with truly

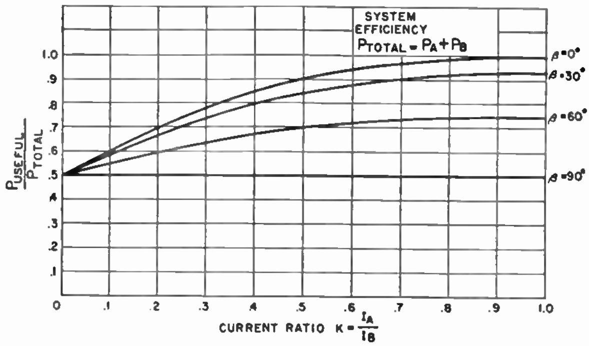

Fig. 3—Power in the useful load in terms of the total power as a function of the ratio of the currents produced by the two generators.

zero phase the power into the useful load will be 1000 watts. Reference to Figure 2 shows that with β equal to 30 degrees, the power into the useful load will be 933 watts. If the useful load is an antenna, the field strength will drop less than four per cent with this degree of phase misadjustment.

The calculations of Figure 2 have been replotted in Figure 3 to better illustrate the relative "flatness" of circuit conditions with variation in the parameter K.

So far in the analysis, it has been assumed that the two generators produced sine waves of identical frequency. It is reasonably apparent that if the signals are of complex wave form, but identical, cancellation in the dummy resistor will still be secured and the additive condition in the useful load realized. If the two generators represent modulated power amplifiers, the necessary conditions of operation are that the carriers are substantially in phase in the useful load and that the modulation of the two output stages is identical and simultaneous.

A pair of output tubes and a combining network may now be considered as a unit, with this unit combined with several other identical units to provide a transmitter with a large number of tubes in multiple operation. In this arrangement each tube is free from interaction with other tubes. Figure 4 illustrates the manner in which eight tubes are combined with seven bridges, or diplexers. It is now apparent that if n tubes are combined, the number n must be 2 raised to an integral power, that is, n must be 2, 4, 8, 16, and so on.

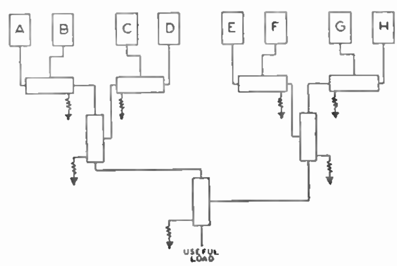

Fig. 4—A combination of eight tubes and seven diplexers to accomplish simultaneous operation into a common load without interaction between tubes.

Also, the number of bridges required is $n-1$.

Since the currents in the outputs of the bridges are additive, it is a simple matter to estimate the overall efficiency and other operating conditions of interest without tracing through the rather involved bridge network. To accomplish this estimate, the following nomenclature has been set up:

P = power output of each output stage.

n = total number of stages.

nP = total power in useful load when all n stages are turned on.

mP = total power available when m stages are turned on.

I_n = current in useful load with n stages turned on.

I_m = current in useful load when m stages are turned on.

$I_m = (m/n) \cdot I_n$.

P_m = power in useful load when m stages are turned on.

$P_m = I_m^2 R = (m/n)^2 (I_n^2 R) = (m/n)^2 \cdot (nP)$.

Ratio of power into useful load with m stages turned on to total power available from n stages $= P_m/(nP) = (m/n)^2$.

Circuit efficiency when m tubes are on $= (P_m/mP) \cdot 100 = m/n \cdot 100$.

Table I—Conditions of Operation in Seven Combining Networks and Eight Output Stages

m (Number of stages operative)	$\frac{mP}{nP} = \frac{I_m}{I_n}$	$P_m/(nP)$
8	1.0	1.0
7	0.875	0.766
6	0.75	0.562
5	0.625	0.391
4	0.5	0.25
3	0.375	0.141
2	0.25	0.063
1	0.125	0.016
0	0	0

As an example, suppose that each of eight tubes is capable of delivering 100 watts. Then a total of 800 watts is available for the useful load. With any three of the tubes turned on, the power delivered to the networks is 300 watts. The current in the useful load is m/n or $3/8$ of the current found there when eight tubes are on. The power into the useful load, P_m, is 112.5 watts, and the combining circuit efficiency is 37.5 per cent.

Table I illustrates still further the conditions of operation for n tubes, with n equal to eight.

Practical Bridge Circuits for Low-Frequency and High-Frequency Operation

A simple method of applying the bridge circuit principles at low frequencies, of the order of a few megacycles or less, is illustrated in Figure 5. It is apparent that this elementary application is a one-step bridge circuit which does not lend itself to the repetitive use outlined above.

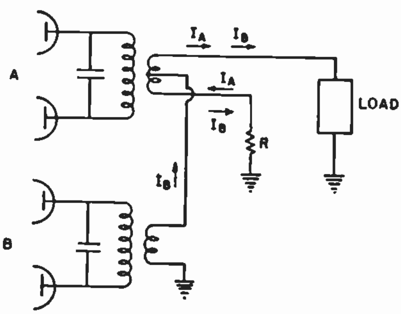

Fig. 5—A means of combining four tubes at a low frequency.

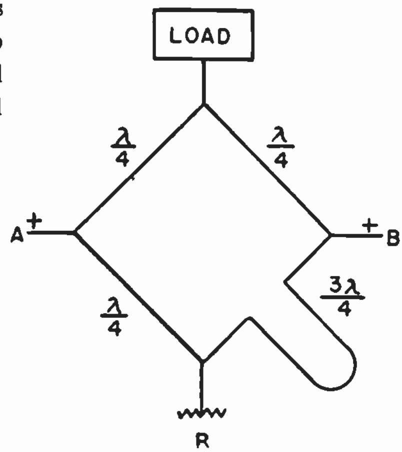

Fig. 6—A bridge arrangement constructed of lengths of transmission line.

A circuit much better suited to cascading at low frequencies may be best developed by referring to Figure 6. In this particular diagram, the arms of the bridge are coaxial transmission lines. Three of the arms are each one-quarter wave length at midband, while the fourth arm is three-quarters of one wave length. For the sake of simplicity, only the inner conductors are shown in Figure 6. Perfection of uncoupling between points A and B depends upon the exactness of these line lengths and the device is strictly limited to a narrow band of frequencies. When the useful load and the dummy resistor have a resistance of R ohms, and the characteristic impedance of the transmission line arms is chosen as $\sqrt{2} \cdot R$, the resistance looking in at points A or B will be R ohms at midband. While the above choice of characteristic impedance plays no part in the balancing action at midband, affecting only the input impedance, this same choice does help in broadbanding the circuit.

The circuit of Figure 6 may now be used as a guide in forming a lumped-circuit network for use at low frequencies. This has been done in Figure 7. Each one-quarter-wave line has been replaced by a Pi network consisting of two capacitors and one inductance coil. The inductance and capacitance values have been so chosen that X_L equals X_C at midband. The fourth arm is formed from two inductances and one capacitance to be the equivalent of the three-quarter wave long branch. When $X_L = X_C = MR$, the input impedance at A and B is

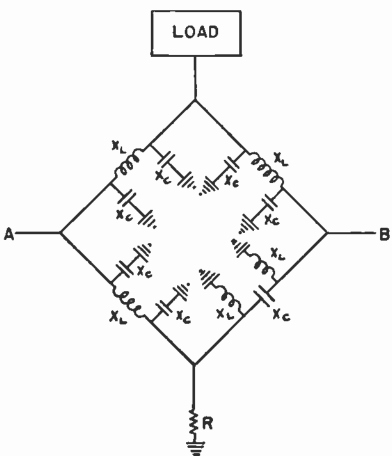

Fig. 7 — A bridge circuit constructed of lumped elements, to be the equivalent of Figure 6.

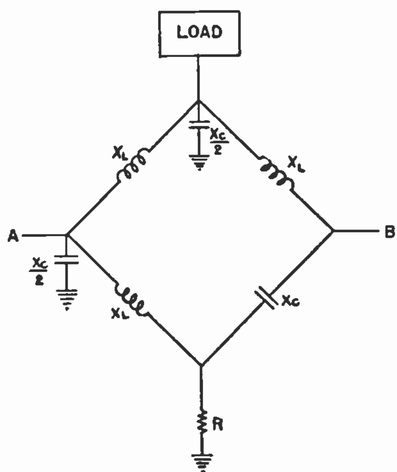

Fig. 8—A simplified circuit, equivalent to Figure 7.

$M^2R/2$ ohms. Again a choice of M equal to $\sqrt{2}$ gives maximum broadbanding for the input impedance. Figure 7 may be considerably simplified as shown in Figure 8 when it is noted that an inductive reactance in parallel with an equal capacitive reactance forms a parallel resonant circuit and both elements may then be omitted from the system.

The narrow-band limitations of Figure 6 may be avoided by the use of the circuit shown in Figure 9. Here the balance between feed points is independent of frequency and the variation of impedance at the feed points establish the limits on the frequency band.

If a circuit balanced to ground is desired, the circuit shown in Figure 10 is useful. Here the arms of the bridge may be either parallel

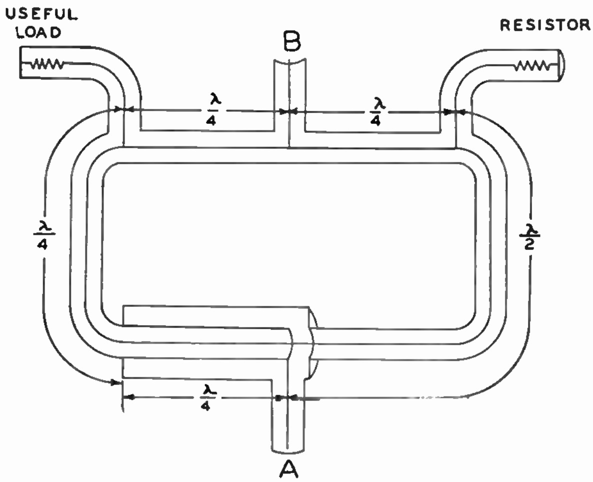

Fig. 9—A high-frequency bridge which is not sensitive to frequency change.

wire lines or two coaxial cables. This circuit does not depend on the line lengths being one-quarter wave for balance. The line lengths shown do establish the midband frequency as far as the input impedance is concerned. Here again a choice of characteristic impedance equal to MR ohms yields an input impedance of $M^2R/2$ ohms and a value of M equal to $\sqrt{2}$ gives the broadest input impedance characteristic. The circuit of Figure 10 with M equal to unity is described by Westcott.[4] The authors have found that a choice of M equal to $\sqrt{2}$ yields a much more desirable input impedance versus frequency characteristic.

For higher frequencies, particularly the ultra-high frequencies, the authors have found the bridge or diplexer of the slotted type shown in Figure 11 to be the simplest and most easily balanced. This is the same diplexer which has been used so very successfully to diplex a Turnstile antenna, a method of operation which permits both picture and sound transmitters to be fed to a single Turnstile.

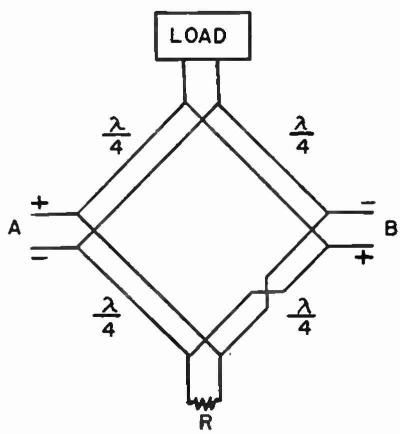

Fig. 10 — A bridge arrangement which is balanced to ground.

[4] C. H. Westcott, "Transmission-Line Bridge," *Wireless Engineer*, Vol. XXV, No. 298, p. 215, July, 1948.

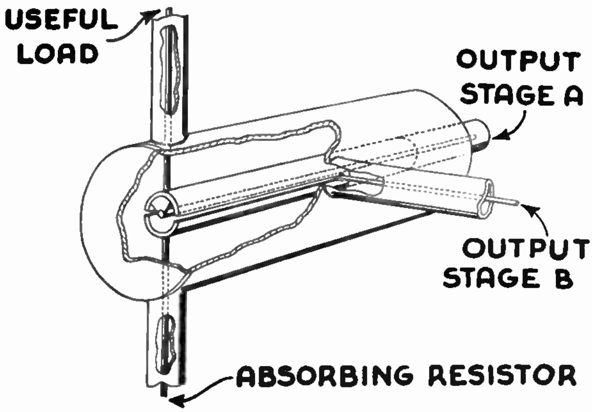

Fig. 11 — A slotted diplexer or bridge which is particularly useful at ultra-high frequencies.

AN ULTRA-HIGH-FREQUENCY TELEVISION TRANSMITTER APPLYING THE BRIDGE ARRANGEMENTS FOR MULTIPLE OPERATION OF OUTPUT STAGES

To demonstrate the principle of multiple operation, the authors developed a television transmitter using four RCA-5588 tubes in the output stages, to produce a total power of 400 watts. The circuit arrangement is shown in Figure 12. A crystal oscillator operated at a frequency of 7870.4 kilocycles. When the output of this oscillator was passed through suitable frequency multiplying stages, the exciter developed a signal with a frequency of 212.5 megacycles. At this point, the chain was broken into four parallel paths. Each path then led through two doubler stages to furnish driving voltage to a final amplifier stage at 850 megacycles. Three diplexers and three absorbing resistors were used as shown to combine the four output stages into a

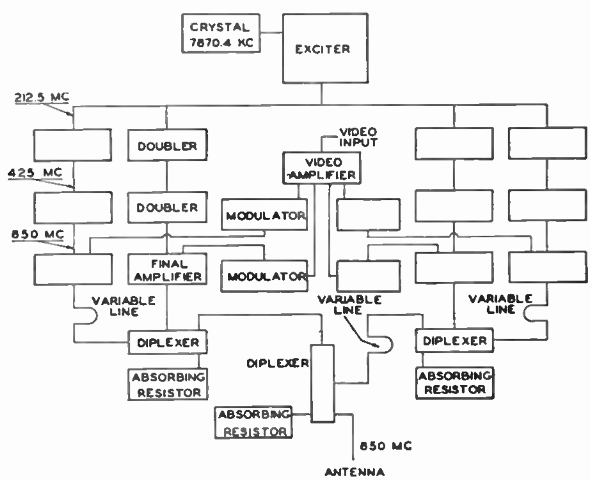

Fig. 12—A block diagram of an 850-megacycle television transmitter using the principle of multiple operation.

single antenna. Each final amplifier was cathode modulated from a single picture source to accomplish simultaneous modulation of each final amplifier. The diplexers were of the type shown in Figure 11.

Three variable length lines, shown in Figure 12, were used to provide phase adjustment of the radio-frequency carriers. These variable length lines were constructed of overlapping tubes, adjusted by a rack and pinion. Experience has indicated that these lines may be omitted and small phase adjustments accomplished by slight tuning of amplifier tank circuits in the doubler stages.

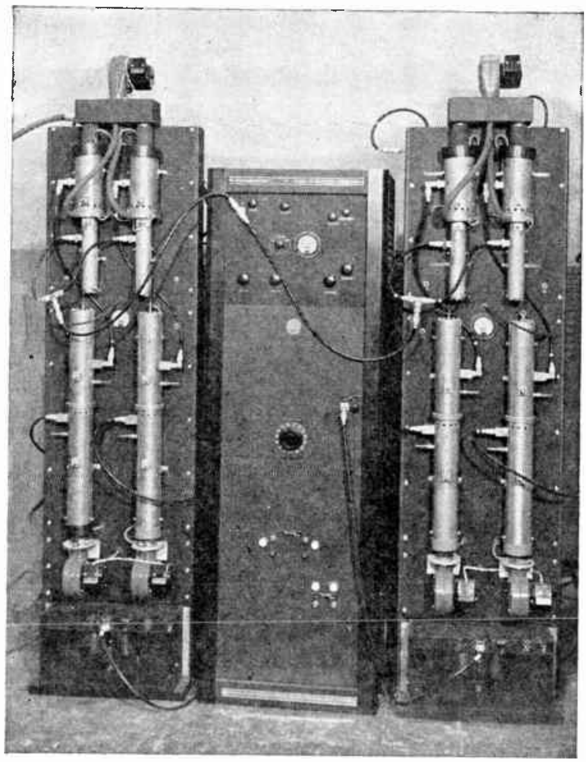

Fig. 13—The experimental 850-megacycle television transmitter.

One or more of the final amplifier stages could be turned on or off without reaction on the remaining stages. Signal strength in the antenna circuit changed according to theoretical predictions.

The transmitter is shown in Figures 13 and 14, with the diplexers visible in Figure 14. This transmitter was operated under an experimental license as Station W3XCY in Washington, D. C. in the fall of 1948 and has since been in operation in Princeton, N. J., as Station KE2XAY for the purpose of providing further test data in connection with the principle of multiple operation.

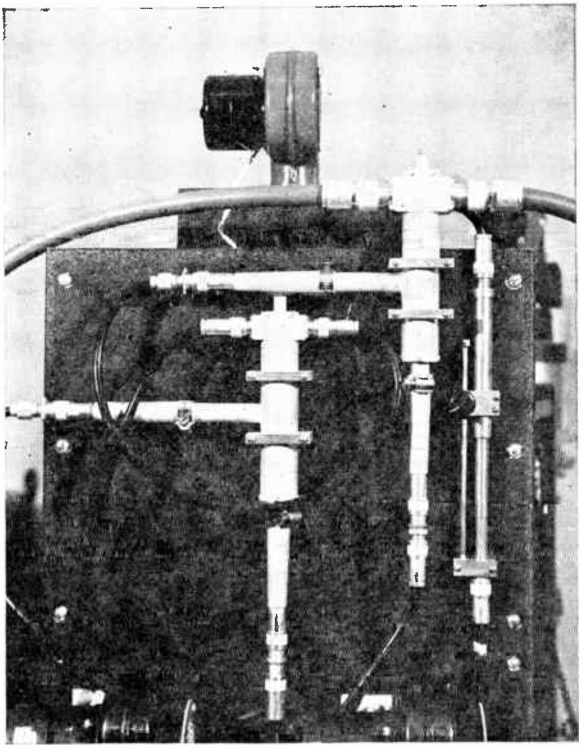

Fig. 14—Another view of the 850-megacycle television transmitter, showing the diplexers used in securing multiple operation.

CONCLUSION

A combining network has been developed which allows transmitting tubes to be operated simultaneously into a common load without interaction between tubes and without reduction in band width. A transmitter has been constructed which shows the method to be applicable to the design of ultra-high-frequency transmitters for television use. While relatively low power tubes were used in this demonstration transmitter, it was done for purposes of expediency and to demonstrate the principle of operation. The authors do not mean to imply that several small tubes are to be preferred to one large one. However, when the largest tube available does not approach the power desired for the particular service under consideration, multiple operation with diplexing circuits seems to be indicated as a practical solution.

EXPERIMENTAL ULTRA-HIGH-FREQUENCY TELEVISION STATION IN THE BRIDGEPORT, CONNECTICUT AREA*

BY

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Editor's Note: This paper constitutes the first in a series of reports on the NBC UHF field tests at Bridgeport, Connecticut. The second of the series—"An Experimental Ultra-High-Frequency Television Tuner"—appears on pages 68-79 of this issue.

It is currently planned to include the following two papers in the June 1950 issue of RCA Review:

"A New Ultra-High-Frequency Transmitter"

"Ultra-High-Frequency Antenna and System for Television Transmission"

Subsequent papers will include reports on propagation studies, service area surveys, service operating characteristics of equipment, and subsequent equipment and other technical developments.

Summary—The engineering considerations involved in the construction and operation of an experimental television broadcast station in the ultra-high-frequency band are presented. The transmitter, KC2XAK, is located in the Bridgeport, Connecticut area and operates in a standard bandwidth of six megacycles from 529 to 535 megacycles with a newly developed transmitter and high-gain antenna. Programs are picked up directly from Station WNBT in New York on Channel 4 and are demodulated, processed and retransmitted on the ultra-high-frequency band.

INTRODUCTION

As World War II drew to its close, it became apparent that great expansion in radio service was imminent, particularly in the very-high-frequency (VHF) and ultra-high-frequency (UHF) spectrums. It was also evident that the whole field of frequency allocations in these spectrums should be reviewed in preparation for these new and extensive services of the future. One of the most important of the new services under consideration for the postwar period was television broadcasting. Accordingly, the Federal Communications Commission (FCC) held a public hearing which began on September 28, 1944, with the purpose of reviewing existing allocations in the light of future needs. As a result of this hearing and subsequent developments, commercial television is now assigned twelve channels in the VHF band on which there are currently in operation approximately one hundred stations. But it has been apparent that twelve

* Decimal Classification: R588×R310.

channels do not permit adequate television service for a truly nationwide system.

For future use the FCC set aside a block of ultra-high frequencies in the 475-890 megacycle band for television. Insufficient information was available with which to adopt standards and allocate frequencies at that time. The need for this information has been most apparent and both government and the industry have undertaken to obtain such information. It is necessary to determine whether or not the television transmitter standards presently used for VHF could be adopted for UHF service. It is also necessary to determine the propagation characteristics of the ultra-high frequencies. These two generalized fields require a large amount of data which must be obtained and integrated. The areas which could be served by UHF television transmitters with practicable radiated powers and antenna heights must be determined. Moreover, the propagation characteristics which determine the minimum separation which can be tolerated between cochannel stations, and the characteristics of transmitting and receiving apparatus and the propagation characteristics which determine the minimum separation necessary between adjacent-channel stations must be known.

During the last several years, much has been learned from work done in the development of apparatus and the studies of propagation undertaken by various government and private laboratories and by manufacturing and operating companies.

Radio Corporation of America has conducted, concurrently, a number of projects[1,2] to determine the propagation characteristics and the television service potential of the ultra-high frequencies, particularly in the band from 475 to 890 megacycles, and has made this information available to the FCC and the industry at large.

Upon completion of the tests described in Reference (2), it was decided that an experimental ultra-high-frequency television transmitting station should be erected in a representative city which was not adequately served by a local VHF transmitter. It was felt that such a station should be a full scale custom built prototype of future commercial installations in the UHF band so that the results obtained would be truly indicative of the practical possibilities of ultra-high-frequency broadcasting in the type of community in which many of these stations would be operated. Accordingly, such a project was initiated.

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[1] G. H. Brown, J. Epstein and D. W. Peterson, "Comparative Propagation Measurements; Television Transmitters at 67.25, 288, 510 and 910 Megacycles", *RCA Review*, Vol. 9, No. 2, p. 177, June 1948.

[2] G. H. Brown, "Field Test of Ultra-High-Frequency Television in the Washington Area", *RCA Review*, Vol. 9, No. 4, p. 565, December 1948.

Selection of the Site

After considerable investigation, an area having Bridgeport, Connecticut as its approximate center was selected for the station site. Application for a construction permit for an experimental television station, to transmit in the band from 529 to 535 megacycles and specifying Bridgeport as the general location, was made to the FCC on February 8, 1949. The permit was granted May 4, 1949, assigning the call KC2XAK.

The city of Bridgeport is located on Long Island Sound (see Figures 1 and 2) at the mouth of the Poquonock River and has a population, for the metropolitan district, of approximately 216,600 according to the 1940 census. While fringe reception of television stations in New York and New Haven is obtained in this area, there is no locally originated television service.

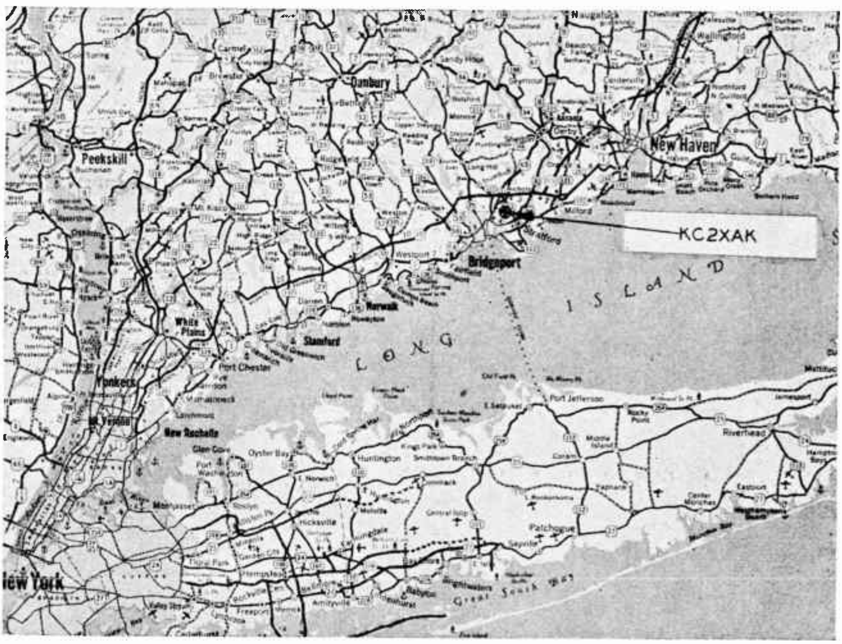

Fig. 1—Map of general area surrounding Bridgeport transmitter location.
Courtesy Esso Standard Oil Company (Copyrighted by General Drafting Company, Inc., New York.)

As can be seen from Figure 2, the city is ringed with a series of hills all of which are about 200 feet high. Of these, Success Hill, in Stratford, Connecticut, which is north-northeast of the center of Bridgeport and just outside the city limits, was chosen after extensive surveys as the most suitable site for the installation of a television transmitter. From the standpoint of covering not only Bridgeport, but

Fig. 2—Topographic map of Bridgeport and immediate vicinity.

(Geological Survey, United States Department of the Interior.)

Fig. 3—Transmitter Building (front view).

also neighboring communities such as Stratford, Devon and Milford adequately, Success Hill appeared to be the most attractive location.

Application was thereupon made to the FCC, modifying the Construction Permit to show Success Hill as the exact location of the transmitter site. This modification was granted by the FCC on October 12, 1949, with a proviso that construction should start on or about December 12, 1949, and be complete on or before June 12, 1950.

Transmitter Building

The transmitter building resembles a conventional Cape Cod cottage from the exterior, as shown in Figure 3. The floor plans of this structure are shown in Figure 4. The useful floor area of the apparatus rooms is 1164 square feet.

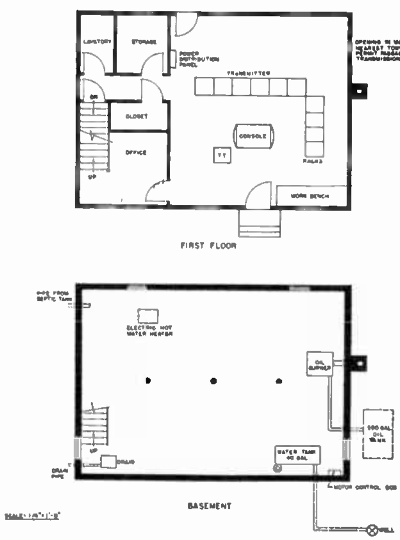

Fig. 4—Floor plans of transmitter building.

The peak power requirement of the installation was initially estimated as 50 to 60 kilovolt-amperes at 240-120 volts alternating current and it was therefore necessary to provide a 400-ampere service in order to achieve adequate regulation. Ventilation of the transmitting equipment is secured by means of vents located in the first floor ceiling which are arranged to accept the air exhausted from the transmitter racks. The attic in turn is ventilated by a pair of two speed 24-inch exhaust fans which are operated whenever it is undesirable to retain the equipment heat within the building. The construction of the transmitter building started on September 22, 1949 and was completed on November 15, 1949.

As a supporting structure for the transmitting antenna, a steel tower 210 feet in height with a base 24 feet square was erected next to the transmitter building as shown in Figure 3. In order to adequately ground the tower as protection against lightning discharges, copper straps six inches wide were attached to three of the four tower legs and these were separately bonded to the well casing. A considerable reduction in cost of the tower lighting system was effected by employing a newly developed weatherproof cable, Simplex Anhydrex, for the lighting cable runs, rather than conventional cable in conduit. These cables were secured to the tower ladder risers by means of worm-type hose clamps and this arrangement permitted installation of the tower wiring in a fraction of the time normally required.

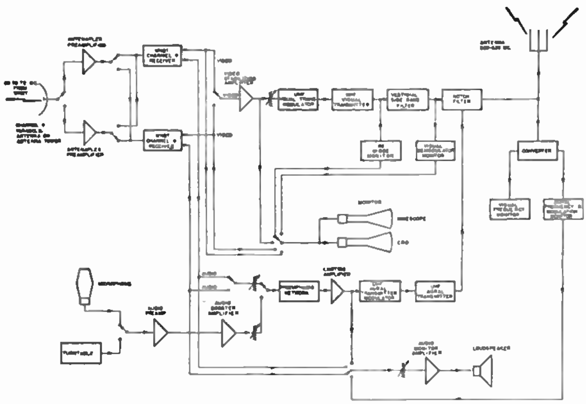

Fig. 5—Block diagram of transmitting system.

An efficient ground system for the radio transmitting equipment installed in the building was secured by situating all equipment racks on a continuous copper sheet placed over the flooring. This sheet was grounded to the well casing in the manner described above. Erection of the tower was started on November 17 and completed on November 24.

Transmitting System

The transmitting system employed at Station KC2XAK is outlined in the overall system block diagram shown in Figure 5. The transmitter is arranged to operate as a satellite of the VHF television station WNBT in that the visual and aural program signals are picked up directly from WNBT and retransmitted on the UHF band. While provisions are incorporated for local aural station identification, none are included for local origination of test pattern or other video signal.

The manner in which this system of satellite operation is accomplished may be seen from Figure 5. The signal from WNBT is picked up directly from the transmitter located atop the Empire State Building in New York City at a distance of approximately 54 miles by means of a parabolic antenna located on the antenna tower. The signal is then fed to a preamplifier located in the station building which in turn provides the radio-frequency signal for two specially constructed receivers tuned to Channel 4.

The video signal is taken from the receiver through a built-in isolation amplifier and is then processed by a stabilizing amplifier which improves the quality of the signal and corrects any degradation which may have occurred in the synchronizing information. The video signal is then applied to the UHF visual transmitter modulator and final power amplifier. The output of the visual transmitter is filtered by a vestigial sideband filter to attenuate the lower sideband in accordance with the television transmission standards established by the FCC.

The demodulated audio signal from the receiver is taken from the audio output stage of the receiver, pre-emphasized, and fed into a limiting amplifier which prevents over-modulation of the transmitter by audio surges. In the event that a local station identification is to be made, the output of either a microphone or a turntable may be introduced into the system at this point as indicated in Figure 5.

The output of the limiting amplifier is then applied directly to the frequency-modulation exciter which constitutes the UHF aural transmitter modulator. The outputs of the UHF aural and visual transmitters are combined in a notch type cross coupling filter and the resultant signal is used to excite a newly developed high gain UHF television transmitting antenna through a coaxial transmission line. Thus, programs originating from Station WNBT on the present VHF band are received, demodulated, processed and retransmitted on the UHF band.

In addition to the facilities just described, monitoring equipment has been provided which permits observation of either picture or sound quality anywhere within the system as indicated by Figure 5. Also, frequency monitoring equipment has been installed which makes it possible to hold the frequency of the visual and aural carriers well within the tolerances set for standard television broadcasting.

Input Equipment Features

Although certain components in the overall transmitting system are standard units, most contain innovations which are of technical interest and will therefore be described.

The parabolic receiving antenna is located on a platform on the

160-foot level of the antenna tower as shown in Figure 6. It consists of a dipole which is mounted in the focus of a circular, parabolic screen 10 feet in diameter. The parabola has a focal length of 29½ inches and provides an antenna gain of 3.5 decibels as compared to a simple dipole. The receiving antenna transmission line consists of an ATV-225 balanced shielded pair which has a nominal characteristic impedance of 225 ohms and a loss of 2.3 decibels per hundred feet at 50 megacycles. Although the line loss incurred with this cable is greater, the use of shielded transmission line affords a degree of noise immunity not possible with the standard unshielded 300-ohm line.

Fig. 6—Transmitter building (side view) and antenna tower. The Channel 4 parabolic receiving antenna can be seen at the 160 foot level of the tower.

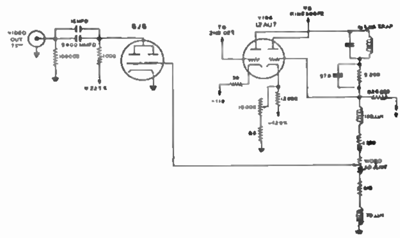

Fig. 7a—Revised video output circuit of Channel 4 receivers.

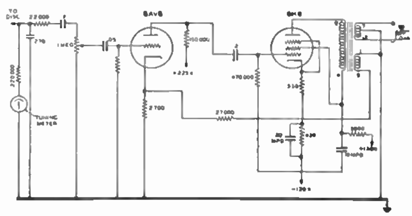

Fig. 7b—Revised audio amplifier circuit of Channel 4 receivers.

The preamplifiers which precede the Channel 4 receivers are fixed tuned amplifiers and are identical with the type SX9A amplifier normally furnished as part of the RCA Antenaplex television distributing system. The amplifier consists of two 6J6 push-pull cross-neutralized amplifier stages in cascade which provide a gain of approximately 20 decibels, as well as an improvement in the signal to noise ratio. The output stages of these amplifiers are equipped with isolation pads which permit a signal to be fed to one or both of the 300-ohm receiver inputs without undesirable interaction.

The Channel 4 receivers used for the actual detection of the Channel 4 signal are standard RCA 9T246 receivers with 10-inch screens which have been specially modified for this particular application. The major changes made in this receiver are modifications of the audio and video output circuits and these are illustrated in Figures 7a and 7b.

The revised video output circuit is shown in Figure 7a. The normal plate load of the first half of the 12AU7 video amplifier has been divided in such a way that the input of the 6J6 isolation amplifier may be shunted across a portion of it without undesirably affecting normal operation of the receiver video circuits. This isolation amplifier will provide an output signal of 2 volts peak to peak at an output impedance of 75 ohms when the "video adjust" and receiver automatic gain controls are properly adjusted.

Figure 7b indicates the modifications made in the receiver audio amplifier. The power amplifier has been altered to provide a standard audio output impedance of 600 ohms and the overall distortion characteristics of the amplifier have been improved by the addition of an inverse feedback loop from a separate winding on the output transformer to the cathode of the penultimate audio amplifier stage. In addition, a tuning meter has been connected across the output of the discriminator which permits adjustment of the fine tuning control of the receiver for minimum audio distortion, which point is indicated by a zero indication on the meter. By this means an audio output of +5 dbm* with but .4 per cent second harmonic distortion can be obtained. The receivers have also been completely shielded, housed in a manner suitable for rack mounting, and have been aligned for optimum reception on Channel 4.

The video stabilizing amplifier employed in the system is a type ND-329 Clamp and Sync Amplifier. It incorporates clamp circuits which stabilize the video signal and contains provisions for the independent adjustment and restoration of sync pulse amplitude so that the video signal fed to the visual modulator will at all times meet FCC standards. The remaining input equipment consists mainly of standard commercial audio amplifiers.

UHF Transmitter

The television transmitter, which constitutes the major link in the transmitting system, is an RCA type TTU-1A. The TTU-1A, which operates in the UHF band, meets the FCC standards for VHF television broadcast transmitters.

The transmitter is housed in six racks, as is shown in Figure 8, the left hand 3 racks containing the aural portion of the transmitter and the right hand 3 racks containing the visual portion. The transmitter includes, as an integral part, a slightly modified RCA type

* Decibels referred to a zero level of 1 milliwatt in 600 ohms.

TT-500B television transmitter which comprises the center two racks in Figure 8.

In the visual section of the transmitter, the radio-frequency output of the 4X150A's in the final stage of the TT-500B is used to drive a tripler employing eight type 4X150A tubes in parallel in a single cavity. The output of the tripler in turn drives a power amplifier containing eight additional 4X150A tubes and employing a cavity design similar to the tripler. These cavities can be seen in the second rack from the right in Figure 8. The video modulator circuits of the TT-500B have been modified to drive a direct-coupled cathode follower stage, consisting of eight 6L6 tubes, which serves as a modulator for the UHF power amplifier.

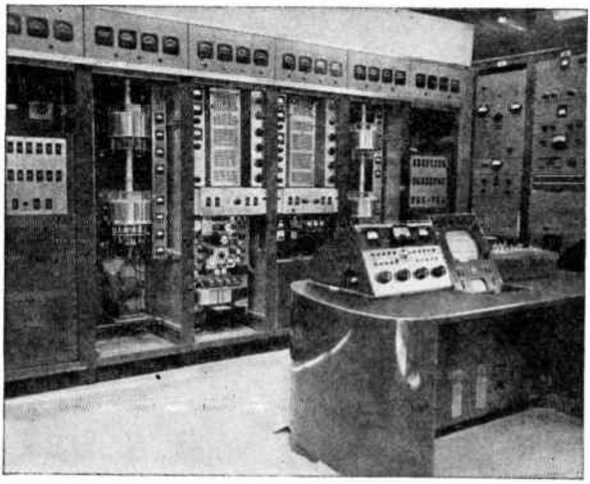

Fig. 8—The RCA TTU-1A transmitter, input and monitoring racks and control console.

In the aural section of the transmitter the arrangement of tripler and power amplifier cavities is identical to that just described, and these are located in the second rack from the left in Figure 8. Modulation, of course, takes place in the frequency modulation exciter which is located in the TT-500B.

The transmitter operates in a standard six-megacycle band, 529 to 535 megacycles, the visual and aural carrier frequencies being 530.25 and 534.75 megacycles respectively. This requires that the TT-500B output stages which drive the UHF triplers operate at 176.75 and 178.25 megacycles. An overall frequency multiplication of 108 occurs between the oscillator and the final frequency in both the visual and aural transmitters. The visual carrier is held within $\pm.002$ per cent of the assigned operating frequency and the sound carrier is automatically maintained 4.5 megacycles above the visual by means of a novel system of frequency control which eliminates relative drift between the visual and aural carriers. This arrangement makes it unnecessary to hold transmitter carrier frequency stability closer than the present FCC requirements for carrier frequency stability in the UHF band even though the intercarrier sound reception feature may be incorporated in future UHF television receivers.

Each of the transmitter sections is designed to work into a 51.5-ohm, 1⅝-inch transmission line, and will deliver normal power outputs of .5 kilowatt aural and 1.0 kilowatt peak visual.

Frequency Monitoring Equipment

The arrangement of frequency and aural modulation monitoring equipment used in conjunction with the UHF transmitter is illustrated in Figure 9. As can be seen, the method used to monitor the transmitter is to convert the transmitter signal frequencies to 45.25 megacycles for the picture carrier and 49.75 megacycles for the sound carrier and then monitor at these frequencies with a standard RCA WF-49A and WF-50A visual frequency monitor and General Radio type 1170-AT FM aural monitor. The converter heterodyning signal is supplied by a standard frequency crystal oscillator at 30.3 megacycles and a multiplying chain. The crystal used in this oscillator is of a special design that has a long term stability of better than 2 parts per million per 30 days. In order to check the actual amount of frequency drift, means have been provided for the calibration of the oscillator against WWV. For this purpose, a 250-kilocycle oscillator is used as a secondary frequency standard and is calibrated against WWV at 5 or 10 megacycles by the use of an external communications receiver. The 250-kilocycle signal frequency is then multiplied and the resultant signal frequency compared to that of the heterodyning signal so that the heterodyne signal can be set to exact zero beat.

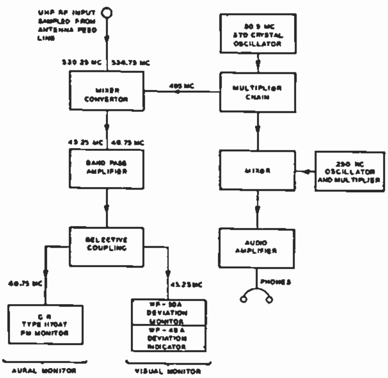

Fig. 9—Block diagram of transmitter monitoring equipment.

Output Network, Transmission Line and Antenna

The coaxial vestigial sideband filter and notch filter indicated in the block diagram of the system (Figure 5) are located behind the

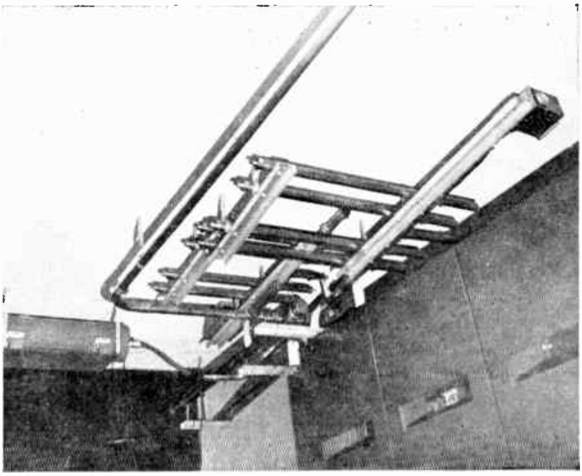

Fig. 10 — Coaxial vestigial sideband filter and diplexer suspended on ceiling behind the transmitter.

transmitting equipment in the station building as shown in Figure 10. The output of these networks is fed to the antenna by means of special 3⅛-inch UHF transmission line pressurized with nitrogen, similar to that employed in the UHF transmission tests in the Washington area.[2,3]

The antenna itself is an RCA type TFU-20A. It is similar in structure to the standard pylon antenna (See Figure 11) but is operated as a slot antenna. The slot radiators are covered by polyethylene covers which are the protrusions which appear on the antenna pole in Figure 11.

The antenna has gain of 17, a diameter of 10¾ inches and an overall height of forty feet above the pole socket in which it mounts. Because of the nature of the antenna structure, efficient operation requires that ice formation over the slot radiators be prevented. A thermostatically controlled de-icing system consisting of a hot air

Fig. 11—The RCA TFU-20A UHF antenna in test position before erection on tower.

[3] D. W. Peterson, "Notes on a Coaxial Line Bead", *Proc. I.R.E.*, Vol. 37, No. 11, p. 1294, November, 1949.

blower has therefore been installed to prevent icing conditions and losses due to extreme humidity. The output of the blower is forced into the antenna at the base of the antenna pole and is exhausted at the top of the antenna just below the beacon. The electrical power input to the blower heaters can be increased to as much as 14 kilowatts if conditions demand. The overall height above ground of the antenna and antenna tower is 250 feet and the overall height above mean sea level is 440 feet. The height of the antenna radiation center above average terrain is 330 feet.

SYSTEM EFFICIENCY

The overall system efficiency from the transmitter to the antenna is approximately 80 per cent. With an antenna gain of 17 and a visual transmitter power output (peak) of one kilowatt, the effective radiated power of the system will be roughly 13.9 kilowatts peak visual. It is estimated that the power costs for the entire system for 16 hours per day, 7 days per week operation would be roughly $4000 per year.

Additional developments are under way to increase the efficiency of the various UHF system components and raise the antenna gains. It would be possible, for example, to employ wave guide operating in the $TE_{0,1}$ mode for the antenna feed system. Such a wave guide would have dimensions of roughly $15 \times 9\frac{1}{2}$ inches and would have an efficiency of about 92 to 93 per cent for the length of guide necessary.

FIELD TESTS

Experimental television station KC2XAK went on the air with full power and full modulation on December 29, 1949. A number of television receivers and converters designed to receive KC2XAK are being installed in homes in the Bridgeport area. Field tests will include observations in homes throughout the service area, at distances and under conditions which will determine the extent of coverage of the station. Various types of receiving antenna will be tested, shadow areas and multipath problems investigated and extensive field intensity measurements will be made.

Such measurements will be taken at representative receiver locations and will include actual voltages obtained at receiver terminals. In addition, measurements will be conducted along various radials and an investigation made of field intensity versus receiving antenna height under various conditions. Upon completion of the surveys, the results obtained will be disclosed to the FCC and the industry in subsequent papers.

ULTRA-HIGH-FREQUENCY ANTENNA AND SYSTEM FOR TELEVISION TRANSMISSION*

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Summary—An omni-directional, horizontally polarized ultra-high-frequency antenna with a power gain of 17.3 is described. Some of the performance characteristics and development problems are given. A vestigial sideband filter and notch diplexer constructed of concentric transmission line is described. Measured performance characteristics and design considerations are also given. The concentric transmission line used to feed the transmitter output to the antenna is described briefly. A broad frequency characteristic is obtained by use of compensated undercut insulators supporting the inner conductor. Waveguide, elbows, and transitions to concentric line have been developed for operation in the 500 to 750 megacycle band. Considerations which make waveguide attractive for ultra-high-frequency television service and measured characteristics of the components developed are presented. The antenna system was developed for the operation at ultra-high-frequency experimental television station KC2XAK at Bridgeport, Connecticut. The antenna system involves the antenna, transmission line or waveguide, a notch diplexer and vestigial sideband filter.

Antenna

An effective radiated power of 10 to 20 kilowatts was desired for the experimental ultra-high-frequency (UHF) television broadcast station at Bridgeport, Connecticut.[1] A study of the problems involved indicated the antenna should have as much gain as practical, consistent with stability of radiated signal. Deflection of the antenna and supporting tower by high winds causes the vertical pattern of the antenna to tilt from the horizontal. This may cause the received signal to decrease to an unacceptable value in some receiver locations during high winds. Previous experience and study of the vertical pattern accompanying a given gain indicated a gain of 20 to 25 would give acceptable coverage. Experience obtained

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* Decimal Classification: R326.81×R117.2.

† The material in this paper for which the author is directly responsible, together with certain other material, forms the basis of a Doctor's thesis in Electrical Engineering to be presented to the faculty of the Graduate School of the University of Pennsylvania.

[1] R. F. Guy, J. L. Seibert and F. W. Smith, "Experimental Ultra-High-Frequency Television Station in the Bridgeport, Connecticut Area," *RCA Review*, Vol. XI, No. 1, pp. 55-67, March, 1950.

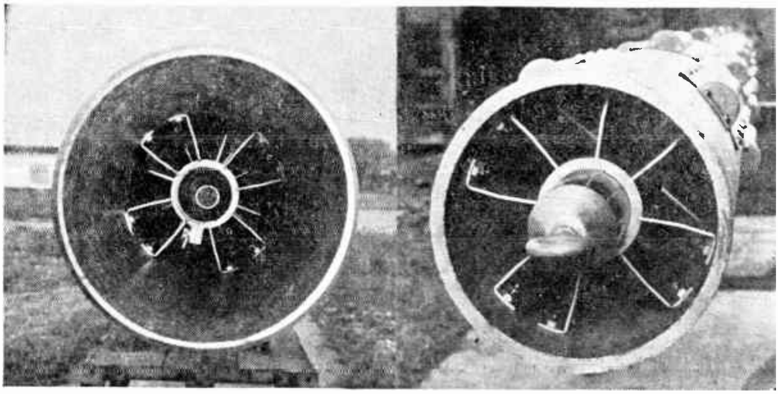

Fig. 1—End views of the antenna.

with the antenna herein described, which has a power gain of 17.3, will determine the practicability of higher power gains.

Very-high-frequency television transmitting antennas utilize a branching-type feeder system to each radiating element to obtain the desired bandwidth and pattern characteristics. An extension of this practice to UHF high-gain antennas would result in a very complicated feeder system due to the large number of radiating elements involved. In the case of the antenna constructed for Bridgeport, the radiating system would have required 88 branch lines; this clearly was not desirable and a feed system of greater simplicity was developed.

The antenna consists of a 3⅛-inch outside diameter copper tube mounted coaxially inside a 10¾-inch outside diameter, ½-inch wall steel tube (Figure 1). The steel

Fig. 2—Antenna installed in the normal vertical position in a rotating socket for measuring horizontal patterns.

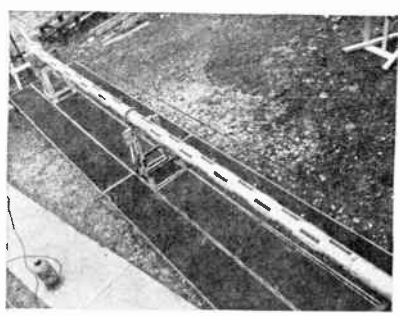

Fig. 3—Antenna mounted on horses horizontally over space cloth to absorb ground reflections.

tube is in two lengths, each approximately 20 feet long joined by means of flanges at the radiation center of the antenna (Figures 2 and 3).

Each layer of the radiating system consists of four half-wave slots equally spaced around the circumference of the steel tube. There are twenty-two such layers of half-wave slots, making eighty-eight individual slots in the steel tube.

The antenna is divided into two electrically identical groups of half-wave slots, the slots being spaced approximately a half wave length between ends or approximately a full wave length between centers. The upper and lower groups of slots consist of eleven layers each with a space of 1.66 wave lengths between the ends of the two groups. Each successive layer of slots is rotated 45 degrees to suppress transmission of the $TE_{1,1}$ and other noncylindrical modes within the steel cylinder. The modes which do not have cylindrical symmetry would cause unequal excitation of individual slots in a layer resulting in a non-circular horizontal pattern. Horizontal patterns deviating considerably from circular were observed during development work and the staggering of the layers was found to eliminate such variations from circular. It was interesting to observe, when only one layer was driven, that the pattern was closely circular, checking theoretical calculations.[2]

The 3⅛-inch copper tube installed within the antenna tube acts as a transmission line to distribute the transmitter output to the various layers of slots. The slots are driven by radial probes fastened within the antenna tube on one edge opposite the center of each slot. The current passing through the probe capacity also passes through the driving point impedance of each radiating slot. A set of radial probes is used between each layer to obtain an impedance bandwidth at the input of each succeeding layer of the antenna which is approximately equal to the bandwidth of the end layers.[3] A similar antenna may be constructed without the tuning probes between layers. However, it is observed that the system input bandwidth becomes progressively narrower as the number of layers, and consequently the gain, is increased.

The 3⅛-inch diameter copper tube within the antenna has an inner conductor within the lower half which serves as a transmission line to carry the transmitter output to the center of the antenna feed system. The center feed avoids any tilt or other dissymmetry in the vertical pattern with changes in frequency which would be character-

[2] G. Sinclair, "Patterns of Slotted Cylinder Antennas," *Proc. I.R.E.*, Vol. 36, No. 12, p. 1487, December, 1948.

[3] H. J. Riblet, "Microwave Omni-directional Antennas," *Proc. I.R.E.*, Vol. 35, No. 5, p. 477, May, 1947.

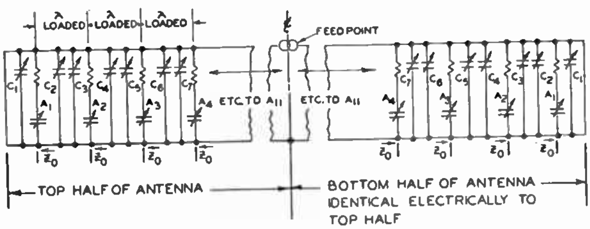

Fig. 4—Diagram of electrical distribution system.

istic of an end fed broadside array. The feed point may be shifted from the center of the array if desired to produce a phase difference of the currents in the upper and lower half of the antenna. Any such phase difference between currents in the two halfs of the antenna is accompanied by a corresponding tilt in the vertical pattern. This adjustable tilt in vertical pattern may be used to advantage in any particular installation to adjust for particular terrain conditions or cover densely populated valleys, etc.

Figure 4 is a line diagram illustrating the electrical operation of the feed system used in the antenna. The tuning and feed probe settings for successive layers are different. The probe settings required for matched impedances in the corresponding layers of the top and bottom half of the antenna were usually different due to small mechanical variations in the antenna structure.

A study of the characteristics of the antenna was made using a microwave model at approximately 5000 megacycles (Figure 5). A matched load was inserted in the end of the microwave antenna tube. The relative magnitude and phase of the voltage across the load was measured by conventional methods. Phase and magnitude measurements made with one layer of the antenna matched by means of the tuning probes for various settings of the slot probes gave a curve of additional phase retardation versus relative power absorbed by the radiating layer. This curve was used to determine the proper spacing between layers to compensate for the additional phase shift introduced by the tuning probes. The proper amount of power absorbed by each layer of slots relative to the power transmitted to the subsequent layers is determined as follows for the eleven layers in the top and bottom halves of the antenna.

The end layers, top and bottom extremity of the complete 22-layer antenna, have no succeeding layers to which power must be trans-

Fig. 5—One-tenth scale microwave model of one-half of the antenna (top or bottom half).

mitted. Consequently, they must absorb all of the power which is contained in the incident wave. This is accomplished by adjusting the four slot feed probes and the shorting plug at the end of the antenna until this layer matches the characteristic impedance of the concentric line within the antenna, which is approximately 68 ohms. The second layer from each end must absorb ½ of the power and transmit ½ of the power to the end layers. A combination of feed probe and tuning probe settings is then found which will give this power distribution; i.e., reduce the voltage on the matched load in the end of the antenna to .707 of the value obtained when no tuning or feed probes are inserted. The tuning probes must be adjusted for an impedance match each time the feed probe setting is changed. The setting of the feed probe for the third layer transmits ⅔ of the incident power and absorbs ⅓, the fourth layer absorbs ¼ and transmits ¾ of the power etc. to the 11th layer which absorbs $1/11$ of the power and transmits $10/11$. The settings of the feed and tuning probes and the spacing between each layer of slots determined by this experiment were made and the vertical pattern and gain of the model were measured. The measured vertical pattern closely resembled the expected pattern for 11 layers in phase with equal currents. The power gain measured by substitution of a standard horn reference for the microwave model was 11.4. This checked the gain obtained by integration of Poynting's vector obtained from the measured vertical pattern.

The spacing of successive layers of the full scale UHF antenna was obtained by scaling the dimensions of the microwave model.

The tuning probe locations and settings could not be successfully scaled from the model, and it was necessary to repeat the work of adjusting the tuning probes individually for each layer. The probes were adjusted to obtain an impedance match leaving the slot feed probes set at the dimension determined by scaling the model. Since the magnitude and phase of the currents in each layer of the full scale antenna were not checked, some degradation of vertical pattern and gain of the full size UHF antenna resulted. The gain of the full size antenna was 17.3 instead of 23 or 24 which would have been obtained if the individual layers were directly adjusted for equal phase and currents. The variation of results between the model and full size antenna is attributed to mechanical variations between the model and the full size antenna. It has been found quite difficult to duplicate, electrically and mechanically, all of the practical requirements of the full scale antenna in a one-tenth scale model.

Each layer of the full scale antenna was adjusted to obtain an impedance match at 531 megacycles. Difficulty in adjustment of the entire antenna by adjusting each successive layer from each end simultaneously led to the discovery that the tuning probe settings for corresponding layers from each end were not the same if the input impedance was matched. The proper reference conditions on the slotted measuring line for each half of the antenna were obtained by shorting out one-half of the antenna a wave length from the feed point and substituting a matched load in the unshorted half of the antenna after all feed probes and tuning screws had been removed. The matched load was constructed of four radial fins of phenolic laminate about 10 feet long mounted on a sleeve which would permit the fins to slide freely on the 3⅛-inch diameter inner conductor of the antenna and just clear the inside of the outer steel tube. The fins were covered on both sides with 377 ohms per square space cloth and had a linear taper on the input end about three wave lengths long. When the load was adjusted for a match, the input impedance to the antenna feeder system did not change when the load was moved on the 3⅛-inch inner conductor. After the reference conditions were obtained with the matched load, it was possible to tune each half of the antenna either by using the matched load in one half and tuning the other half or shorting out the one half with the shorting disk installed at a point located an integral number of half waves from the feed point.

Experiments with the microwave model and a large sheet of metal to simulate a perfect conducting earth indicated an error in free space impedance measurements of about ± 6 per cent would occur if the UHF antenna were mounted horizontally 6 feet above the earth's surface. This was considered undesirable, and experiments using the microwave model with space cloth having the impedance of free space (377 ohms per square) and a metal ground sheet with space cloth equivalent to a full size sheet of space cloth (9×40 feet) placed under the antenna ¼ wavelength above the ground plane indicated that impedence measurements could be made with the horizontal antenna above the ground if the 9×40 foot space cloth absorber were placed ¼ wave above the ground. The full size space cloth absorber was constructed on a group of wooden frames with space cloth on the top and wire screen on the bottom to assure terminating the wave in a good conductor placed at the earth's surface. The setup of the antenna for impedance tests over the space cloth absorbers is shown in Figure 3. Later impedance measurements on the horizontal antenna 6 feet over plain earth without the ground reflection absorbing sheets indicated the ground may not affect the impedance as much as expected (see Figure 6).

Calculations of reflection coefficient of the earth's surface for

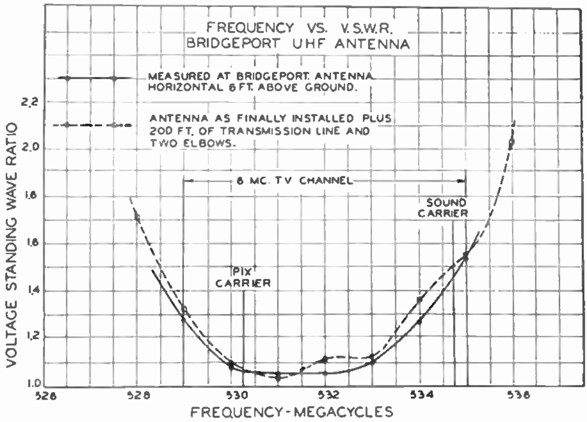

Fig. 6—Standing-wave-ratio characteristic of the antenna.

normal incidence using a dielectric constant of 16 for earth gives a reflection coefficient of —.6 compared to the assumed perfectly conducting earth reflection coefficient of —1. If this were the case, one might expect the error in impedance measurements, due to the ground 6 feet distant, to be in the order of ± 3.6 per cent. Observations have indicated that it is probably less than this.

Vertical patterns were measured by horizontally rotating the antenna while mounted on a vertical spindle under the center of radiation as shown in Figure 7. A receiver was set up at a distance sufficiently great from the transmitting antenna that the transmission path length difference to the receiving antenna from the ends and from the center of the transmitting antenna was less than one-tenth wave length. The observer at the receiver setup was in constant communication with the engineer at the transmitting antenna (see right inset, Figure 7). An accurate compass dial and vernier were installed on the transmit-

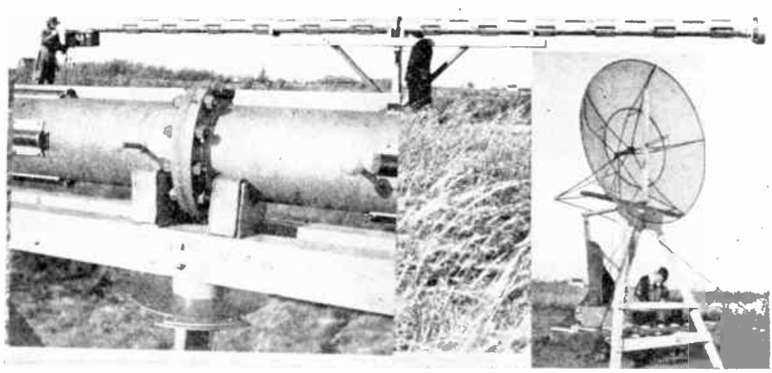

Fig. 7—Vertical pattern test setup for the antenna.

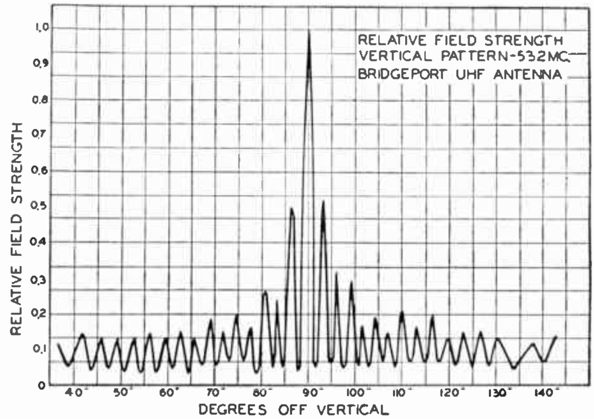

Fig. 8—Vertical pattern measured for the antenna.

ting antenna spinner to indicate the azimuth with better than .1 degree accuracy, as the transmitting antenna was rotated by hand and the signal recorded point by point. The vertical pattern was measured at 529, 532 and 535 megacycles and found to remain about as shown in Figure 8. Poynting's vector integrations using the measured vertical patterns, give the gain frequency characteristic shown in Figure 9. The vertical pattern calculated from the eleven section microwave model pattern measurements is shown in Figure 10. The greater side lobe levels of the full scale antenna and the deviation of the null angles compared to those obtained with the microwave model are largely due to variations of the amplitude and phase of the currents in the various layers of the full size UHF antenna. The side lobe level of the full size antenna may be reduced to compare with the microwave model if the current magnitude and phase of each layer are correctly adjusted.

Calculations were made to determine what transmitter power would give an adequate received signal for the vertical pattern measuring set-up used (Figure 7). Based on the premise that the ground reflection coefficient near grazing incidence was —1, these calculations indicated that a transmitter power in the order of 100 watts would be required. Experiments with the actual test setup and location indicated adequate received signal could be obtained with about 2 milliwatts transmitter power. This suggests the ground reflection coefficient including the effects of brush, grass

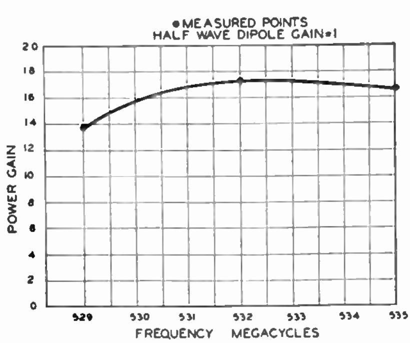

Fig. 9—Power gain versus frequency.

and irregular terrain is nearly zero or is actually positive (see terrain of Figure 7).

The horizontal pattern was measured in the setup shown in Figure 2. The antenna was mounted vertically in a socket and could be readily rotated. The receiver and antenna were mounted at a distance, opposite the radiation center of the transmitting antenna. The measured horizontal pattern varied less than ± 1 per cent from circular, which is in close agreement with the pattern calculated for staggered slots.[2]

Transmission Line

The commercial very-high-frequency, air dielectric, concentric transmission line has a uniform tubular inner conductor with supporting ceramic disc insulators clamped to it at one-foot intervals. This type of line is satisfactory for very-high-frequency service where the characteristic impedance is practically constant. However, the uncom-

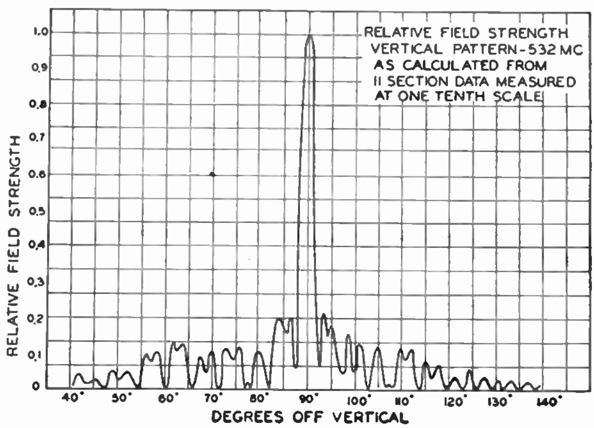

Fig. 10—Vertical pattern for full antenna calculated from measurements on the microwave model.

pensated beads at periodic intervals make the line have characteristics similar to a low pass filter with attenuating bands in the UHF region. The characteristic impedance fluctuates rapidly with frequency and is reactive over much of the UHF band. A line of this type is clearly not suitable for the stringent standing wave requirements of television transmitting antenna systems in the UHF band where a nearly constant resistance load must be used.

There appear to be many approaches to the UHF insulator support problems[4, 5] for concentric line. The method developed by D. W. Peter-

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[4] R. W. Cornes, "A Coaxial Line Support for 0-4000 Mc," *Proc. I.R.E.*, Vol. 37, No. 1, p. 94, January, 1949.

[5] D. W. Peterson, "Notes on A Coaxial Line Bead," *Proc. I.R.E.*, Vol. 37, No. 11, p. 1294, November, 1949.

son was used in the Bridgeport installation because satisfactory line of this type was available. The transmission line in the installation changed the standing wave characteristic of the antenna very little, as indicated by measurements, taken with the antenna mounted on the tower, through 200 feet of transmission line (Figure 6).

The insulator supports on the transmission line used at Bridgeport were mounted in undercut spaces on the inner conductor. Small series inductances were cut in the faces of the undercut to compensate for the step capacity[6,7] as shown in Figure 11. The standing wave characteristic of a sample of the transmission line used is shown in Figure 12. The line was terminated in a resistance of 52 ohms and measured on a 52-ohm measuring line.

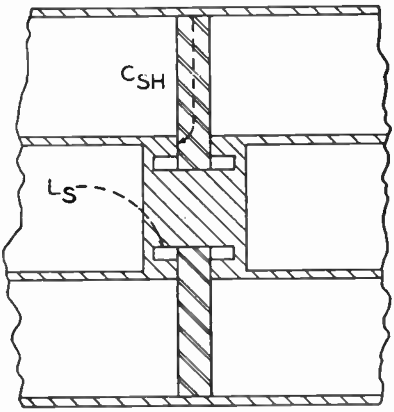

Fig. 11 — Section of compensated undercut insulator for 3⅛-inch diameter coaxial line.

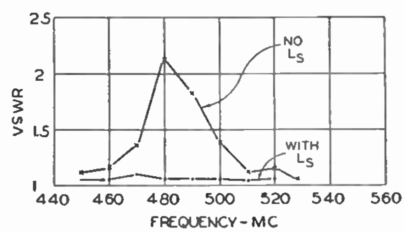

Fig. 12—Standing wave ratio of sample run of UHF transmission line.

Wave Guide

The simplicity and relatively low loss of wave guide led to its consideration for experimental line to transmit power from the transmitter to the antenna. Various wave-guide components were developed for this purpose including transitions for terminating the wave guide in coaxial lines and E and H plane bends (Figures 13, 14, 15 and 16). The problems of using wave guide near the lower end of the UHF band are new, and considerable development work and experimentation is required before it can be utilized commercially.

The wave guide investigated is rectangular having inside dimensions $7\frac{1}{2} \times 15$ inches. It is fabricated from sheet metal. Hot dipped galvanized steel, aluminum or copper clad steel are suitable materials.

A comparison of the loss of wave guide and common size concentric

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[6] HANDBOOK OF DESIGN DATA, Brooklyn Polytechnic Institute, Report No. R-158-47.

[7] J. R. Whinnery, H. W. Jamieson and T. E. Robbins, "Coaxial Line Discontinuities," *Proc. I.R.E.*, Vol. 32, No. 11, p. 695, November, 1944.

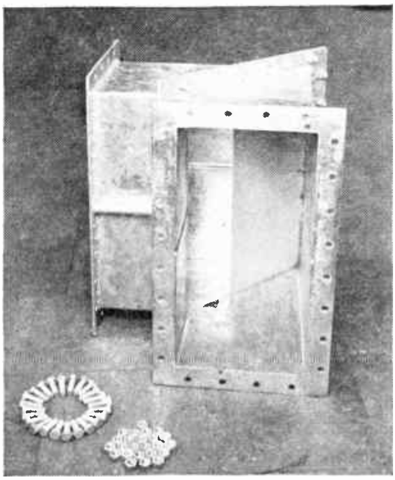


Fig. 13—E plane bend for 7½ × 15-inch wave guide.

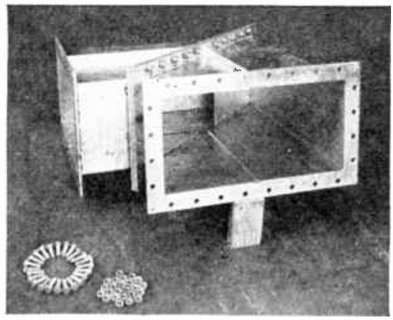


Fig. 14—H plane bend for 7½ × 15-inch wave guide.

transmission lines is given in Figure 17. The standing wave frequency characteristics of some of the wave-guide components developed for experimental work are shown in Figures 18, 19, 20 and 21.

The wave guide and components developed correspond closely to those employed in current microwave practices. However, the large size created new electrical and mechanical problems. It was convenient to do much of the development work on the wave guide and components using ½ and $\frac{1}{10}$ scale models at higher frequencies. No difficulties similar to those observed in scaling the antenna were experienced, since the wave guide was so simple that all important features could be accurately scaled.

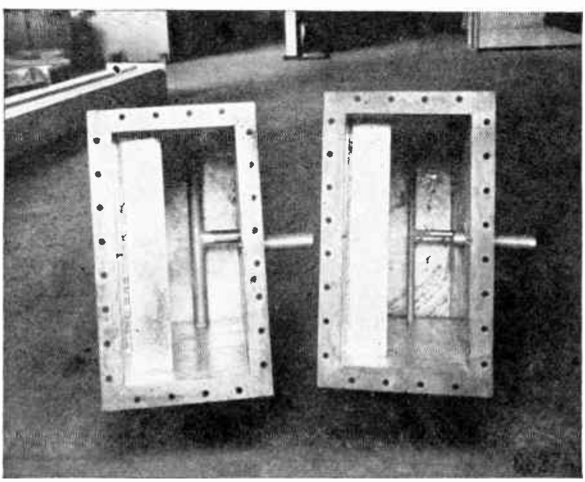


Fig. 15—Transitions from 7½ × 15-inch wave guide to coaxial line.

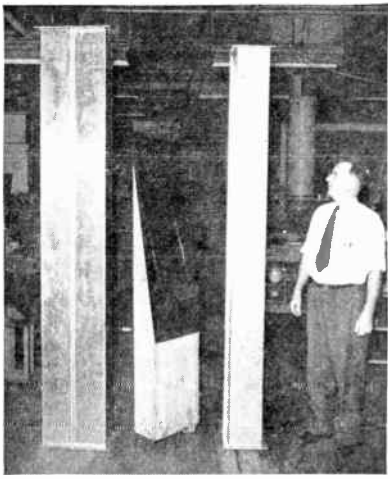


Fig. 16—Straight wave-guide section 7½ × 15 inches.

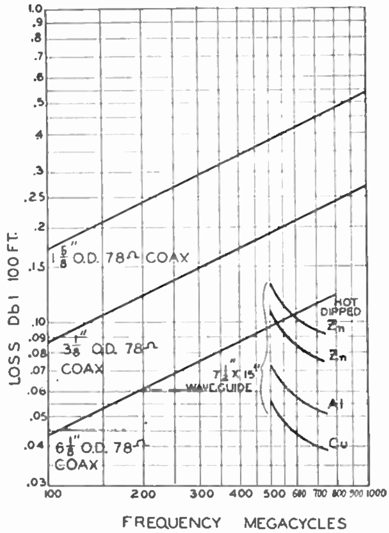


Fig. 17 — Curve showing loss of various 7½ × 15-inch wave guides and standard size coaxial lines.

Vestigial Sideband Filter

A high-level vestigal sideband filter for use on the output of the transmitter was developed. The filter used was designed to present a constant input resistance for picture carrier and both sidebands, although only the upper sideband and a portion of the lower sideband is transmitted. The transmission characteristic of the sideband filter installed at Bridgeport is shown in Figure 22.

The sideband filter was constructed of coaxial transmission line (Figure 23). A line diagram of the circuit arrangement used is shown in Figure 24. Some of the resonant

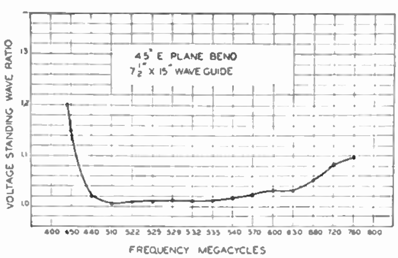


Fig. 18—45-degree E plane bend standing-wave ratio versus frequency.

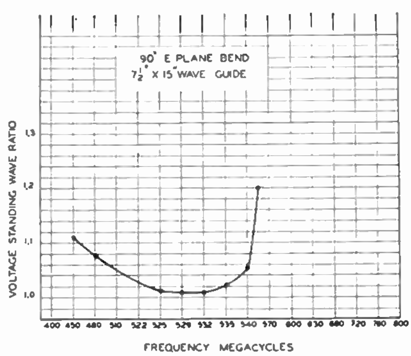


Fig. 19—90-degree E plane bend standing-wave ratio versus frequency.

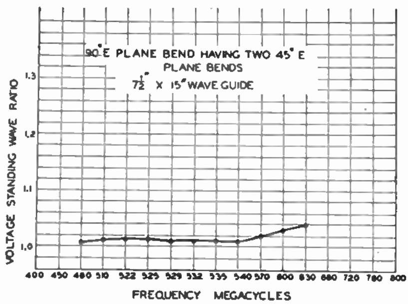


Fig. 20—90-degree E plane bend consisting of two 45-degree bends —standing-wave ratio versus frequency.

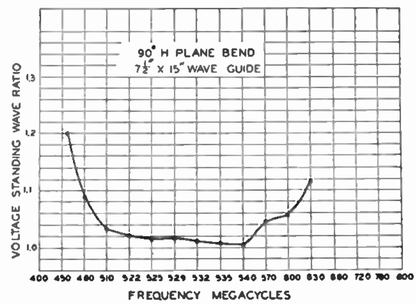
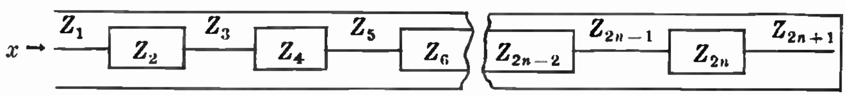


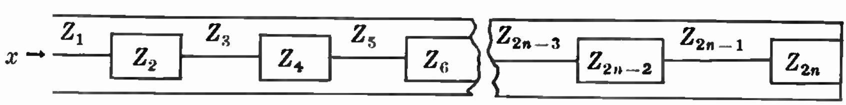
Fig. 21—90-degree H plane bend standing-wave ratio versus frequency.

circuit elements are constructed of stepped quarter-wave sections to obtain a high reactance slope without using long transmission lines. The reactance slope obtained with a stepped open-end transmission line is:



$$\frac{dx}{df} = \frac{\pi Z_1}{2f_o}\left\{1+\frac{Z_1}{Z_2}\left[1+\frac{Z_3}{Z_2}\left(1+\frac{Z_3}{Z_4}\left\{1 + \frac{Z_5}{Z_4}\left[\cdots\left(1+\frac{Z_{2n+1}}{Z_{2n}}\right)\right]\right\}\right)\right]\right\}. \quad (1)$$

The equation for the reactance slope of a stepped shorted-end transmission line is obtained by letting Z_{2n+1} equal zero in Equation (1):



$$\frac{dx}{df} = \frac{\pi Z_1}{2f_o}\left\{1+\frac{Z_1}{Z_2}\left[1+\frac{Z_3}{Z_2}\left(1+\frac{Z_3}{Z_4}\left\{1 + \frac{Z_5}{Z_4}\left[\cdots\left(1+\frac{Z_{2n-1}}{Z_{2n}}\right)\right]\right\}\right)\right]\right\}. \quad (2)$$

All steps in impedance are a quarter-wave long at the resonant frequency.

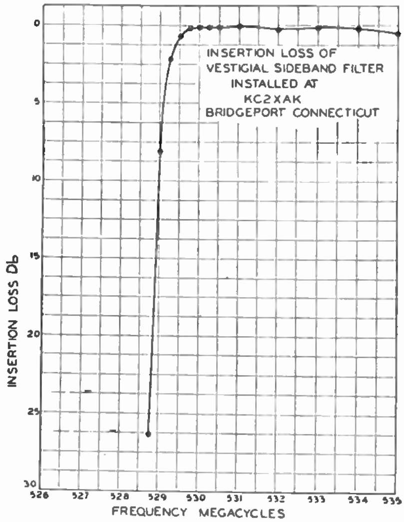


Fig. 22—Sideband filter insertion loss versus frequency.

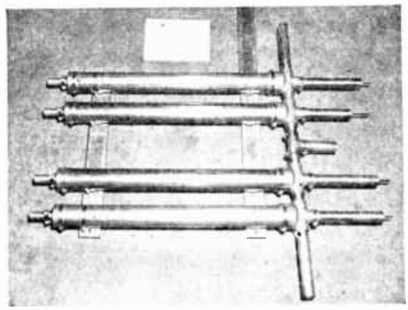


Fig. 23—Sideband filter photograph.

The sideband filter is designed to have five points of input impedance match. One is at the crossover frequency, where the output to the antenna is reduced 3 decibels. This is obtained by making the input impedance of the low pass and high pass portion of the sideband filter complementary. The normalized conductance of the input to the low pass and high pass sections are equal to + ½ at the crossover frequencies. The normalized susceptances are equal and opposite in sign, thus making the input admittance to the filter system a normalized conductance of one with zero susceptance at crossover. The rejector circuits in the high pass portion of the filter are parallel resonated at one frequency $f_0 + \Delta_2$ in the pass band. The impedance is matched at a frequency very close to this parallel resonant frequency. The rejector circuits in the low pass portion of the filter are similarly parallel resonated at a frequency $f_0 - \Delta_2$ in the reject band resulting in a similar matched impedance point in the reject band. The reject circuits in the high pass and low pass portions of the filter are tuned to different frequencies. Proper selection of the reactance slopes and reject frequencies causes one rejector circuit to compensate for the mismatch introduced by the other to obtain an additional matched impedance point in the pass and reject band. The ideal transmission characteristic of

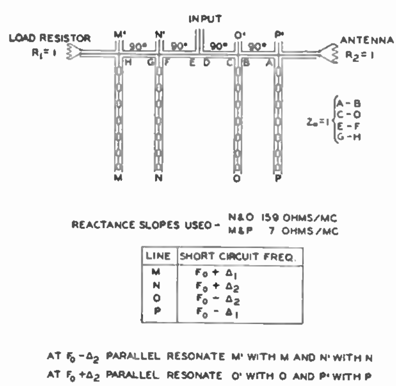


Fig. 24—Sideband filter line diagram.

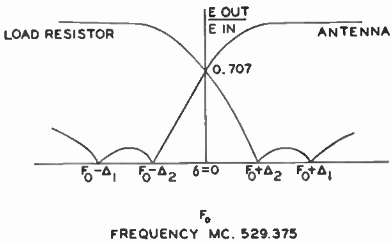


Fig. 25—Sideband filter ideal response.

the vestigial sideband filter is shown in Figure 25.

Notch Diplexer

A notch diplexer is a filter circuit which is required to feed the sound and picture transmitter output into the single antenna transmission line without interaction between the transmitter outputs. A coaxial line circuit was developed for the Bridgeport installation (Figure 26). A line diagram of the circuit arrangement used is shown in Figure 27. The circuit used does not differ greatly from the notch diplexers used for very-high-frequency television.

The operation of the notch diplexer is as follows, referring to Figure 27: The lines of length $11\lambda/4$ are tuned for a short circuit at picture frequency on junctions D and E. The short circuit at picture

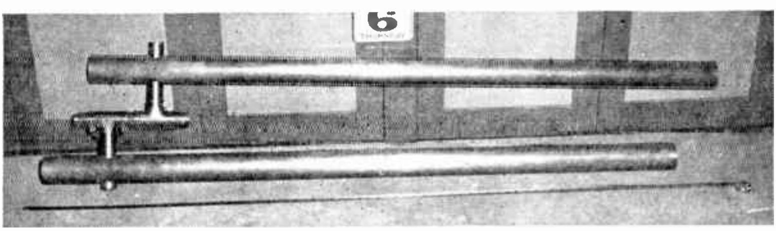


Fig. 26—Notch diplexer photograph.

frequency for junction D is transformed to an open circuit at junction F by a quarter-wave line DF. The open circuit at junction F permits the picture transmitter output to go to the antenna. The line CA is selected a quarter-wave at picture carrier and therefore does not shunt junction C at picture carrier. The length between junction F and C is selected to obtain a good impedance match for picture fre-

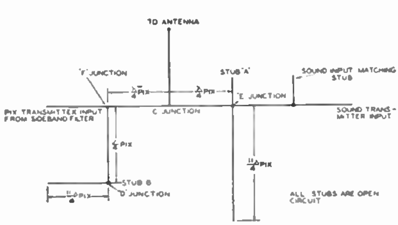


Fig. 27—Notch diplexer line diagram.

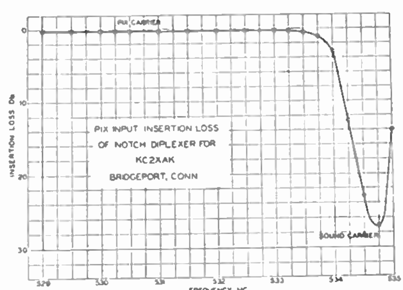


Fig. 28—Notch diplexer insertion loss.

quencies just below the sound notch thus obtaining a sound notch with steep skirts and a sharp shoulder. Stubs A and B are adjusted so that when each $11\lambda/4$ circuit is attached at junction D, maximum rejection of sound frequency on the picture input exists. This adjustment does not correspond to minimum standing wave ratio on the sound input and an additional stub is used on the sound input line to obtain a matched input at sound carrier frequency. The standing wave ratio and transmission characteristics obtained on the picture input for the coaxial line diplexer installed at Bridgeport, Connecticut are shown in Figures 28 and 29.

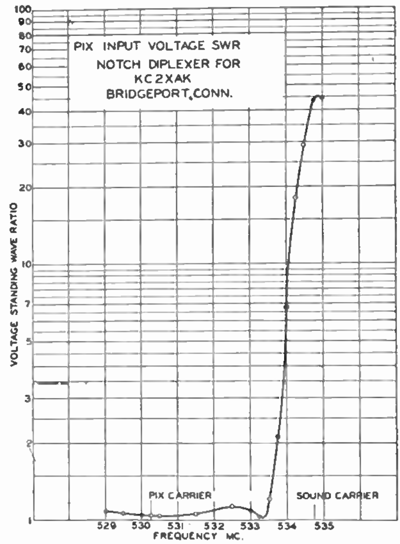


Fig. 29—Notch diplexer standing-wave ratio versus frequency.

Conclusion

The various components of the UHF television antenna system have been constructed and tested. The practicability of UHF television transmitting antenna equipment has been demonstrated by the experimental test results obtained and will be confirmed by operational experience with the experimental UHF Television Station KC2XAK in Bridgeport, Connecticut. Adequate power gain can be realized in a practical structure to permit adequate service with low power transmitters. The antenna development indicated no fundamental obstacles in the design of a practical commercial antenna for ultra-high-frequency television. The antenna described is a prototype of the improved type TFU-20A commercial ultra-high-frequency transmitting antenna.

Acknowledgment

The developments described were the contributions of many engineers of the RCA Victor Division, RCA Laboratories Division and the National Broadcasting Company, Inc.

A NEW ULTRA-HIGH-FREQUENCY TELEVISION TRANSMITTER*

By

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Camden, N. J.

Summary—A description is given of a new television transmitter which operates in the proposed ultra-high-frequency band at a frequency of 529-535 megacycles with a power output at the peak of the synchronizing pulses of 1000 watts. Several novel circuits have been developed, including multi-tube cavities operating as tripler and power amplifier stages. The transmitter conforms to all the standards pertaining to transmitters operating in the very-high-frequency band and is comparable in performance to present day very-high-frequency transmitters.

Introduction

The increasing need for additional television channels placed greater emphasis on the utilization of the ultra-high-frequency channels which have been proposed by the Federal Communications Commission. Practical employment of these frequencies was hampered by the lack of operational experience which could only be obtained by scheduled operation at power levels capable of delivering a useful signal in an urban area. While propagation tests and demonstrations were conducted during the Summer of 1948, it became obvious that further tests under practical operating conditions would be required to demonstrate the suitability and possible limitation of television broadcasting in these channels. It was decided that an output power of the order of one kilowatt would be satisfactory for this purpose.

In November 1948, the development of such a television transmitter was undertaken. The frequency was to be 529-535 megacycles, the power output was to be of the order of one kilowatt using currently available tubes, and the design was to be such that the transmitter could be used in commercial operation and would conform with present very-high-frequency (VHF) television standards. Thirteen months later, on December 29, 1949, the completed transmitter, designated TTU-1A, went on the air at Bridgeport, Connecticut.

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* Decimal Classification: R583.4 × R310.

The first consideration in the development of this transmitter was to find a type of tube suitable for the job. In the interest of expediting the design it was decided to employ multiple operation of readily available tubes. Two tubes which offered possibilities were the 4X150A and the 5588. To make comparison tests, single tube cavities were built, and the characteristics of each tube were measured at 530 megacycles. The final decision was in favor of the 4X150A for two reasons. First, the 4X150A, a tetrode, would have a higher stage gain and be easier to modulate than the 5588 triode. Second, the 4X150A was much more stable than the 5588 which would definitely require neutralization. Even though it might become necessary to neutralize the 4X150A, neutralization of a tetrode would be much simpler and much less critical in practical operation.

The second consideration was to determine the type of radio-frequency circuits to be used. The most straightforward way would be to use several single tube cavities and combine their outputs by suitable passive networks. Another possibility would be to arrange the tubes around a circle in a single cavity and drive them in parallel. A test cavity was built to see if the proper mode could be set up in such a cavity. The encouraging results obtained led to the decision to use this type of circuit.

A tentative radio-frequency tube lineup was decided upon, and from these basic decisions, the TTU-1A transmitter was developed.

Description

The TTU-1A transmitter is housed in six cabinets fastened together to form a single assembly as shown in Figures 1 and 2. On the left is the aural section and on the right is the visual section.

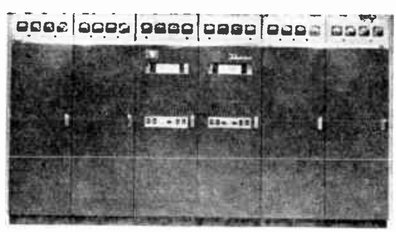


Fig. 1—Front view—doors closed.

The TTU-1A is built around the RCA TT-500B Television Transmitter, a commercial transmitter operating on channels 7 to 13 with a peak visual power output rating of 500 watts and an aural power output rating of 250 watts. The TT-500B serves as a driver for the ultra-high-frequency stages and the video modulator as shown in block diagrams Figures 3 and 4. The visual radio-frequency chain comprises a crystal oscillator and four multiplier stages followed by an amplifier all of which are part of the TT-500B transmitter. These in turn drive an eight-tube tripler consisting of eight type 4X150A tubes in parallel in a single cavity. The power amplifier consists of eight type 4X150A tubes in a similar cavity. The output of the power amplifier is fed to the antenna through a reflectometer which measures the incident and reflected waves on the transmission line and thus indicates the standing wave ratio. The reflectometer is calibrated in peak output power so that the output of the transmitter may be continuously metered.

The video modulator chain consists of three low-power video amplifiers, including a clamp circuit, all located in the TT-500B transmitter and direct coupled to the modulator through a regulated "bucking bias" power supply to provide the necessary bias to the modulator. The modulator consists of eight type 6L6 tubes fed in parallel and connected as cathode followers with the output of each tube directly coupled to the grid of a single power amplifier tube. A video monitor amplifier is supplied to provide the necessary phase reversal for viewing on a monitor tube.

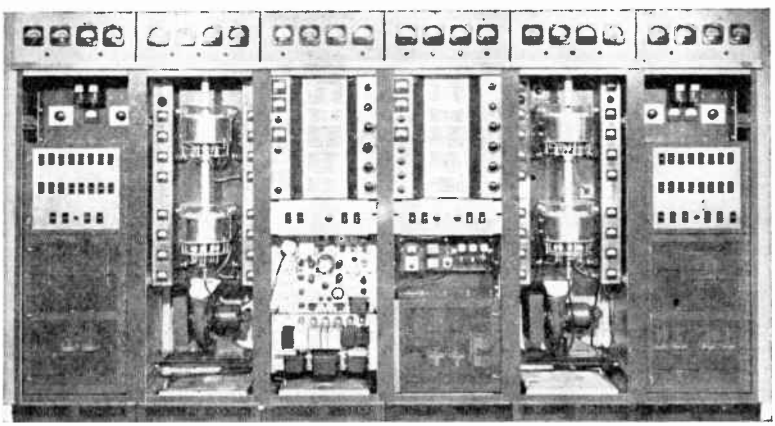


Fig. 2—Front view—doors removed.

The aural radio-frequency chain comprises the frequency-modulation exciter, to be described later, followed by two multiplier stages and an amplifier, all located in the TT-500B transmitter. The output of the TT-500B, which is at one third carrier frequency is followed by an eight-tube tripler cavity and an eight-tube power amplifier cavity identical to those used in the visual transmitter.

The tripler and power amplifier cavities are designed as plug-in units so that in the event of the failure of any component other than a tube, the faulty cavity may be removed and replaced by a spare cavity.

The control circuits are conventional except that indicators are provided to assist in the rapid location of a faulty tube in the cluster. Individual circuit breakers are connected in the cathode circuit of each tripler and power amplifier tube so that, in the event of an over-current

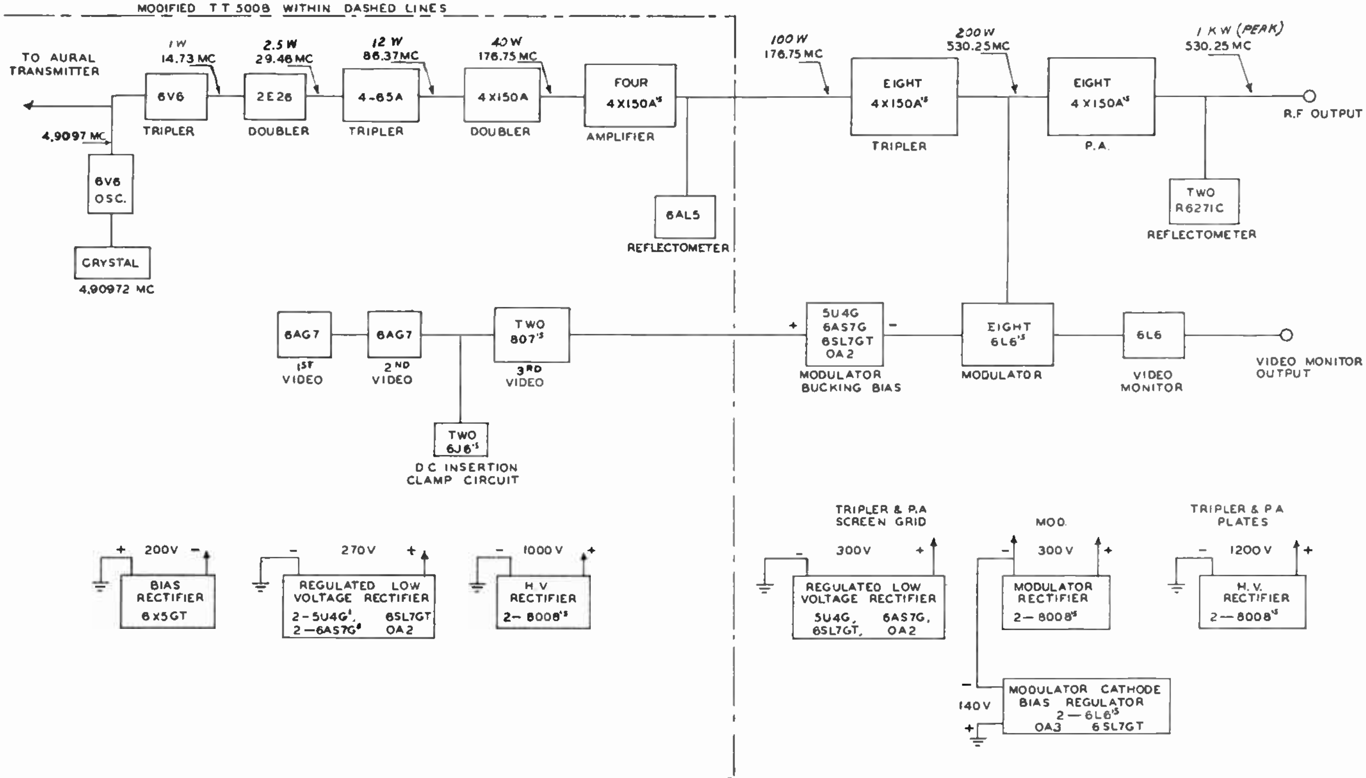


Fig. 3—Block diagram of visual transmitter.

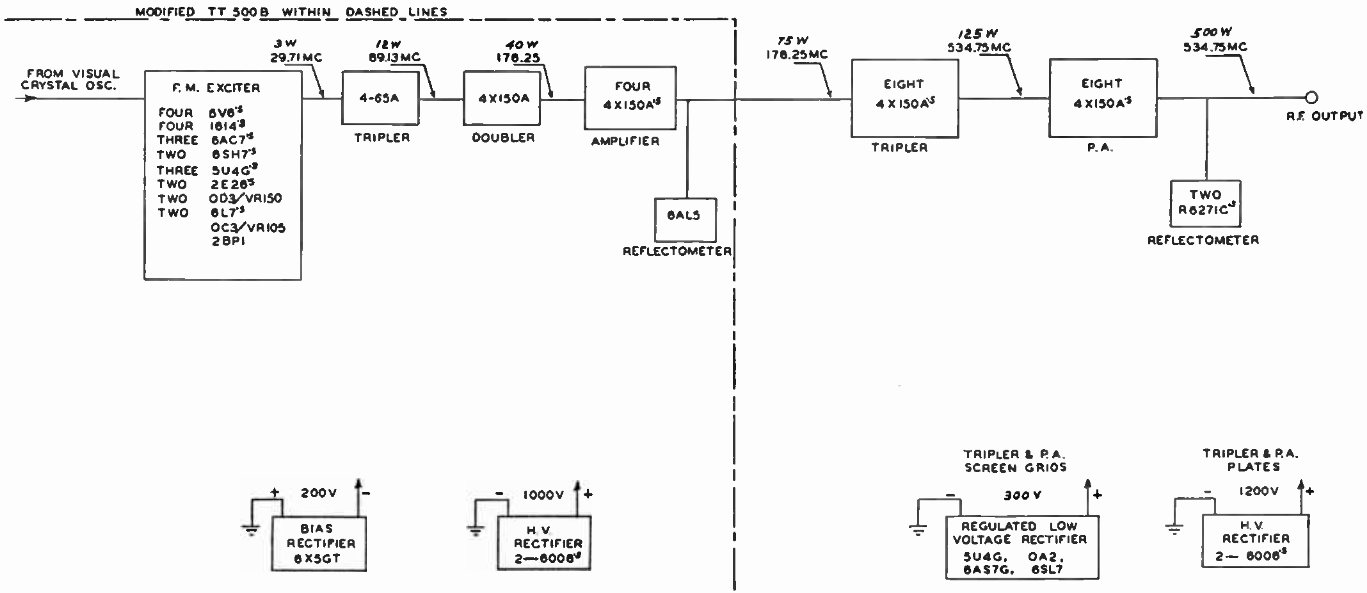


Fig. 4—Block diagram of aural transmitter.

in any tube, the associated circuit breaker trips instantly, providing the desired indication and at the same time removing the plate voltage from all tubes. Indicator lamps are also provided so that there is an overload indication even when the doors are closed.

A separate meter is provided to measure the cathode current of each tripler and power amplifier tube to facilitate the tuning of the cavities.

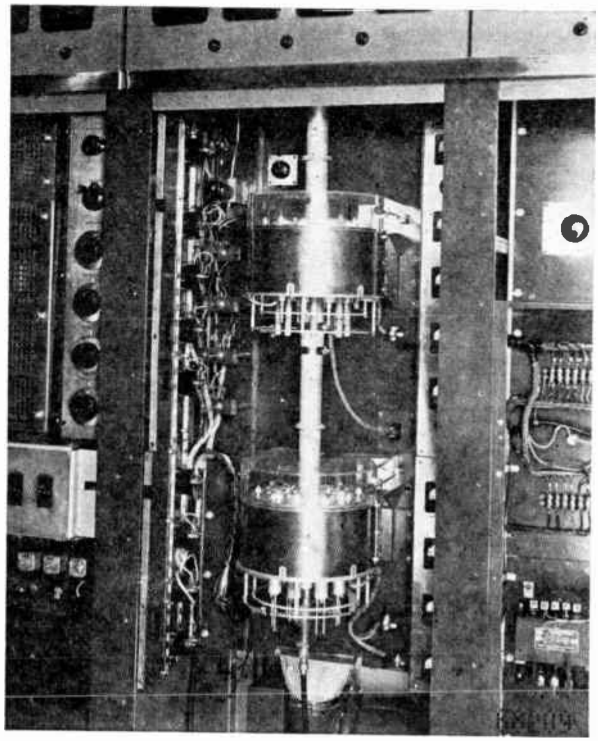


Fig. 5—Tripler and power amplifier stages.

The power supply circuits are conventional except that all circuits in the visual transmitter which are critical to voltage variations are supplied by regulated power supplies. These circuits include all video amplifier stages as well as the screen grids of the intermediate power amplifier, tripler, and power amplifier tubes.

Radio-Frequency Cavity Description

The high-frequency tripler and power amplifier are operated as grounded cathode-grounded screen amplifiers. Each uses eight 4X150A tubes mounted in a single cavity. Figure 8, a cross-sectional view of the cavities and the coupling between them, shows that the power

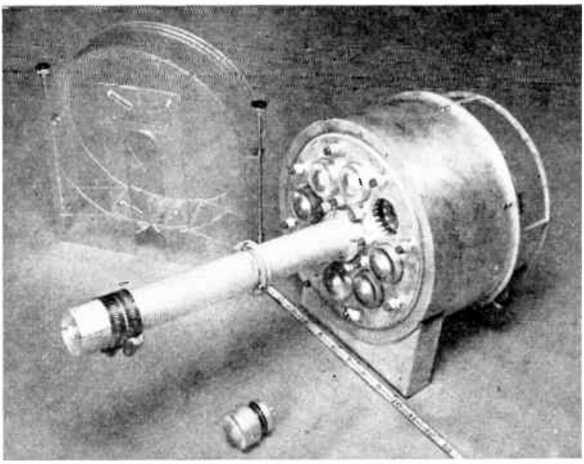


Fig. 6—Power amplifier cavity.

amplifier and tripler cavities are identical except for the grid circuits. Each cavity consists of three concentric cylinders shorted at one end and capped by three flat circular plates, referred to as the anode, screen, and grid plates. These plates provide a base on which the tubes and sockets are mounted.

A sheet of silvered mica placed over the "anode-plate" of the cavity forms the plate blocking capacitor. A thin phosphor-bronze sheet which is cut and formed to provide spring contact fingers for the anodes of all the tubes is laid over the mica and a metal plate provides rigidity to the assembly. The entire assembly is fastened together by means of insulated studs.

The screen grid of each tube is terminated in a ring at the base of the tube. These rings are connected through the screen contact fingers to a flat circular plate. A sheet of mica between this plate and the "screen plate" of the cavity forms the screen by-pass capacitor.

The center conductor of the output transformer is connected to the screen grid. That part of the conductor inside the cavity is an induct-

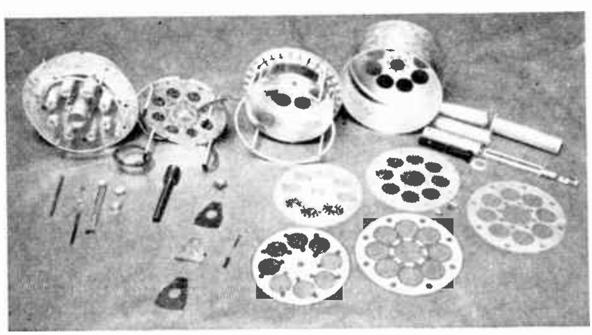
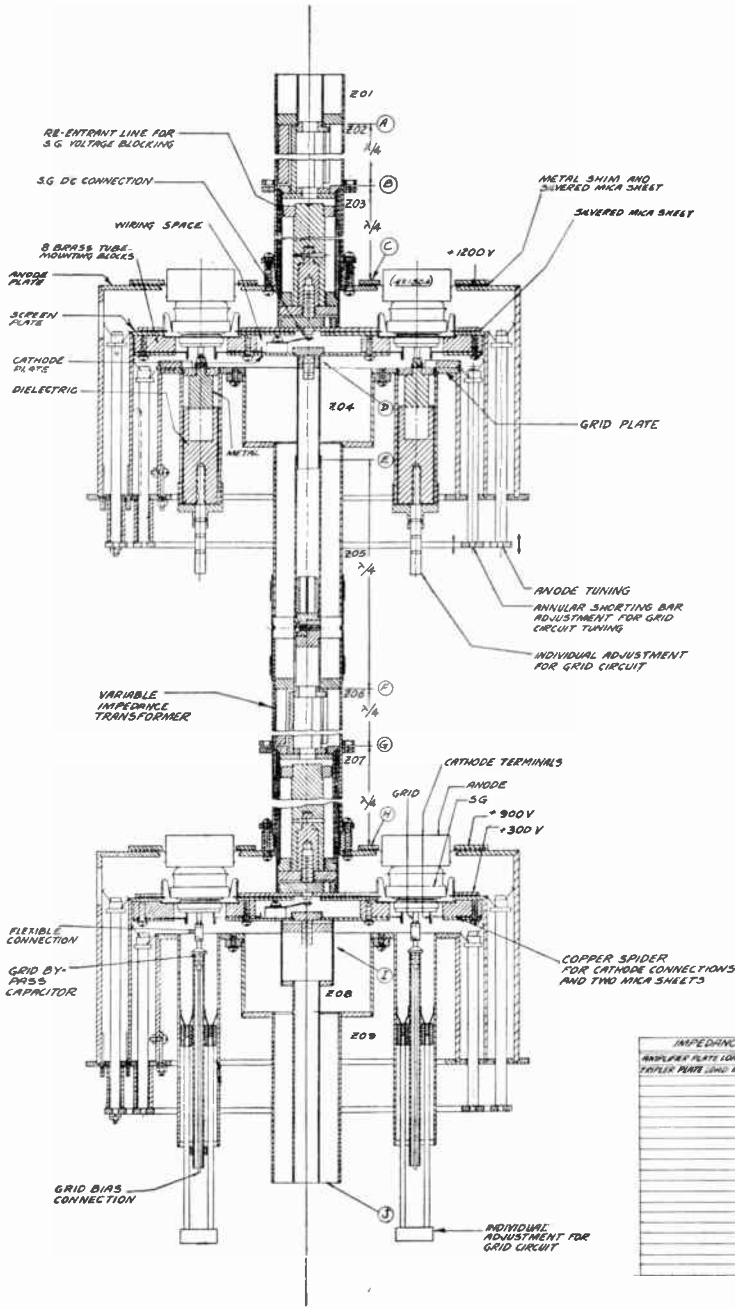


Fig. 7—Disassembled cavity.



IMPEDANCE TABLE	
AMPLIFIER PLATE LOAD IMPEDANCE	385 Ω
TRIPLER PLATE LOAD IMPEDANCE	655 Ω
Z01	515 Ω
Z02	[illegible] Ω
Z03	25 Ω
Z04	100 Ω
Z05	515 Ω
Z06	[illegible] Ω
Z07	25 Ω
Z08	515 Ω
Z09	515 Ω
ZA	515 Ω
ZB	36 Ω
ZC	047 Ω
ZD	100 Ω
ZE	100 Ω
ZF	265 Ω
ZG	231 Ω
ZH	028 Ω
ZI	415 Ω
ZJ	515 Ω

Fig. 8—Cavity cross section diagram.

ance which is a part of the resonant circuit of the cavity. If the cavity were to be cut into eight pie sections, each tube would be in the center of a half-wave resonant line foreshortened by the tube capacity. The equivalent electric circuit is shown in Figure 9. The circuit has been designed so that when it is resonated at 530 megacycles, approximately half the tube displacement current will flow out toward the shorting bar, and half will flow in toward the transformer. This tends to keep the current distribution uniform in the tube seals. The output impedance of the power amplifier is about ½ ohm at full power output, while the impedance in the tripler is about ¼ ohm.

The cathode of each tube is individually by-passed to a "cathode plate" which is connected electrically to the "screen plate." Thus the direct-current component of the cathode current of each tube may be metered separately. There is no radio-frequency field in the space between the "screen plate" and the "cathode plate"; therefore this space is used for the direct-current and 60-cycle leads to the filament, screen, and cathode. These leads are brought out from the cavity through four copper tubes which are mounted along the inside of the "screen cylinder."

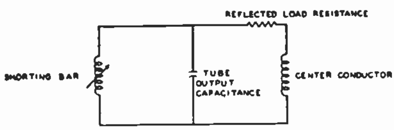


Fig. 9—Equivalent plate circuit of a single tube.

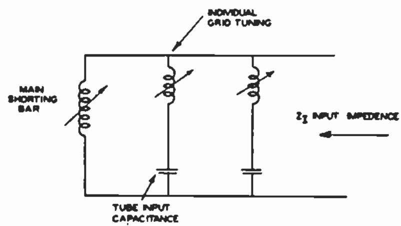


Fig. 10—Equivalent circuit of tripler grid cavity.

In the tripler, the grid of each tube is connected to a feed-through type by-pass capacitor which is screwed into the center conductor of each tuning stub. The outer conductor of each tuning stub is soldered to the "grid plate" of the cavity. The equivalent electrical circuit is as shown in Figure 10 where only two of the eight grids are shown. By adjusting the main shorting bar and the individual tuning adjustments, the grid cavity can be resonated and the drive on each tube can be balanced. By differentially tuning the main shorting bar and all eight of the individual tuning adjustments, the input impedance, Z_I may be varied for proper matching. The tripler has been designed to match a 51.5-ohm coaxial line.

The design of the power amplifier grid tank is different from that in the tripler because at 530 megacycles, the reactance of the cathode and grid leads is approximately equal to the input capacitive reactance.

In order to resonate the power amplifier grids with the main shorting bar, an equivalent capacitance is inserted in series with the grid. This is done by the individual open circuited tuning stubs the capacitance of which is made adjustable by means of a movable dielectric sleeve. The equivalent circuit for the grid tank is shown in Figure 11.

The input impedance to the grids can be adjusted by differentially adjusting the main shorting bar and individual tuning adjustments as described for the tripler, but the power amplifier grid circuit is designed to have an input impedance of 100 ohms.

The power amplifier grid bias connection is not shown in Figure 8. The requirements of such a circuit are that it prevent all radiation outside the cavity, and present a high impedance to ground for the video frequencies. Eight quarter-wave chokes do this quite effectively. Since the radio-frequency impedance is low at the grid terminal of the socket, the center conductor of each choke is connected at this point. The choke is mounted in an insulating block adjacent to the individual grid tuning stub and the outer conductor of the choke is connected to the modulator output.

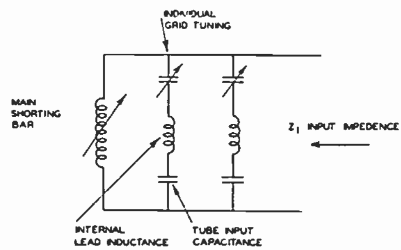


Fig. 11—Equivalent circuit of power amplifier grid cavity.

Both the anodes and grids of the 4X150A tubes require forced air cooling to prevent overheating of the seals. To accomplish this, air is forced through a screened slot in the side of the cavity in the anode tank. Inside the cavity, the air splits into two paths. Part of the air is forced out of the cavity through the anode radiators, and the rest is forced down across the grid seals of the tubes and out the grid cavity. To increase the air flow across the grid seals, three short bakelite pins are cemented into the tube socket. These raise the tube slightly above the socket and allow the air to pass across the base of the tube and through the center hole in the socket into the grid cavity. A plastic shield fastened over the top of the cavity creates a back pressure which further increases the flow of air across the grid seals and in addition provides a safety cover for the anodes of the tubes which would otherwise be exposed.

OUTPUT COUPLING TRANSFORMERS

The output coupling of the power amplifier cavity consists of a two-section impedance matching transformer. Each section is a quarter wavelength long, and together they transform the 51.5-ohm line impedance down to ½ ohm at the output of the power amplifier cavity. Since the inner conductor of the first section is at the same direct-current potential as the screen of the tubes, a re-entrant quarter wave section blocks the direct-current screen voltage but provides a low impedance radio-frequency path between the inner conductors of the first and second sections of the output coupling transformer.

Transforming from 51.5 to 0.5 ohms would ideally require two sections of surge impedance, 1.59 and 16.2 ohms respectively for maximum broadbanding, but if a 1⅝-inch line is used as the outer conductor a section of 1.59-ohm surge impedance is mechanically impractical especially since a potential of about 300 volts exists between the outer and inner conductor. As a compromise, a surge impedance of 2.5 ohms was chosen for the first section. This fixes the surge impedance of the next section at 25 ohms, but in order to vary the loading on the power amplifier, this section was made variable. A two-to-one change in surge impedance of this section results in a four-to-one change in power amplifier output impedance, which was considered adequate. The surge impedance of the second transformer section should therefore be adjustable between 18 ohms and 36 ohms.

A cross section of a tentative design was laid out to scale and a flux plot was made to determine the capacitance per unit length. A flux plot gives a fairly accurate determination of the capacity if it is done carefully. To make one, the electric and magnetic lines are sketched free-hand following a knowledge that:

1. The electric lines terminate perpendicular to the conductor.

2. Electric and magnetic lines are mutually perpendicular.

If the electric and magnetic lines are drawn to approximate squares, then:

$$C = \frac{\text{Number of ``squares'' along magnetic lines}}{\text{Number of ``squares'' along electric lines}} \times 8.854 \text{ micromicrofarads per meter.}$$

A cross-sectional view of the final design is shown in Figure 12. The outer conductor consists of three square rods soldered to the inside of the 1⅝-inch tube and the inner conductor consists of three half-round rods supported by a plate at each end as shown. To change the surge impedance, the outer conductor is rotated with respect to the inner conductor. A slotted flange at one end limits the travel from maximum to minimum surge impedance. From the flux plot, $Z_{0\,\text{max.}}$ and $Z_{0\,\text{min.}}$ were found to be 38.7 ohms and 18.1 ohms respectively. After the transformer was built, the surge impedance was measured on a slotted line. The realized values of $Z_{0\,\mathrm{max.}}$ and $Z_{0\,\mathrm{min.}}$ were 37.3 and 20.2 ohms respectively.

Although the output impedance of the tripler cavity is ¼ ohm, it is desirable to use the same type of output coupling. If this is done, the variable section of the output coupling must be terminated in 25 ohms. Since the input impedance to the power amplifier grid cavity is 100 ohms, a quarter-wave section of 51.5-ohm line will provide the proper match. The cavities are mounted far enough apart to do this, and a short section of 100 ohm line completes the connection.

MODULATOR

The modulator design is based on the TT-500B modulator which is designed to operate from a standard RMA composite video signal. The TT-500B modulator (two type 807 tubes in parallel) is capable of fully grid modulating four 4X150A tubes with a bandwidth of about 5 megacycles. Since the TTU-1A transmitter requires that the eight 4X150A tubes be similarly modulated, it is clear that sufficient voltage can be obtained for modulation, but at a reduced bandwidth due to the additional grid capacity of the extra tubes. In order to reduce the capacity that the 807 tubes must work into, they are followed by a cathode follower which in turn delivers its output to the grids of the power amplifier tubes. Several advantages are gained by the use of the cathode follower. The input capacitance of the cathode follower is considerably reduced because of degeneration provided by the cathode resistor and the linearity is very much improved. In addition, with a given capacitance load and bandwidth, it is possible to use a larger load resistor, which results in less modulator plate current.

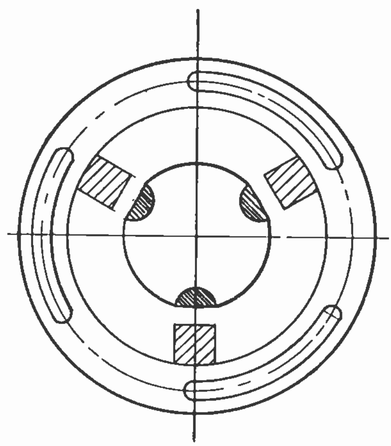


Fig. 12—Cross section of variable impedance transformer.

The choice of a modulator tube was the next problem. It was decided to use eight modulator tubes with the grids connected in parallel for video signals, rather than one larger tube for several reasons. First, by providing a separate bias control for each modulator tube and direct coupling each modulator tube to a single power amplifier tube it becomes quite simple to compensate for any unbalance in static plate currents. Secondly, an unreasonably large tube operating at a fairly high plate voltage would be needed to deliver the required 71 volts peak to peak to the 300-micromicrofarad load presented by the grids of the eight power amplifier tubes. The one objection to the use of a multiplicity of modulator tubes was the high input capacitance which made the design of the input circuit more difficult. The number of components in the input circuit becomes quite large when the direct-current grid potentials must be separately adjustable as noted above. Thus, it becomes essential that the components be as small as possible and the location of all grid circuit components becomes very critical.

A simplified schematic diagram of the modulator is shown in Figure 13. The video output tubes of the TT-500B transmitter are direct coupled to eight 6L6 tubes with grids connected in parallel for video frequencies and each tube is direct coupled to a single power amplifier tube.

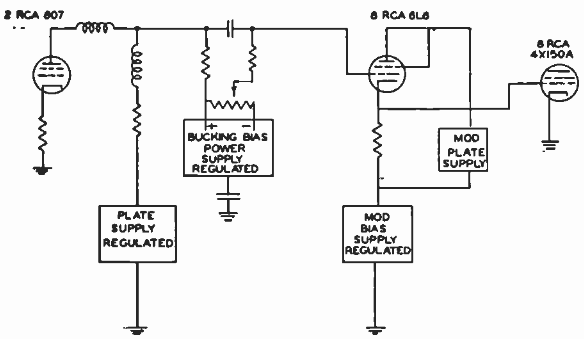


Fig. 13 — Simplified schematic diagram of modulator.

A "bucking bias" power supply was connected between the plates of the 807 tubes and the grids of the 6L6 tubes in order to overcome the positive voltage at the plates of the 807 tubes. Since the power supply has a considerable capacity to ground and is at video potential above ground it was necessary to isolate the supply by means of high resistance in both positive and negative leads. This was possible since the modulator tubes do not draw grid current. In order that the shunt impedance presented by these resistors and the power supply capacitance be held constant over the video frequency range, the power supply capacity was increased to make the RC constant of the shunt network equal to 0.025 second. Thus, for all frequencies above approximately 40 cycles there is a loss of about 2 per cent compared to the direct-current response.

A video amplifier is provided for monitoring the modulator output. The input to the monitor amplifier is obtained from the grids of the power amplifier tubes by means of eight voltage dividers. These are connected together at the low-potential end which is connected to the grid of the amplifier. Thus, the video input is the average of the signals applied to the eight power amplifier tubes. The resistors used for the voltage dividers are of the deposited carbon type which have good high frequency characteristics and a shunt capacitance of only 0.8 micromicrofarads. This value of capacitance is almost sufficient for high frequency compensation of the divider. However, a capacitance of 1.5 micromicrofarads across only one of the resistors results in nearly complete compensation up to about 5 megacycles. By proper choice of resistance values, the voltage divider also serves as a meter multiplier, so that the average grid voltage of the eight power amplifier tubes may be read on a single meter.

In modulating the power amplifier it was found that the response of the modulated wave had a deep valley at approximately 3.5 to 4 megacycles, although the video voltage applied to the grids of the power amplifier was flat to 5 megacycles. This was traced to resonance in the cathode circuits of the power amplifier between the cathode bypass capacitors and the cathode lead inductance. These leads are of necessity quite long because the cavities are plug-in. The inductance was reduced as much as possible by shortening the lead length and shielding the leads and in addition each cathode circuit was critically damped by means of a blocking capacitor and damping resistor connected to ground. This eliminated all traces of cathode resonance.

Sound Exciter

The frequency-modulation sound exciter is designed to accept the standard RMA audio input signal of 10 dbm* ± 2 dbm at 600 ohms and deliver to the multiplier stages of the transmitter a frequency modulated signal at 1/18 the output frequency of the transmitter and at a power level of about three watts. The exciter is similar to the standard frequency-modulation exciter[1] developed for the RCA FM broadcast transmitters except that the center frequency is stabilized so that the output frequency of the aural transmitter is maintained 4.5 megacycles higher than that of the visual transmitter within ± 450 cycles. This relative stability is considerably in excess of that which is required for the satisfactory operation of receivers of the inter-carrier-sound type. The tolerance proposed by several RMA technical committees is ± 5 kilocycles.

Figure 14 shows a block diagram of the frequency control system.

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* Decibels referred to a zero level of 1 milliwatt in 600 ohms.

[1] N. J. Oman, "A New Exciter Unit for Frequency Modulated Transmitters", *RCA Review*, Vol. 7, No. 1, pp. 118-130, March, 1946.

The visual and aural transmitters are driven by a crystal oscillator operating at 4.9097 megacycles and a master oscillator operating at 4.9514 megacycles respectively. The master oscillator is frequency modulated by a reactance tube modulator.

The frequency of the master oscillator is accurately maintained at 1/108 of the transmitter output frequency by means of a motor controlled capacitor connected in the oscillator tank circuit. The motor is controlled by a frequency detector which compares two frequencies derived respectively from a crystal oscillator and from the difference frequency between the visual and aural oscillators. Thus, any error in the master oscillator frequency causes a correcting force to be applied to the tuning capacitor.

One of the reference frequencies operating the frequency detector is obtained directly from an auxiliary crystal oscillator operating at a frequency of 104.165 kilocycles through a frequency divider with a division ratio of five to one to produce the reference frequency of approximately 20.8 kilocycles.

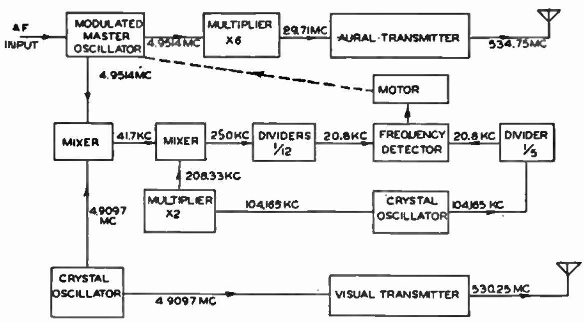

Fig. 14 — Block diagram of frequency control system.

The other reference frequency is obtained by combining the outputs of the visual and aural oscillators in a mixer to produce a frequency of approximately 41.7 megacycles which is the difference between the oscillator frequencies. This frequency is mixed again with the second harmonic of the auxiliary crystal oscillator to obtain a new frequency of 250 kilocycles, which after being divided by 12 becomes the second reference frequency of 20.8 kilocycles. Thus any deviation of the master oscillator frequency from 4.9514 megacycles causes a change in the second reference frequency which in turn causes the motor to rotate and correct the oscillator frequency. In this manner, the frequency stability of the visual transmitter is determined only by the visual crystal oscillator. The center frequency of the aural transmitter is stabilized with reference to the visual output frequency so that the frequency difference is maintained within the specified limits.

Tank Circuit

During the initial operating tests of the cavity, a power output of 475 watts was obtained with a power input of 2000 watts. The difference between output and input power represents loss due to plate dissipation in the tubes and also radio-frequency losses in the tank circuit. The tank circuit losses were found in the following manner.

The unloaded Q of the cavity was measured to be 94 and an output capacity reactance of 8 ohms was calculated for the eight tubes in parallel. Substituting these values in the formula

$$P_d = \left[\frac{E^2_{\text{max.}}}{2\,X_c}\right]\left[\frac{1}{Q_d}\right], \tag{1}$$

and assuming a radio-frequency output voltage of 900 volts peak, which is consistent with the plate voltage applied, the power dissipated in tank circuit losses is shown to be 540 watts.

It is interesting to note that the loaded Q of the cavity can also be calculated. Restating (1) and substituting $P = 475 + 540 = 1015$ watts gives

$$Q_0 = \frac{E^2_{\text{max.}}}{2X_c\,P} = 49.7, \tag{2}$$

where Q_0 is the loaded Q of the cavity.

The above calculations are not strictly accurate because they neglect the energy stored in the electric field of the cavity itself, but they do give a fair enough approximation to indicate an excessive amount of circuit loss and in addition point to a possible method of obtaining a quantitative breakdown of the various losses in the cavity.

Since the Q of any electrical circuit is the ratio of the stored energy to the energy dissipated in the circuit, the quantity $1/Q$ is proportional to the power loss in that circuit. Thus, to find the loss factor, $1/Q$, of any component, the Q of the cavity can be measured with the component in and out. As long as the stored energy in the circuit has not been changed, the loss factor for the particular component will be the difference between the two $1/Q$ factors. It was decided to break down the circuit losses by successive Q measurements to determine which parts of the cavity contributed the most loss.

Originally, it was supposed that the low current densities resulting from the size of the cavity would allow the top plates to be fitted to the cylinders by a press fit. These seams were soldered and a considerable improvement was measured.

Various other possible losses were measured in order to isolate each loss as much as possible. For example, the loss in the shorting bar was measured by opening half the finger contacts and measuring the increase in loss. This increase is equal to the loss in the shorting bar. In this manner, many of the losses in the cavity were measured.

A summary of these results is given in Table I. The values represent the average of several readings, and are probably accurate to within 10 or 15 per cent.

Table I

Loss	Loss Factor — ($1/Q$)	
Total		87.5×10^{-4}
Plate—cylinder seams	30.5×10^{-4}	
Plate mica by-pass	24.4×10^{-4}	
Screen mica by-pass	6.4×10^{-4}	
Tubes	11.7×10^{-4}	
Transformer section	3.3×10^{-4}	
Main plate shorting bar	6.5×10^{-4}	82.8×10^{-4}
Unaccounted for		4.7×10^{-4}

It should be noted that the unaccounted loss contains the accumulated error of all readings and is approximately 5 per cent of the total loss.

Both the plate and screen micas were silvered, and the direct-current screen terminal was changed from the tube sockets to the center point of the screen finger plate (see Figure 8). Table I shows that these improvements together with the improvement made by soldering the seams should have increased the Q to 380 if all losses in the plate and screen by-pass capacitors had been eliminated. When the unloaded Q was again measured, it was found to be 350. With these changes a power output of 1050 watts was measured.

There are two other items of importance in connection with the design of the radio-frequency cavities. The first is the spurious responses in the plate cavity. When the cavity is operated in the desired mode, the electric field is constant around the cavity. Since the cross section of the cavity represents an electrical half-wave circuit, it acts like a wave guide at cut-off, and there can be no mode propagated around the cavity at the fundamental frequency. A few megacycles above the resonant frequency however, such modes can be propagated and a great many spurious resonances can occur above the fundamental resonance because of the multiple reflections of the curved walls of the cavity.

The frequencies of these spurious resonances are far enough away from the fundamental frequency so that they are not excited by either the carrier or the radio-frequency modulation components.

One other item of interest is the method of neutralizing the tubes. Although the 4X150A proved very stable as far as oscillation was concerned, it was found necessary to neutralize the amplifier in order to reduce the "feed through" power and improve the modulation characteristic at the cut-off end. Since the screen by-pass capacitance was fixed, it was decided to neutralize by means of a Bridged-T circuit. The neutralizing voltage is provided by the inductance of the screen contact fingers. A single-tube cavity was set up and several different types of screen fingers were measured for minimum feed-through.

Eight-Tube Cavity Performance

The following discussion of the performance of the eight-tube cavity is divided into two parts. First, a comparison of tube performance at 530 megacycles and at low frequencies, and second a comparison of the performance of an eight-tube cavity versus the performance of a single-tube cavity.

There are four principal differences between the tube performance at 530 megacycles and the calculated performance which is based on low frequency operation. These are:

1. Increased driving power,
2. Lower apparent plate efficiency,
3. Increased radio-frequency grid voltage required for a given value of plate current,
4. Back heating of the cathode.

The first two effects were determined to be almost entirely due to circuit losses. If all the known circuit losses are taken into account by the method previously described, the measured radio-frequency power is 90 per cent of the calculated value. The remaining 10 per cent is due to transit time effects and to variation in tube characteristics.

The increased radio-frequency grid voltage and cathode back heating are probably both due to transit time effects at ultra-high frequencies. Measurements showed that the grid drive was about 50 per cent more than the calculated value for any specified plate current. The back heating was easily compensated for by reducing the filament voltage from 6.0 to 5.2 volts.

An accurate comparison between the operation of an eight-tube cavity and a single-tube cavity requires the operation of eight tubes, one at a time, in a single-tube cavity and comparing this operation with the operation of the same eight tubes in an eight-tube cavity. Although this was not done, the performance of a single-tube cavity had been observed on numerous occasions, and enough data was available to compare the two types of operation under equivalent conditions. For example, a 4X150A in a single-tube cavity developed 150 watts power output at 240 milliamperes plate current, and under the same conditions each 4X150A developed an average of 137 watts or 92 per cent of the single-tube output in the eight-tube cavity. The unloaded Q of the single-tube and eight-tube cavities were found to be 488 and 350 respectively. Loaded Q of a single-tube cavity was not measured, but assuming it to be the same as the value measured for the eight-tube cavity ($Q = 58$) it can be shown that the circuit efficiencies of the eight-tube and single-tube cavities are 83.5 and 88 per cent respectively. From this one would expect the output of the eight-tube cavity to be 95 per cent of the output of the single-tube cavity, a discrepancy of 3 per cent from the measured value.

The conclusion is that when cavity losses are taken into account the operation is essentially the same in either a single-tube or an eight-tube cavity.

PERFORMANCE

The TTU-1A transmitter is designed to meet the proposed standards for commercial ultra-high-frequency television transmission. The performance limits for VHF transmitters as stated in RMA standards and in the FCC "Standards of Good Engineering Practice Concerning Television Broadcast Stations," have been met wherever they apply. In addition the relative stability of the carrier separation has been greatly improved over that required by current or proposed standards.

The visual transmitter is conservatively rated at a peak output power of one kilowatt when transmitting a black picture. Under these conditions the output tubes are operating considerably below the maximum allowable plate dissipation. The regulation of the output is such that the peak output varies less than 5 per cent from an all black picture to an all white picture.

The output of the transmitter contains both sidebands; however, the lower sideband is removed by means of a vestigial sideband filter externally connected. The radiated signal has the standard RMA vestigial sideband characteristic.

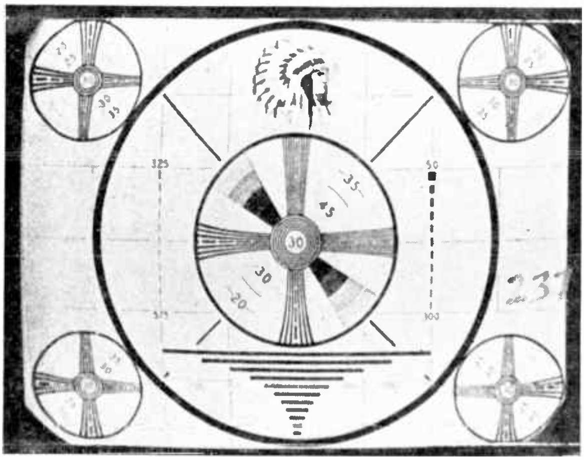

Fig. 15—Picture at input of transmitter.

The aural transmitter is rated at a power output of 500 watts. It is frequency modulated and is capable of more than ±50 kilocycles swing with less than 5 per cent distortion from 30 cycles to 15 kilocycles; ±25 kilocycles swing represents 100 per cent modulation. The distortion at 100 per cent modulation is less than the limits stated in RMA Standard TR104-A.

The carrier frequency of the visual transmitter is held constant to within 0.002 per cent of the assigned value. The frequency difference between the aural and visual carrier frequencies is constant to within ±450 cycles which compares very favorably with the stated requirement of ±5 kilocycles.

The bandwidth of the plate circuit of the power amplifier is sufficiently broad so that the picture quality is not affected by plate circuit

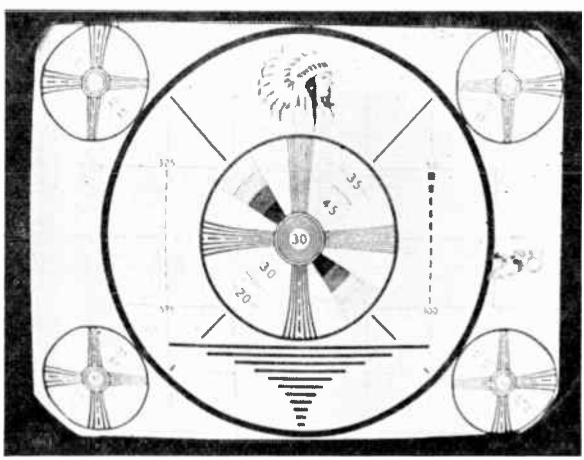

Fig. 16—Picture at output of transmitter.

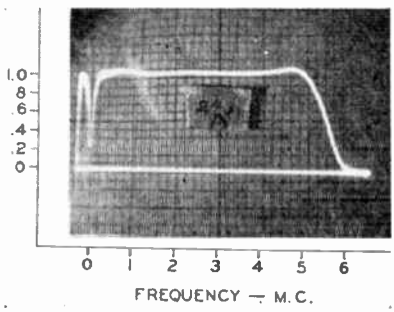

Fig. 17—Modulator video response.

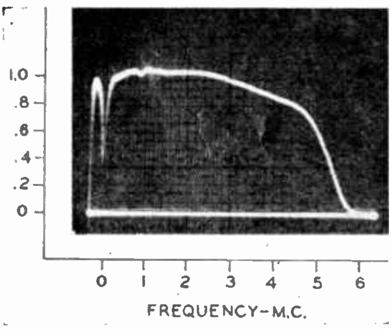

Fig. 18—Overall video response.

tuning. In tuning the transmitter, the plate circuit is tuned for maximum output. Optimum broadbanding of the grid circuit is readily obtained by slightly detuning on the high frequency side of resonance.

The performance of the visual transmitter is shown by Figures 15 to 20. Figures 15 and 16 show the input and output signal respectively when the standard test pattern is transmitted and viewed on a commercial master monitor.

Inspection of Figure 16 shows a vertical resolution of about 460 lines. Figure 17 shows the video response of the modulator measured at the grids of the power amplifier tubes. The response is flat to approximately 5 megacycles. Figure 18 shows the overall frequency response of the transmitter when operating into a dummy load and Figure 19 shows the response to a square wave pulse with a rise time of 0.07 microsecond. It indicates an increase in rise time to 0.11 microsecond. Figure 20 shows the overall modulation characteristic and indicates a feed-through power of only 0.3 watt. A maximum power of 1460 watts is indicated.

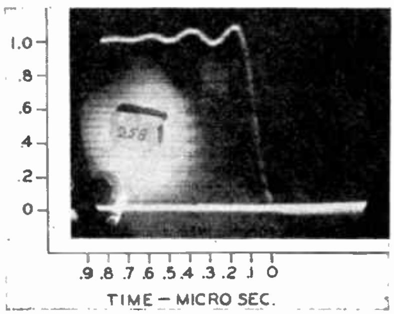

Fig. 19—Square wave response.

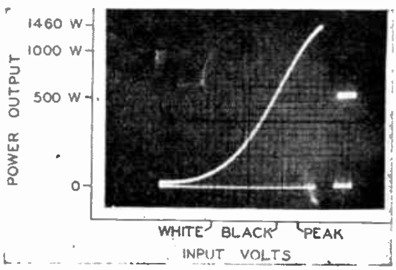

Fig. 20—Modulation characteristics.

Conclusion

The TTU-1A transmitter is capable of radiating a signal in accordance with the current FCC "Standards of Good Engineering Practice Concerning Television Broadcast Stations," and the RMA "Electrical Performance Standards for VHF Television Transmitters."

The quality of the transmitted picture is equal to that transmitted by commercial transmitters currently in operation on the VHF channels. In construction and design this transmitter is entirely suited for commercial operation from both the operational and performance standpoints.

A SIX-MEGACYCLE COMPATIBLE HIGH-DEFINITION COLOR TELEVISION SYSTEM*

A Report

BY

RCA LABORATORIES DIVISION, PRINCETON, N. J.

Editor's Note: This report comprises Exhibit No. 209 submitted by Radio Corporation of America at the Hearing before the Federal Communications Commission in Docket Nos. 8736, 8975, 9175 and 8976, September 26, 1949 et seq.

Previous reports on the new color system are contained in Exhibit Nos. 206 and 207 submitted to the Federal Communications Commission on August 25 and September 6, 1949.

Additional information on various aspects of the system is under preparation. It is planned to publish this information as it becomes available.

INTRODUCTION

THE color system described herein has its roots in the simultaneous method first disclosed on October 30, 1946 and subsequently described in detail at a Hearing before the Federal Communications Commission in Docket No. 7896 and in various published technical papers.[1,2]

The new system, as in the case of the wide-band simultaneous system, is completely compatible with the current black-and-white television system.

In addition, the new system includes later developments which, in essence, compress the simultaneous system into a 4-megacycle band suitable for a total channel assignment of 6 megacycles. Not only is the system so compressed, but no detail is lost in the process. This in turn insures a high-definition color picture, while at the same time preserving the normal definition of the black-and-white picture.

The compression of the simultaneous system is accomplished by a combination of two processes:

(a) use of the mixed-highs principle; and

(b) color-picture sampling and time-multiplex transmission.

These band-saving techniques are described in the following pages.

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* Decimal Classification: R583.

[1] RCA Laboratories Division, "Simultaneous All-Electronic Color Television," *RCA Review*, Vol. VII, No. 4, p. 459, December, 1946.

[2] R. D. Kell, G. C. Sziklai, R. C. Ballard, A. C. Schroeder, K. R. Wendt and G. L. Fredendall, "An Experimental Simultaneous Color Television System," *Proc. I.R.E.*, Vol. 35, No. 9, p. 861, September, 1947.

Studio and Related Equipment Characteristics

A block diagram of the color television broadcasting station is shown in Figure 1.

Studio Apparatus

The color camera (live, film, or slide), its related equipment, and the synchronizing generator are the same components used in the wide-band simultaneous system. These were described in Dockets No. 7896 and No. 8976 and in References (1) and (2).

This studio apparatus provides three signals, one for each of the primary colors (green, red and blue). Each of these signals may contain frequency components out to a maximum of four megacycles, and in addition an average or dc component.

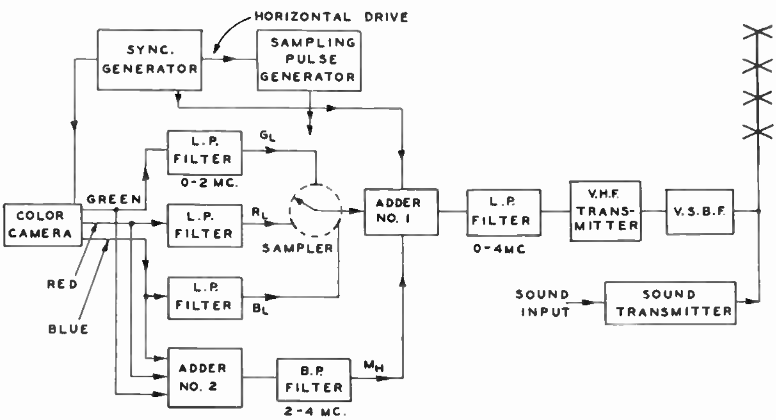

Fig. 1—Block diagram of the color television transmitter.

Signal Routing

For one signal routing of Figure 1, each color signal passes through a low-pass filter which eliminates frequency components above two megacycles. The green-channel signal coming out of its particular low-pass filter is designated as G_L on Figure 1, indicating that at this point the signal contains the dc component and ac components with frequencies of two megacycles or less. The three low-frequency signals, G_L, R_L and B_L are then sent into an electronic commutator or sampler (discussed below).

For the second signal routing of Figure 1, the three-color signals from the camera are combined in electronic Adder No. 2 and then are passed through a band-pass filter. The output of this filter contains frequencies from two to four megacycles, with contributions from each of the three color channels. The signal at the output of the band-pass filter is designated as M_H, the mixed-high signal. The mixed-high frequencies are fed to Adder No. 1 which, as will be seen, is also receiving the signal from the sampler and from the synchronizing generator.

Mixed-Highs

The principle of mixed-highs, referred to above, was described in Docket Nos. 7896 and 8976. It has been demonstrated that the mixed-highs procedure is successful and satisfactory in a wide-band simultaneous system.

In the new system, at the transmitting end, the sampling process (discussed below) is capable by itself, of providing high-frequency components of each color signal. Since the sampling frequency determines the highest frequency which will be passed, when the high-frequency components of each color signal are combined the resulting band-width does not exceed four megacycles.

However, photocells used with flying-spot devices in film and slide scanners inherently have poor high-frequency response, particularly in the red. It is often necessary, therefore, to peak the higher-frequency response of such devices. This peaking, of course, raises the level of the high-frequency noise components. When the entire frequency band of zero to four megacycles is sampled, the 3.8-megacycle sampling frequency beats the high-frequency noise to low frequency and vice versa. Because, in the case of pickup devices of the type referred to above, the high-frequency noise has been peaked, an even interchange of noise components does not occur, as it would if the noise spectrum were uniform. Consequently, after the sampling process has taken place, the low-frequency noise components will have been accentuated. This causes a coarse-grain structure in the picture which may be objectional to the eye.

Experience has shown, therefore, that there is a definite advantage gained in sampling only the lower half of the video band (up to two megacycles) and using the principle of mixed-highs for the upper half of the video band (from two to four megacycles), and this procedure is used at the transmitting end.

In pick-up equipment with uniform noise characteristics, no such effects as above described exist. This means that either the dual procedure given in the above paragraph or sampling only may be used. This also applies to receiving equipment.

Sampling and Combining Process

The sampling pulse generator, which embodies time-multiplexing techniques, is an integral part of the electronic commutator and makes use of the trailing edge of the horizontal synchronizing pulse to time the sampling of each of the color signals.

In the sampler, each color signal is sampled for a very short time, at a rate of 3.8 million times per second for each color. The samples for each scanning line are timed with respect to the horizontal synchronizing pulse for that line. However, for alternate line scans the timing of the sampling pulse is such that samples of any one color

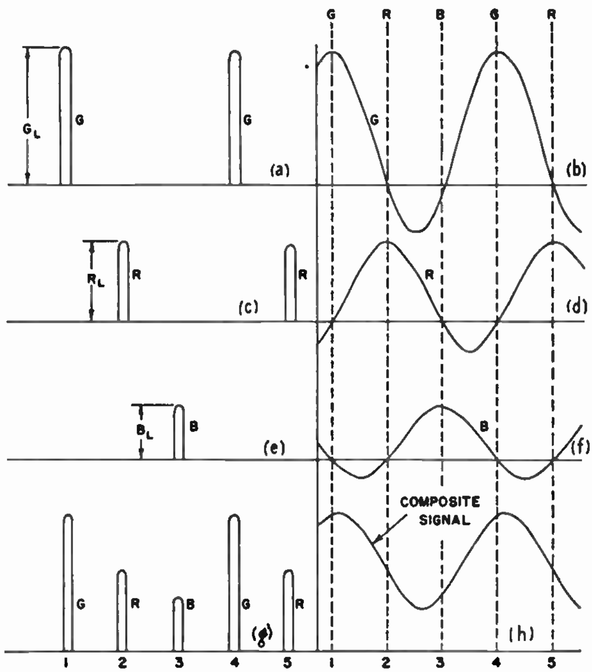

Fig. 2—Functioning of the sampling system at the transmitting end.

are taken at a point midway between samples of the same color in the line above. Alternate scans of the same line are displaced in this same way. Thus, the second samples are taken midway between the first samples, resulting in dot interlacing (to be discussed in more detail in Part IV).

Figure 2 illustrates the functioning, at the transmitting end, of the sampling system in the pickup of large uniform polychromatic areas, with the three primary colors represented by three different signal strengths. Figure 2(a) shows the output of the sampler due to the green signal only. The green channel signal is sampled every

0.263 microsecond ($0.263 = 1/3.8$). At a time 0.0877 microsecond after a green sample, a sample is taken of the red signal. This time delay is one-third of the time between successive green samples. The red samples continue to be taken 0.263 microsecond apart as shown in Figure 2(c). The blue samples are taken at the same rate and follow the red samples by a time of 0.0877 microsecond, as indicated in Figure 2(e). The composite output of the sampler consists of a superposition of the green, red, and blue trains of pulses or samples. Figure 2(g) shows the signal in the circuit at the output of the sampler.

From the sampler the signals pass to an electronic combining device called Adder No. 1 in Figure 1. Standard synchronizing signals from the synchronizing generator are also applied at this point.

This signal and the synchronizing pulses from Adder No. 1 feed into the low-pass filter. In the case of large area color only, the mixed-highs signal is not present.

The narrow green pulses of Figure 2(a), occurring at a rate of 3.8 million pulses per second, are smoothed by the low-pass filter to give the result shown in Figure 2(b). This wave consists of a dc component, which is the average of the pulse sample, plus a sine wave which has a frequency of 3.8 megacycles (the filter having removed the higher order harmonics). The 3.8-megacycle sine wave and the dc component change together, as the green signal changes in strength, in such a way that the signal of Figure 2(b) always passes through zero at the same interval of time after the peak regardless of the strength of the green signal. The smoothed sample of the green signal may be expressed as: $\frac{1}{3} G(t) \; [1 + 2\cos{(2\pi f t)}]$ where $G(t)$ is the green signal as a function of time, and f is the sampling frequency, namely 3.8 megacycles. A study of this expression reveals that the smoothed green sample goes through zero 120 and 240 electrical degrees after the signal has reached its maximum value.

The above equation is an excellent approximation of the conditions existing in the circuit when the duty cycle of the samples for a given color is 15 per cent or less. Accordingly, the duty cycle of the system is maintained within these limits.

The red samples of Figure 2(c) are smoothed by the filter to yield the result shown in Figure 2(d). This again is made up of a dc component and a sine wave with a frequency of 3.8 megacycles.

Smoothing of the blue sampling pulses results in the contribution shown in Figure 2(f). It should be noted in Figures 2(b), 2(d), and 2(f) that when any one color signal reaches its maximum value, the other two responses are crossing the zero axis.

While the curves of Figures 2(b), 2(d), and 2(f) have been shown separately for illustrative purposes, it should be remembered that the pulse train of Figure 2(g) goes into the low-pass filter. Thus the composite signal of Figure 2(h) comes out of this filter. In this figure, the dc component is the sum of the dc components of the green, red, and blue signals, while the 3.8 megacycle sine wave is the sum of three sine waves of the same frequency. This results in a composite 3.8-megacycle sine wave with a new amplitude and phase position superimposed on the composite dc component.

The action of the system in the pick-up of varying color areas is illustrated by means of Figure 3. In Figure 3(a), the three color signals are shown as they enter the sampler, with the appropriate sampling pulses as they come out of the sampler indicated by vertical lines. These same pulses are shown in Figure 3(b), with the envelope indicating the result of smoothing in the filter. It will be appreciated that Figure 3 is not quantitative and is used purely for illustrative purposes. Fine detail of the color picture, carried by the higher frequency components, is, of course, now supplied by the mixed-highs signal, but this cannot be easily shown in the diagram.

TRANSMITTER

The output (time-multiplexed color signal, mixed-high color signal, and synchronizing signals) of the low-pass filter (zero to four megacycles) in the right center of Figure 1 is applied to the modulator of a conventional VHF or UHF television transmitter.

From this point on, including the transmitter itself, together with the vestigial sideband filter, diplexer, sound channel and antenna, the system is that of a normal black-and-white television station, with no changes necessary.

The signal transmitted is consistent with the "Standards of Good Engineering Practice Concerning Television Broadcast Stations."

RECEIVING EQUIPMENT CHARACTERISTICS

Figure 4 is a block diagram of one type of color television receiver. The radio-frequency circuits, the picture intermediate-frequency amplifiers, the second detector, the sync separator, the sound intermediate-frequency amplifiers, the discriminator, and the audio circuits are identical with those of a conventional black-and-white receiver.

The composite video and synchronizing signals from the second detector enter the sync separator, which removes the video signal and sends the synchronizing pulses to the deflection circuits and to the sampling pulse generator.

Sampling and Smoothing Process

The sampling pulse generator utilizes the trailing edge of the horizontal synchronizing pulse to actuate the receiver sampler in identical fashion and in synchronism with the transmitter sampler.

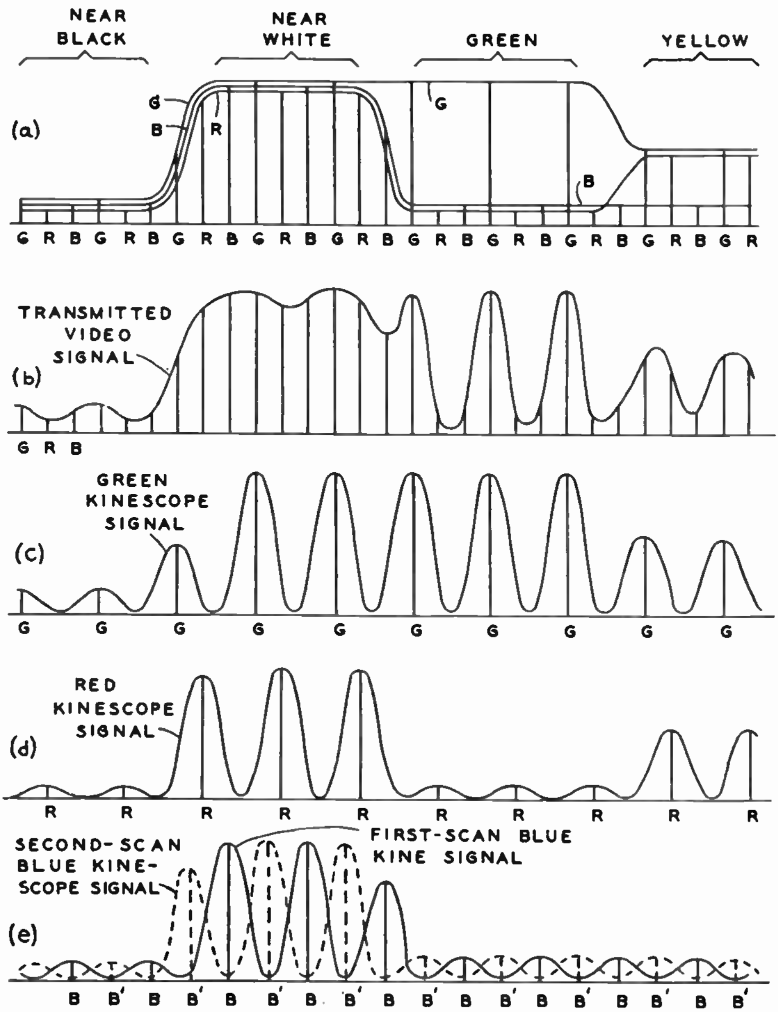

Fig. 3—Action of the system in the presence of varying color areas.

The signal from the second detector also enters the sampler. It has the same form as the composite signal of Figure 2(h), or as the solid

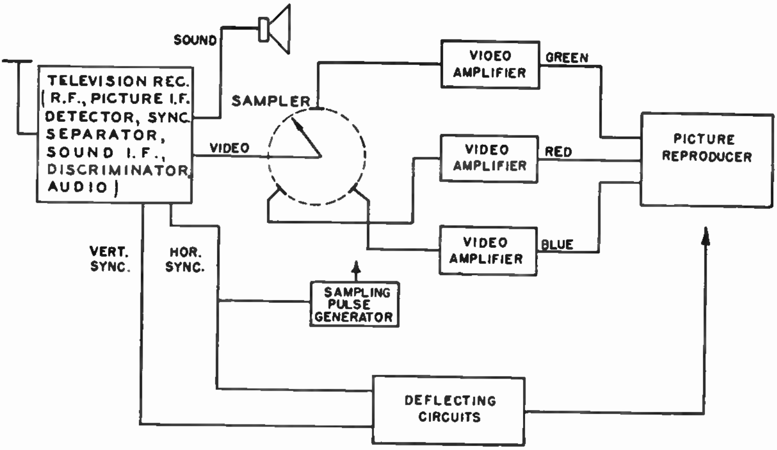

Fig. 4—Block diagram of one type of color television receiving equipment.

envelope of Figure 3(b). For ease of reference, Figure 2(h) has been reproduced as Figure 5(a). Again, the case of large uniform polychromatic areas is used for illustrative purposes.

The electronic commutator samples the composite signal every 0.0877 microsecond, producing the short pulses shown in Figure 5(a). The amplitude of each of these pulses is determined by the amplitude of the composite wave at that particular instant.

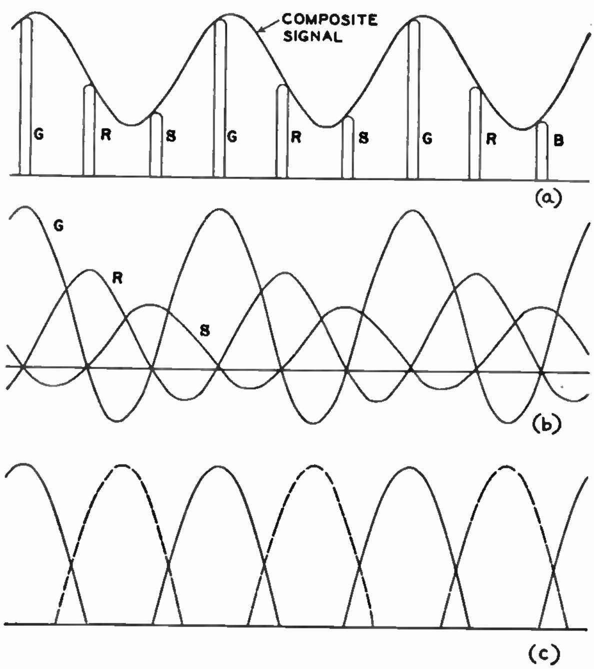

Fig. 5—Functioning of the sampling system at the receiving end.

The commutator feeds these pulses into three separate video amplifiers which in turn control the picture-reproducing apparatus which may consist of three cathode-ray tubes or kinescopes having appropriate color-producing phosphors. This method for portraying the single color picture with three kinescopes is similar to that demonstrated to the Commission during the Hearing on Docket No. 7896 and in References (1) and (2).

The video amplifiers have a flat frequency response to four megacycles, and must cut off completely at 7.6 megacycles. (Reference here is to the frequency response of the video amplifiers only and not to channel requirements.)

The sampler sends the pulses to each of the video amplifiers and its attendant kinescope in succession. For instance, in Figure 5(a), the first pulse shown in green goes to the green kinescope, the next pulse goes to the red, while the third pulse is sent to the blue. The green receives the fourth, seventh, tenth, and so on. Thus, while the individual pulses coming out of the sampler are 0.0877 microsecond apart, the green pulses going to the video amplifier for the green picture repeat every 0.263 microsecond. The green channel pulses of Figure 5(a), in passing through the video amplifier, lose all frequency components except the fundamental frequency of 3.8 megacycles and the dc component. The resultant smoothed signals are shown in Figure 5(b). The green, red, and blue signals are shown in superposition on this figure for illustration. It should be remembered that at this point the green signal shown is that fed to the green kinescope, while the red and blue signals are applied to their individual kinescopes.

Examination of Figures 2(b), 2(d), 2(f), and 2(h), has already revealed that, when the green signal is maximum, the red and blue signals are passing through zero. Hence, since the composite signal is sampled for green by a narrow pulse at the receiver at this exact instant, the receiver sampling pulse is a true measure of the green signal and includes no dilution from the red or blue signals. Likewise, the red and blue samples are each taken at points on the composite signal where no crosstalk is contributed from the other two color signals.

The above statement concerning absence of crosstalk holds good for all frequency components up to one-half the sampling frequency. For frequency components approaching the sampling frequency in order of magnitude, from a purely circuit aspect, crosstalk is present. However, the physiological characteristics of the eye which make possible the application of the mixed-highs principle apply equally well to the crosstalk of the higher-frequency components. Consequently, crosstalk in the fine detail is of no consequence.

Assuming that the kinescope actually cuts off with negative applied signal, and neglecting the nonlinearity of the input control-voltage versus light-output characteristic of the kinescope, the solid lines of Figure 5(c) may be regarded as the effective light intensity along one line scan in green. Figures 3(c), 3(d), and 3(e) show the effective signals for the green, red, and blue kinescopes, again for a single line scan.

Picture Dot Interlacing and Scanning Sequence

Returning now to Figure 5(c), it may be seen that a single line scan on the green channel lays down a series of green dots on the screen as shown by the solid lines. As was indicated above, these dots occur at a rate of 3.8 million times per second. If fine detail were involved to such an extent that two adjacent pulses in the green channel in a single line scan were of different amplitude, it is basic that the highest frequency component of use in establishing picture detail would be a sine wave which went from a crest to a trough in the time between the two adjacent green pulses. This sine wave would then have a frequency of 1.9 megacycles.

The fact that each pulse has a rise equivalent to twice this frequency allows the use of picture-dot interlacing to secure full detail up to a frequency band 3.8 megacycles wide. This is accomplished by shifting the sampling pulses the next time that the same line is scanned so that the dots are then laid down between the dots that were laid down in the first scan. This second series of green dots is shown by the broken curves in Figure 5(c). In this figure, the dots shown by broken curves are the same amplitude as the dots shown by the solid curves. For resolution of very fine detail. the dots laid down in the first scan would differ in amplitude from the dots laid down in the second scan of this same line.

Inspection of Figure 5(d) reveals that while a single line scan lays down a series of green dots on the screen with space between dots, this space is filled at the same time by red and blue dots, with great overlapping of the dots. The effect of the successive scans of a single line, Figure 5(e), shows even more clearly the complete covering of the line area with picture dots of three colors.

The scanning and interlace pattern used in the new color television system is illustrated in Figure 6. Each letter represents the center of a color dot area on the screen. The actual areas, of course, overlap to a great extent as discussed above.

During the first scanning field, illustrated in the upper diagram in Figure 6, the odd numbered lines are scanned in order. Colored dots are laid down in order along line 1 as shown. Next, line 3 is scanned with a displacement for each color dot shown, in the same fashion as described for the sampling at the transmitting end. The remaining odd lines are scanned in order. This scanning of the first field takes place in one-sixtieth of a second.

During the second field, the even lines are scanned, first line 2 with the colors laid down as shown, then line 4, and so on. The dot pattern laid down during the third field is shown by the lower diagram, where the odd lines are scanned in succession. During the fourth field, the even lines are again scanned in succession with the color dot pattern shown.

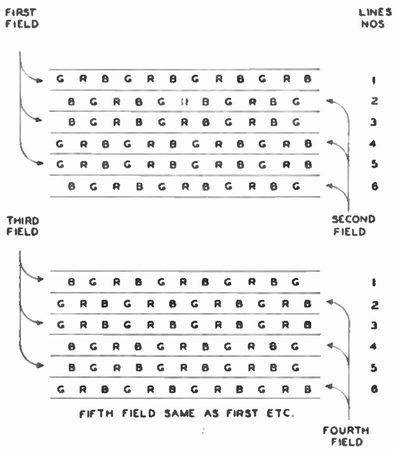

Fig. 6—Scanning and interlace pattern.

Thus, the odd lines are scanned during the first field, but dots of the same primary color are separated by spaces. The even lines are scanned during the second field, again with spaces between like color dots. During the third field, the odd lines are again scanned but the color dots displaced so that the spaces are filled. The even lines are scanned during the fourth field, with the color dots displaced to fill in the spaces left during the second field scanning. Four scanning fields are required to completely cover the picture area, with all spaces filled, with say, green dots. Simultaneously, the area is being covered with red dots and with blue dots. Since there are 60 fields per second, it may be said that there are 15 complete color pictures per second.

It should be remembered that the effective rate for large-area flicker is 60 fields per second, the same as for current black-and-white receivers. At viewing distances such that the picture line structure is not resolved, the effect of small-area flicker due to line interlace and picture-dot interlace is not visible.

Receiving Systems

In the receiver shown in Figure 4, the total signal consisting of the sampled signal plus the mixed highs has been inserted in the receiver sampler and picture-dot interlacing has been used to achieve high definition as discussed in detail above.

Another receiver arrangement is possible. In such a receiver, shown in Figure 7, the entire signal is fed into the sampler as before, but, in this case, low-pass filters with cut-off frequencies of approximately two megacycles are inserted between the sampler and the kinescopes. The low-frequency filters smooth out the pulses of Figure 5(c), so that the adjacent dots of a single color in one line scan now almost completely overlap. Because the pulses have been broadened by the two-megacycle filters in this receiver, horizontal resolution will not be increased by picture-dot interlacing at the receiver. Full resolution, however, is restored by obtaining mixed highs from the signal ahead of the receiver sampler and by-passing the mixed highs through a band-pass filter to the green, red, and blue kinescopes.

The color television receiver of Figure 4 and the alternate receiver of Figure 7 are examples of the flexibility afforded by this color system.

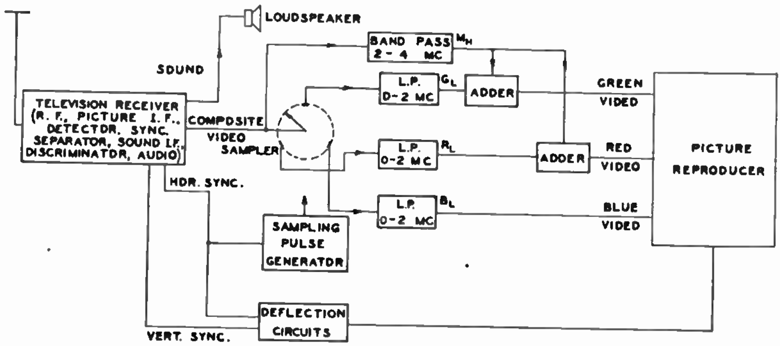

Fig. 7—Block diagram of color television receiver using by-passed highs

Reception in Black-and-White

When the color television signal is received on a current black-and-white receiver, the output of the second detector is represented by Figure 2(h), or, when the picture is of varying color, by the envelope of Figure 3(b). With mixed highs also transmitted as shown in Figure 1, the black-and-white receiver then develops on its kinescope a black-and-white picture with full resolution. The 3.8-megacycle sine wave superimposed on the picture signal produces a dot pattern on the kinescope in high chroma areas, but the dots are not visible at normal viewing distance. Examination of Figure 2 shows that in white areas, where the dot pattern would be objectionable if present, the three color signals are of the same amplitude and the composite signal consists of the dc components only. Hence, there is no dot pattern.

For color transmissions received in monochrome on a current black-and-white receiver, no band saving is involved, but because the transmitted signal contains all the resolution which a black-and-white signal of the same scene would have, the resulting monochrome picture will have the full resolution of the current standards.

Using the standard wedge pattern to test horizontal resolution, the same resolution figure has been obtained when reproducing the color transmission on an unchanged current model black-and-white receiver as may be obtained with the same receiver on a well-designed, well-adjusted black-and-white system using present broadcast standards. The vertical resolution is also consistent with current black-and-white standards.

When a color receiver is tuned to a television broadcasting station transmitting a black-and-white signal, the picture will appear in black-and white with full resolution on the color receiver picture reproducer. The successive pulses delivered to the three kinescopes will all be of equal magnitude, and, hence, will produce varying intensities of white —or a normal black-and-white picture.

Receivers and Color Converters*

Picture Reproducing Systems

One method for portraying a single color picture makes use of three kinescopes, reflective optics, and dichroic mirrors in a projection system. This has been previously demonstrated and described during the Hearing in Docket No. 7896 and in References (1) and (2).

Another method also makes use of three kinescopes in a projection system but uses refractive optics and dichroics instead of reflective optics. This system appears to lend itself more readily to compact design.

A third method uses three kinescopes with a pair of dichroic mirrors so arranged to permit essentially direct viewing. This system appears to lend itself more readily to a lower cost design.

A fourth method uses two kinescopes with a single dichroic mirror. This system appears particularly attractive for use in inexpensive receivers and color converters.

Because color receivers will probably be simplified by a color picture

* The various receivers and color converters illustrated and described in this section are research models designed to test and demonstrate the basic principles of the system. While indicative of possible approaches to the design of suitable receiving equipment, they are not intended to represent receivers and color converters of commercial design.

reproducer of the single-tube type, intensive research efforts on the problem are being continued.

Direct-View Receiver

Figure 8 represents a direct-view picture-reproducing system utilizing three kinescopes which are standard in every respect except that the phosphors are green, red and blue, respectively. The green, red and blue signals are impressed on the grids of their respective tubes. The deflecting yokes of the three tubes are connected in parallel, so that the rasters produced on the three screens are identical.

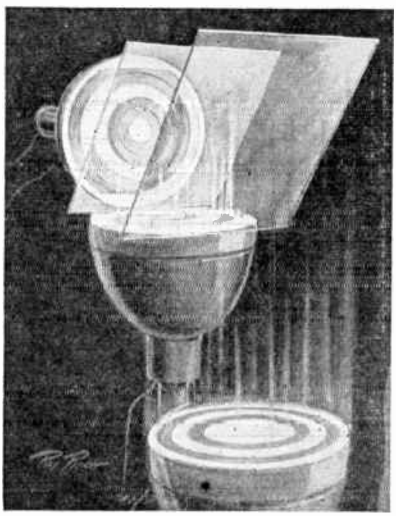

Fig. 8—Direct-view picture-reproducing system using three kinescopes and a pair of dichroic mirrors.

Fig. 9 — Positioning of ten-inch kinescopes for direct-view picture-reproducing system.

The tubes are viewed through dichroic mirrors. The red dichroic mirror reflects the red image from the red tube, the blue dichroic mirror reflects the blue image from the blue tube, and both mirrors are transparent to green light, so that the green tube is viewed directly through both dichroic mirrors. The red dichroic mirror is also transparent to blue light, so it does not interfere with the blue image. The mirrors and tubes are properly arranged so that to the eye the three pictures appear superimposed and are viewed as one picture.

Figure 9 shows three standard-size ten-inch kinescopes and the two dichroic mirrors mounted in a framework for proper viewing from the top, through a fully silvered mirror. A receiver using the above arrangement could be housed in a cabinet of the type shown in Figure 10. It is possible in this arrangement that, if the kinescopes are short-ended, the cabinet size can be materially reduced.

Projection Receiver

Another type of picture reproducing system is shown in Figure 11. This gives a projection picture, 15 × 20 inches. Three projection kinescopes are used which are standard except for the phosphors. Each tube control-grid is connected to the appropriate video channel, and the deflection yokes are supplied from a common source. Each tube is arranged in a reflective optical system. The light rays from each tube first strike a plane mirror, from which they are reflected to a spherical mirror of the proper focal length to produce an image on the screen. Beyond the spherical mirror the rays pass through a correcting lens and thence via a plane reflecting mirror to the projection screen.

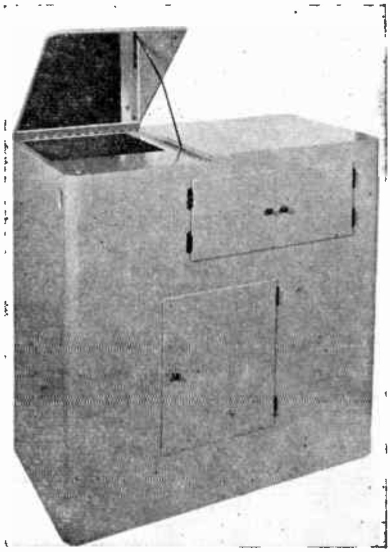

Fig. 10—One type of cabinet for direct-view receiver utilizing three kinescopes and a pair of dichroic mirrors.

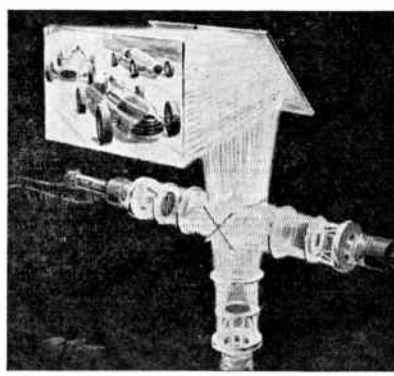

Fig. 11—Projection picture-reproducing system using three projection kinescopes, reflective optics and a pair of dichroic mirrors.

The green image passes through the red-and-blue-reflecting dichroic mirrors. The red image is reflected to the screen by the red-reflecting dichroic mirror. Similarly, the blue image is reflected to the screen by the blue-reflecting dichroic mirror. The complete optical system is so arranged that the three images are superimposed, in register and focus, on the projection screen, where they are viewed as a single color picture.

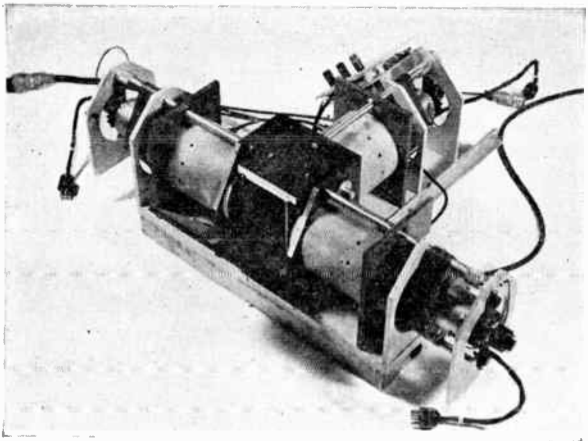

Fig. 12—Arrangement of the projection tubes and optical system used in the projection receiver.

Figure 12 is a photograph of the reflective optical system showing the mechanical arrangement of the projection tubes and the optical system. The receiver employing this system is shown in Figures 13 and 14.

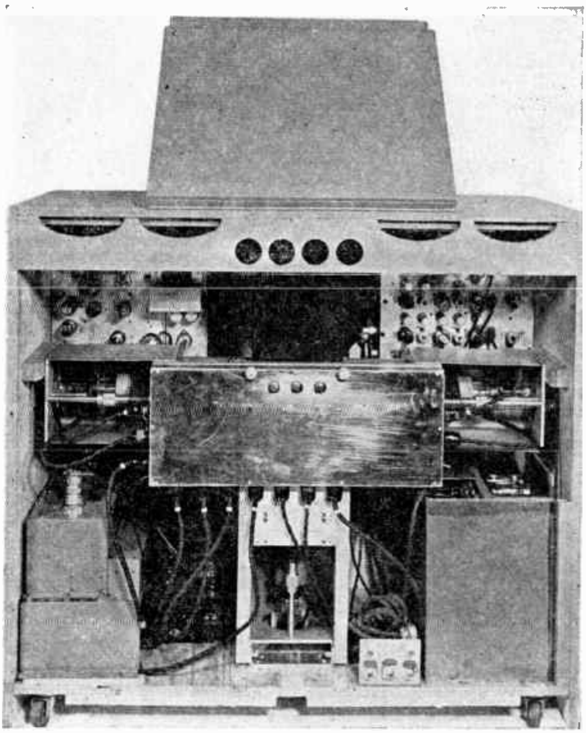

Fig. 13—Rear view of the projection receiver with 15 by 20 inch picture.

Projection Receiver with Magnifying Lens

The same type of projection picture reproducing system, referred to above, is used in the television receiver shown in Figure 15. Here the projected picture is smaller, and is viewed through a magnifying lens.

Color Converters

To convert a current black-and-white receiver to receive color transmissions in color requires the addition of color sampling circuits and a picture reproducer.

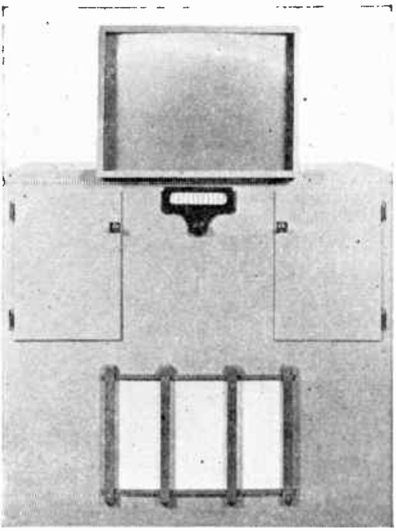

Fig. 14—Projection receiver with 15 by 20 inch picture.

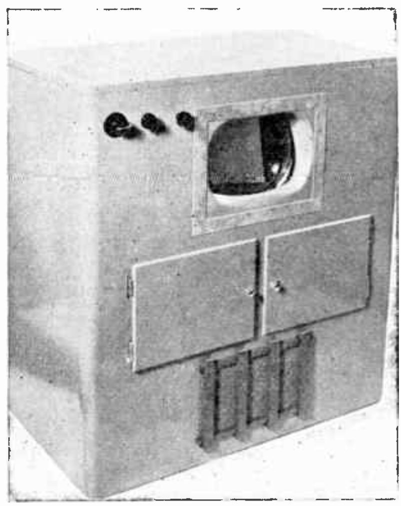

Fig. 15—Projection receiver using reflective optics and magnifying lens.

The size of the color converter for a direct-view picture or for a large projected picture is determined by the size of the kinescopes and the optical system. When the cabinet size and shape have been determined by these elements, the circuit components which need to be added to those already available in the black-and-white receiver can be fitted around the kinescopes and optical system in the cabinet without increasing its size.

A direct-view converter using three 10-inch kinescopes is shown in Figure 16. Interconnections between the standard receiver and the converter are made by a simple harness cable plugged into the tube sockets of the standard reciever. (Reduction in cabinet size through the use of shorter kinescopes, previously mentioned in connection with the direct-view color receiver, applies as well to this color converter.)

A smaller color converter can also be made which gives a projected picture. Three $1\frac{1}{2}$-inch projection tubes are mounted as shown in Figure 17. The complete kinescope and optical system assembly is mounted on the back of a standard television receiver, which can be either a table model or a console. The kinescope (any size) is removed from the black-and-white receiver and the color picture is projected, through the space it occupied, to a screen mounted in the normal picture-mask opening. An additional chassis containing the sampling circuits, reflecting circuits and power supplies is mounted under or back of the television receiver.

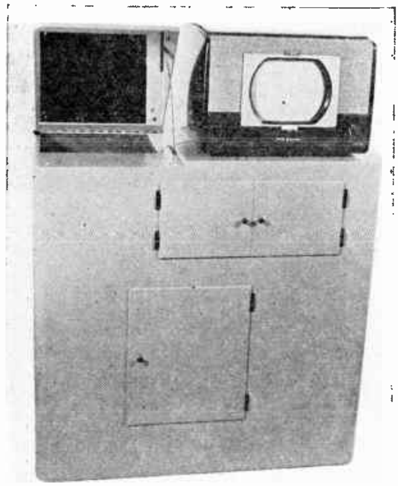

Fig. 16—Direct-view color converter.

Figure 17 shows this system as applied to a color converter, but the principles apply equally well for a color receiver.

Two-Color Systems

Color transmissions can be received on a simplified receiver which reproduces the picture in two colors only, instead of three. The two colors used are green-red and blue-green.

A block diagram of such a receiver is shown in Figure 18. This system is similar to that shown in Figure 4 except that only two video channels are required, and the method of sampling the composite signal is altered.

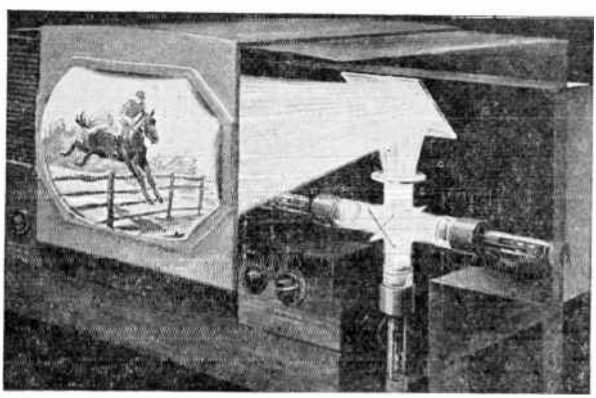

Fig. 17—Color converter using small projection kinescopes and refractive optics.

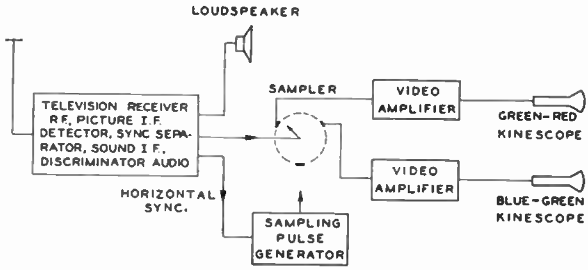

Fig. 18—Block diagram of two-color television receiver.

With reference to Figure 3, it was explained that, for the three-color system, the composite signal was sampled for green at the instant the red and blue components were passing through zero. In like manner, it was sampled for red and for blue when the other two colors in each case were zero. Figure 19 represents the same signals as Figure 2 and shows the different positions of the sampling pulses for a two-color picture-reproducing system. The composite signal is sampled for blue-green at a time when both blue and green components are present in a positive direction. This is indicated by the line labeled *B-G*. The composite signal is sampled for green-red at a time represented by the lines marked *G-R*. As indicated in Figure 19, no sample is taken at the third point. The sampling is repeated for each of the two color combinations once each 0.263 microsecond.

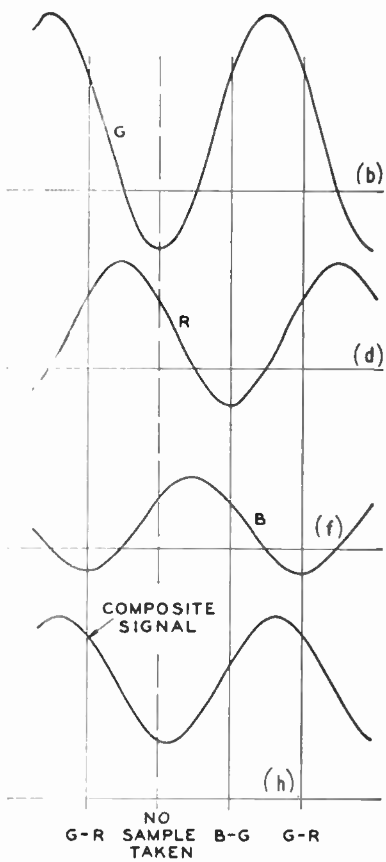

Fig. 19—Receiver sampling positions for the two-color system.

After the composite signal is sampled, the two color signals are amplified in separate video amplifiers having frequency cutoff characteristics as described in connection with the three-color receiver. They are then impressed on the grids of their respective kinescopes.

In the case of a color converter for an existing black-and-white receiver, the black-and-white kinescope in the receiver is used with a suitable color filter placed in front of it. Another kinescope is added and viewed through a dichroic mirror and suitable color filter.

The color converter employing this two-color system is shown in Figure 20. All of the components of the standard black-and-white receiver are used, including the deflecting circuits and second anode power supply. The only equipment that is added is the second kinescope and the sampling circuit. Connections between the television receiver and the color converter are few and are easily made. The foregoing points to the possibilities for a very low-cost color converter.

The principle of the two-color system is illustrated in Figure 21.

The two-color picture-reproducing system is also applicable to a simple and inexpensive color receiver. In this case, however, the two kinescopes will be made with the proper color phosphors and no filters are required.

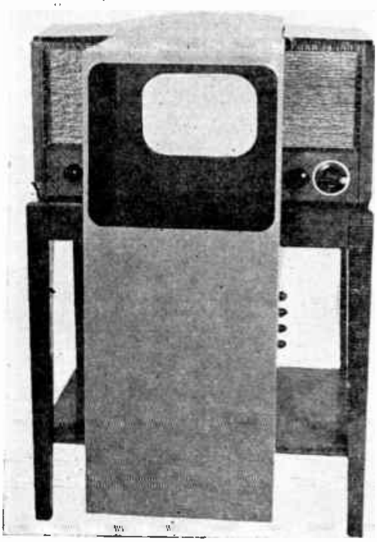

Fig. 20 — Color converter using two-color picture-reproducing system.

Fig. 21 — Two-color picture-reproducing system.

Summary of System Characteristics

The all-electronic color television system described herein is a fully compatible system, employing the current standards of 525 lines, sixty fields per second, and line interlacing. It provides a high-definition color (and black-and-white) picture in the standard six-megacycle channel through the use of the mixed-highs principle and time-multiplexing with picture dot interlacing.

The transmitted signal is consistent with the "Standards of Good Engineering Practice Concerning Television Broadcast Stations." This is the fundamental basis for compatibility and means that a current monochrome receiver will respond in the same way as it would if a standard black-and white camera originated the picture signal.

AN ANALYSIS OF THE SAMPLING PRINCIPLES OF THE DOT-SEQUENTIAL COLOR TELEVISION SYSTEM*

A Report

BY

RCA LABORATORIES DIVISION, PRINCETON, N. J.

Summary—This paper deals quantitatively with a number of aspects of the dot-sequential color television system, namely, the influence of sampling pulse width on color cross talk, the response of standard monochrome television receivers and color television receivers to sinusoidal variations and to step functions, the manner in which the method of mixed highs combines with the sampling procedure to produce high-frequency detail, and circuit methods of eliminating cross talk.

INTRODUCTION

A QUALITATIVE description of the sampling procedure and the use of mixed highs, as well as the dot-interlacing method of increasing detail used in the RCA dot-sequential color television system, has already been published.[1] The reader is referred to that paper for background material. A quantitative discussion of a number of aspects of the system is given in the following pages.

CROSS TALK AS A FUNCTION OF THE WIDTH OF THE SAMPLING PULSE

A block diagram of the color television broadcasting station is shown in Figure 1. The studio apparatus provides three electrical signals, one for each of the primary colors (green, red and blue). Each of these signals may contain frequency components out to at least four megacycles, and in addition an average or dc component.

For one signal routing of Figure 1, each color signal passes through a low-pass filter which eliminates frequency components above a frequency f_A megacycles. Where this paper deals with numerical values, f_A will be taken as 2.0 megacycles. The green-channel signal coming out of its particular low-pass filter is designated as G_L in Figure 1, indicating that at this point the signal contains the dc component and ac components with frequencies of f_A or less. The three low-frequency

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* Decimal Classification: R583.1.

[1] "A Six-Megacycle Compatible High-Definition Color Television System", *RCA Review*, Vol. X, No. 4, pp. 504-524, December, 1949.

signals, G_L, R_L, and B_L are then sent into an electronic commutator or sampler.

For the second signal routing of Figure 1, the three color signals from the camera are combined in electronic Adder No. 2 and then are passed through a band-pass filter. The output of this filter contains frequencies from f_A to f_B megacycles, with contributions from each of the three color channels. For calculation purposes, f_B has been taken as 4.1 megacycles. The signal at the output of the band-pass filter is designated as M_H, the mixed-high signal. The mixed high frequencies are fed to Adder No. 1 which is also receiving the signal from the electronic sampler.

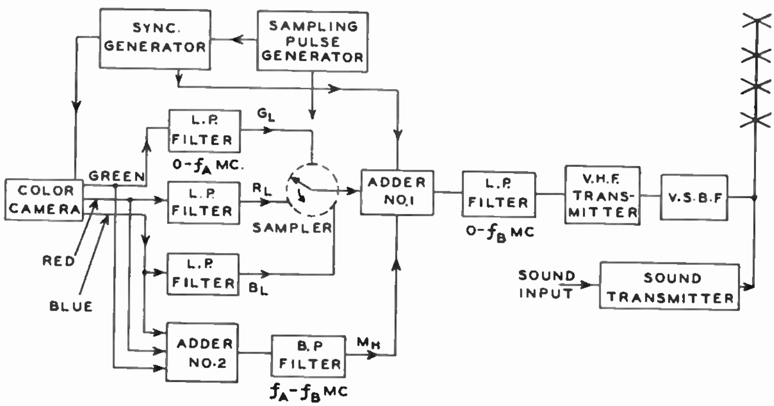

Fig. 1—Block diagram of the color television transmitter.

The frequency relationships in the system are depicted in Figure 2, with the following numerical values chosen for purposes of illustrative calculation:

f_0 = frequency of sampling pulse generator (3.8 megacycles),

f_A = upper limit of frequencies into the transmitter sampler and lower limit of mixed high frequencies (2.0 megacycles),

f_B = maximum frequency component transmitted by the system (4.1 megacycles). This upper limit may be determined by the receiver or transmitter cut-off characteristic, whichever is most restrictive.

$f_B - f_0$ = upper limit of frequencies free from inherent color cross talk without circuit devices (0.3 megacycle).

Figure 3 is a block diagram of one type of color television receiver. The output of the receiver sampler may go through separate video amplifiers to the picture reproducer, which may consist of three separate kinescopes, as indicated in FCC Exhibit 309,[2] or the composite signal may go from the second detector through a single video amplifier to a point where the three kinescope grids are tied in parallel. The keying or sampling is accomplished by applying short negative pulses in sequence to the cathodes of the kinescopes, as described in FCC Exhibit 316.[3]

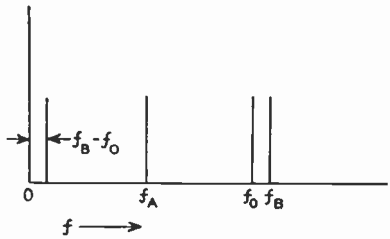

Fig. 2—Frequency relationships in the dot-sequential color television system.

f_B = upper frequency limit of system.

f_o = frequency of sampling of individual colors.

f_A = maximum frequency of signals into the transmitter sampler (also lower frequency limit of mixed highs).

$f_B - f_o$ = upper limit of frequencies free from inherent color cross talk without circuit devices.

The sampling procedure, at either the transmitter or the receiver, may be described in mathematical terms. Suppose that a signal G is applied to one grid of an electronic tube and that the tube has such characteristics that the output signal is always proportional to the signal G. In addition, a second grid is heavily biased except for regular periodic short intervals when this grid is driven to some prescribed positive value. The signal on this second grid thus acts as a gate on the signal G, and the output signal is proportional to signal G when the second grid is positive and the output signal is zero when the second grid is heavily biased.

The output signal may then be regarded as the product of the signal G and the gating signal. A representative gating signal is shown in Figure 4. The period or time between successive gates is T, while the duration of a gate pulse is ΔT. The sampling frequency, $f_0 = 1/T$. The duty factor of the gate may be defined as $F = \Delta T/T$. Then, if the output is proportional to a signal G, the Fourier series for the gated product is

$$G(t) = G \cdot F \left[1 + 2 \sum_{n=1}^{n=\infty} a_n \cos (n\omega_0 t) \right], \tag{1}$$

where $$a_n = \frac{\sin (n\pi F)}{n\pi F},$$

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[2] "A Three-Color Direct-View Receiver for the RCA Color Television System", Bulletin, RCA Laboratories Division, January 9, 1950.

[3] "A Simplified Receiver for the RCA Color Television System", Bulletin, RCA Laboratories Division, February 28, 1950.

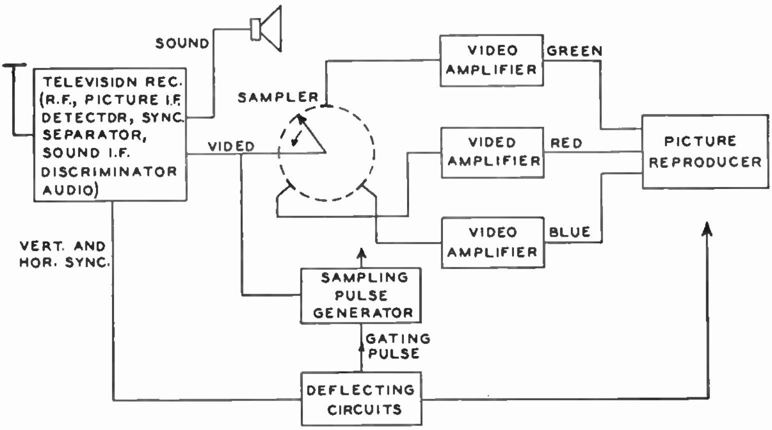


Fig. 3—Block diagram of one type of color television receiver.

and $\omega_0 = 2\pi f_0$.

The input signal G may be varying as a function of time, but for this first consideration of cross talk, G will be constant; that is, a flat green area is scanned.

The Fourier coefficients of the gating pulse shown in Figure 4 are displayed in Figure 5 for $n = 1$ and $n = 2$, as a function of the duty factor F.

Assume that the only signal from the color camera of Figure 1 is for the moment a dc signal from the green camera tube. After sampling at the transmitter, the signal at Adder No. 1 is

$$\frac{G}{3} [1 + 2 \sum_{n=1}^{n=\infty} a_n \cos (n\omega_0 t)]. \tag{2}$$

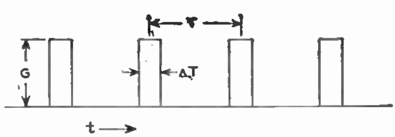


Fig. 4—Flat-top narrow sampling pulse.

$$G(t) = G \cdot F [1 + 2 \sum_{n=1}^{n=\infty} a_n \cos(n\omega_o t)],$$

where F = duty factor $= \frac{\Delta T}{T}$,

$$a_n = \frac{\sin(n\pi F)}{n\pi F}, \ \omega_o = 2\pi f_o, \ f_o = \frac{1}{T}.$$

Since the sampling frequency is 3.8 megacycles and the upper pass limit of the transmitter is considered to be 4.1 megacycles, only the fundamental term of the summation is retained. Then the signal out of the receiver second detector is

$$\frac{G}{3} [1 + 2a_1 \cos(\omega_0 t)]. \tag{3}$$

The sampling of this signal at the receiver for a single color channel may be obtained by multiplying (3) by the Fourier series of (1), but assuming a phase displacement of θ degrees. (Green channel, $\theta = 0°$; blue channel, $\theta = 120°$; red channel, $\theta = 240°$.) Hence the signal at a particular color kinescope is

$$\frac{G}{9}\,[1 + 2a_1\cos(\omega_0 t)]\,[1 + 2b_1\cos(\omega_0 t+\theta) + 2b_2\cos(2\omega_0 t+2\theta) + \cdots]$$

$$= \frac{G}{9}\,[1 + 2a_1 b_1\cos\theta$$

$$+ 2a_1\cos(\omega_0 t) + 2b_1\cos(\omega_0 t+\theta) + 2a_1 b_2\cos(\omega_0 t+2\theta)$$

$$+ 2a_1 b_1\cos(2\omega_0 t+\theta) + 2b_2\cos(2\omega_0 t+2\theta) + 2a_1 b_3\cos(2\omega_0 t+3\theta)$$

$$+ \cdots]. \tag{4}$$

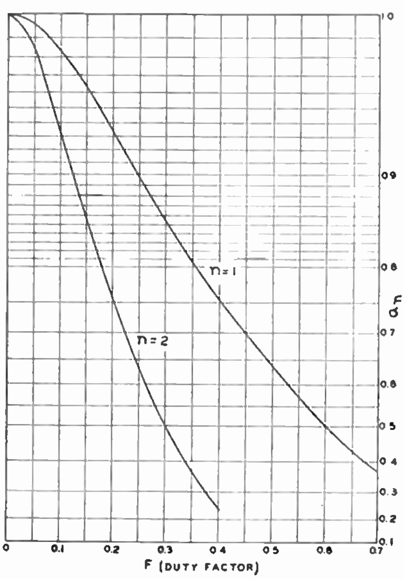


Fig. 5—Fourier coefficients of the sampling pulse of Figure 4 as a function of duty factor.

In (4), b_n has been used for the Fourier coefficients at the receiver sampling to avoid confusion with the a_n values used at the transmitter. Figure 5 applies equally well to b_1 and b_2 as it did to a_1 and a_2.

The terms containing $2\omega_0 t$ or greater may be dropped from (4), with the result

$$\frac{G}{9}\,[1 + 2a_1 b_1\cos\theta$$

$$+ 2a_1\cos(\omega_0 t) + 2b_1\cos(\omega_0 t+\theta)$$

$$+ 2a_1 b_2\cos(\omega_0 t+2\theta)]. \tag{5}$$

The signal on the green kinescope is obtained by setting θ equal to zero in the above expression which then becomes

$$\frac{G}{9}\,[1 + 2a_1 b_1 + 2(a_1+b_1+a_1 b_2)\cos(\omega_0 t)], \tag{6}$$

and the peak signal on the green kinescope (PS_g) is

$$PS_g = \frac{G}{9}\left[1 + 2(a_1b_1 + a_1 + b_1 + a_1b_2)\right]. \tag{7}$$

By setting θ equal to 120 degrees or 240 degrees, and following through the proper manipulation, one may find the peak signal on the red or the blue kinescope. Since the peak signals on the blue and the red kinescopes due to cross talk are equal in magnitude and shifted in time, it is necessary to examine only one of these signals. Cross talk (CT) may be defined as the ratio of the peak signal on the red kinescope to the peak signal on the green kinescope. Then

$$\mathrm{CT} = \frac{1 - a_1b_1 + 2\sqrt{a_1^2 + b_1^2 + a_1^2b_2^2 - a_1b_1 - a_1^2b_2 - a_1b_1b_2}}{1 + 2(a_1b_1 + a_1 + b_1 + a_1b_2)}. \tag{8}$$

Three combinations of duty factor choices are interesting to examine.

Case I. Duty factor of sampling at transmitter equal to duty factor of sampling at the receiver ($a_1 = b_1$).

Equation (8) then becomes

$$\mathrm{CT} = \frac{1 - b_1^2 + 2b_1(1 - b_2)}{1 + 2b_1(2 + b_1 + b_2)}. \tag{9}$$

The attendant cross talk is shown by the top curve of Figure 6.

Case II. Duty factor of sampling at transmitter very small ($a_1 = 1$).

Equation (8) then reduces to

$$\mathrm{CT} = \frac{1 - b_1 + 2\sqrt{1 + b_1^2 + b_2^2 - b_1 - b_2 - b_1b_2}}{1 + 2(1 + 2b_1 + b_2)}, \tag{10}$$

and the cross talk is shown by the middle curve of Figure 6. It may be seen that for a large duty factor at the receiver, the reduction in cross talk achieved by a short duty factor at the transmitter is small.

Case III. Duty factor of sampling at receiver very small ($b_1 = b_2 = 1$)

In this case, Equation (8) reduces to the very simple form

$$\mathrm{CT} = \frac{1 - a_1}{1 + 2a_1}. \tag{11}$$

This cross-talk condition is shown by the lower curve in Figure 6.

The analysis displayed in Figure 6 shows the importance of maintaining a short duty factor at the receiver sampler. Since it is possible to maintain the effect of a short duty cycle at the transmitter sampler both by gating control and circuit adjustment, it would appear that the middle curve of Figure 6 would be applicable for the actual receiver conditions. It may be seen that when the duty factor at the receiver is maintained at less than 0.15, the cross-talk signal remains at least 30 to 1 down from the desired signal.

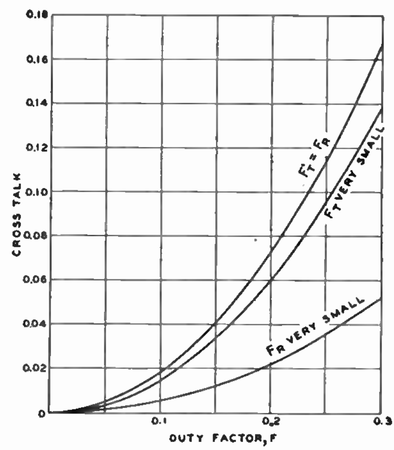


Fig. 6—Cross talk as a function of the duty factor of sampling.

Upper curve—duty factor of sampling at transmitter equal to duty factor of sampling at receiver.

Middle curve—duty factor of sampling at transmitter very small.

Lower curve—duty factor of sampling at receiver very small.

For the remainder of the analysis in this report, it will be assumed that the lessons pointed out by Figure 6 will be well learned. Hence $a_1 = b_1 = b_2 = 1$ will be used in the following analysis. To do otherwise would cloud the results in unnecessary rigor and would add little to the knowledge gained.

The Sampling Procedure Applied to Large Color Areas with a Sinusoidal Variation of the Color

a. A large green area with no variation

The green signal from the camera is assumed to be constant in magnitude, of value G. Figure 7(a) shows this fixed value, where G has been set equal to unity. Under the new assumptions ($a_1 = b_1 = b_2 = 1$), the signal to the transmitter modulator is given by Equation (3) as

$$\frac{G}{3}\left[1 \pm 2\cos(\omega_0 t)\right], \qquad (12)$$

where the plus sign applies for the first scan of the particular line and the minus sign applies to the second scan of the same line. This shift of the sampling by one half of the sampling cycle is accomplished by the methods described in FCC Exhibit 314.[4] The plot of Equation (12) for the two scans of the same line is shown in Figure 7(b).

The signal from the second detector or on the kinescope grid of a conventional black-and-white television receiver will also be given by Equation (12). Hence the solid line of Figure 7(b) may be regarded as the voltage applied to the kinescope grid of a black-and-white receiver during the first scan of a particular line, while the broken line is the corresponding voltage during the second scan of the same line.

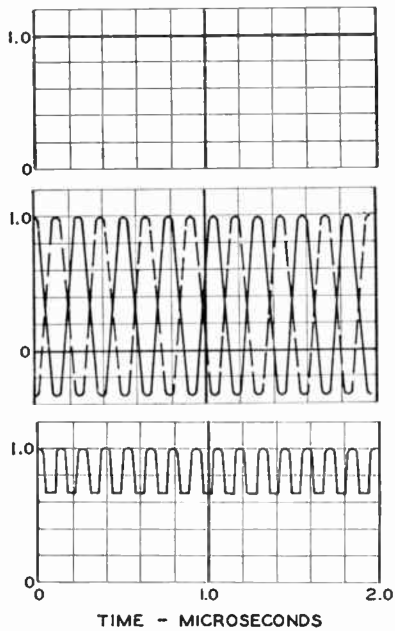


Fig. 7(a) (top)—Constant signal out of green camera tube.

(b) (middle) — Signal to transmitter modulator. Also the signal on the kinescope grid of a conventional black-and-white receiver, as well as the signal on the green kinescope grid of a color receiver.

(c) (bottom)—Combined light intensity of two successive scans of the same line on a conventional black-and-white receiver, and the combined light intensity of two successive scans of the same line on a color television receiver.

Assuming that the kinescope actually cuts off with negative applied signal, and neglecting the nonlinearity of the input control-voltage versus light-output characteristic of the kinescope, the solid line above the axis may be regarded as the effective light intensity along one line scan, while the portion of the dotted line above the axis may be regarded as the effective light intensity along the same line in the next scan. Since the second scan of the same line occurs only one-thirtieth of one second after the first scan, Talbot's law indicates that the light intensities may be added as far as the effect upon the eye is concerned. Figure 7(c) was constructed from the positive values of Figure 7(b) and may be regarded as the response on a conventional black-and-white receiver.

Turning now to the color receiver of Figure 3, the signal on the green kinescope may be obtained directly from Equation (6) by letting $a_1 = b_1 = b_2 = 1$. Then Equation (6) becomes Equation (12), and Figure 7(b) may now be regarded as the voltage on the green kinescope

[4] "Recent Developments in Color Synchronization in the RCA Color Television System", Bulletin, RCA Laboratories Division, February 9, 1950.

grid during the first and second scans of the same line, while Figure 7(c) depicts the light intensity distribution on one line of the green kinescope due to two successive scans of the line.

The signal on the grid of the red kinescope is determined by setting θ equal to 240 degrees in Equation (5), and for the assumed condition of narrow sampling which sets $a_1 = b_1 = b_2 = 1$, the result is identically zero. Similarly, by setting θ equal to 120 degrees, the signal on the grid of the blue kinescope is found to be zero. Hence, with narrow sampling of a dc signal which represents a flat field of a single color, there is no cross talk into the other two color channels.

b. $G + g \cdot \sin(\omega t)$ where $0 < f < f_B - f_o$

In this particular case, the green area is slowly varying so that the electrical signal is made up of a dc component and an ac component of frequency f where $0 < f < f_B - f_o$. This frequency region may be noted on Figure 2. For purposes of illustration, f_B has been chosen equal to 4.1 megacycles and f_o equal to 3.8 megacycles, hence the frequency of variation dealt with in this section must be less than 0.3 megacycle.

The signal out of the green camera tube is $G + g \cdot \sin(\omega t)$ where ω is $2\pi f$. This signal is sampled at the transmitter sampler in the fashion of Equation (3) so the signal at Adder No. 1 in Figure 1 is

$$[G + g \cdot \sin(\omega t)] \cdot \frac{1}{3} \cdot [1 + 2\cos(\omega_o t)]. \tag{13}$$

Equation (13) could be expanded to develop the sidebands generated by the product $\sin(\omega t) \cdot \cos(\omega_o t)$. It would be found that the sidebands have frequencies $f_o + f$ and $f_o - f$, both of which would pass through the filter and the transmitting system. Accordingly, there is no need to make the expansion for this case.

Equation (13) also represents the signal on the kinescope grid of a conventional black-and-white television receiver. Reversing the sign in the second bracket expression yields the equation for the second scanning of the same line.

When $G = 1$ and $g = 1/2$, the signal out of the green camera tube is $G + g \cdot \sin(\omega t) = 1 + 1/2 \sin(\omega t)$. The frequency has been taken as 0.2 megacycle. Figure 8(a) shows the signal out of the green camera tube for this condition.

Figure 8(b) shows a plot of Equation (13) for this same condition and may be regarded as the voltage on the kinescope of a conventional black-and-white receiver for two successive scans of the same line. Figure 8(c) shows a summation which depicts the effective light intensity on the same line of the black-and-white receiver tube.

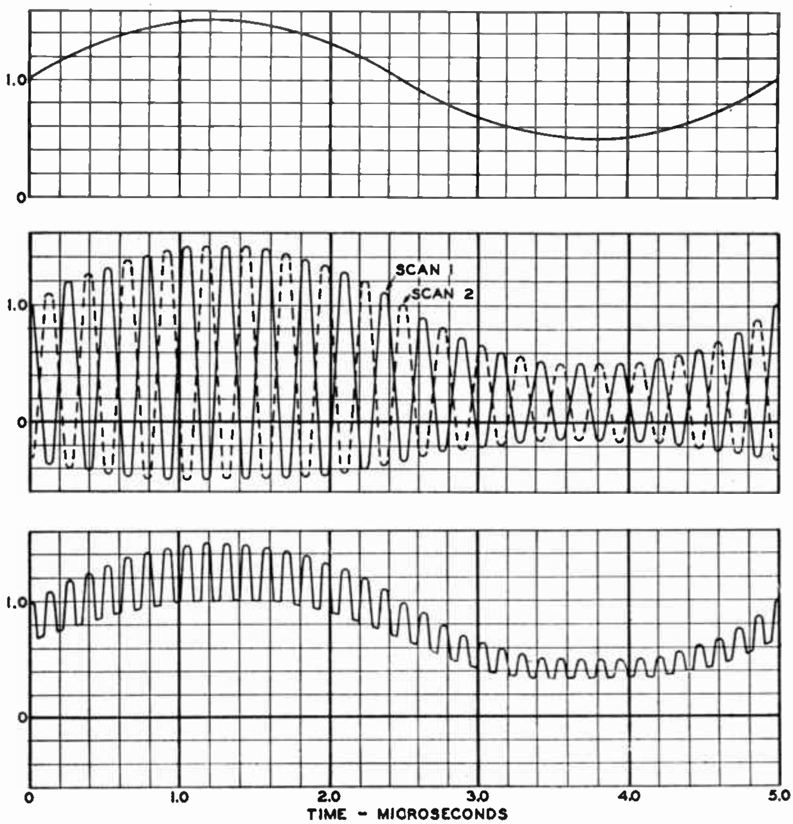


Fig. 8(a) (top)—Signal out of green camera tube; $f = 0.2$ megacycle, $G + g \cdot \sin(\omega t) = 1 + 1/2 \sin(\omega t)$.

(b) (middle)—Signal to transmitter modulator. Also the signal on the kinescope grid of a conventional black-and-white receiver, as well as the signal on the green kinescope grid of a color television receiver.

(c) (bottom)—Combined light intensity of two successive scans of the same line on a conventional black-and-white receiver, and the combined light intensity of two successive scans of the same line on a color television receiver.

Equation (13) is also the signal out of the second detector of the color receiver into the sampler. The sampling of this signal at the receiver for a single color channel may be obtained by multiplying (13) by the Fourier series of (1), but assuming a phase displacement of θ

degrees. Of course, a short sampling is assumed so that the Fourier coefficients are unity. Hence the signal at a particular color kinescope is

$$\begin{aligned}
&\frac{1}{9}\,[G+g\cdot\sin(\omega t)]\,[1+2\cos(\omega_o t)]\,[1+2\cos(\omega_o t+\theta) \\
&\qquad +2\cos(2\omega_o t+2\theta)+\cdots], \\
&=\frac{1}{9}\,[G+g\cdot\sin(\omega t)]\,[1+2\cos\theta \\
&\qquad +2\cos(\omega_o t)+2\cos(\omega_o t+\theta)+2\cos(\omega_o t+2\theta) \\
&\qquad +2\cos(2\omega_o t+\theta)+2\cos(2\omega_o t+2\theta)+2\cos(2\omega_o t+3\theta) \\
&\qquad +2\cos(3\omega_o t+2\theta)+2\cos(3\omega_o t+3\theta)+2\cos(3\omega_o t+4\theta) \\
&\qquad +\cdots].
\end{aligned} \tag{14}$$

Now to find the signal on the green kinescope grid, simply let θ equal zero in (14), which reduces to

$$[G+g\cdot\sin(\omega t)]\cdot\frac{1}{3}\cdot[1+2\cos(\omega_o t)]. \tag{15}$$

Since (15) is identical with (13), it is seen that Figure 8(b) may be regarded as the voltage applied to the kinescope of the green tube in the color receiver for two successive scans of the same line, and Figure 8(c) may be regarded as the equivalent light intensity variation for two scans of the same line.

To find the signal on the grid of the blue kinescope, set θ equal to 120 degrees in Equation (14) and it will be seen that the second bracketed expression goes to zero. If θ is then set equal to 240 degrees, an identical result is found, indicating no signal on the red tube.

Hence, when the frequency of variation is less than $f_B - f_o$, there is no cross talk and the single-color field is reproduced correctly in magnitude and position by the sampling procedure.

c. $G+g\cdot\sin(\omega t)$ where $f_B - f_o < f < f_A$

In this case, the green area is varying so that the electrical signal is made up of a dc component and an ac component of frequency f, where $f_B - f_0 < f < f_A$. This frequency region may be noted on Figure 2, and for illustrative purposes lies between 0.3 megacycle and 2.0 megacycles.

The signal out of the green camera tube is $G+g\cdot\sin(\omega t)$. For purposes of illustration, f has been chosen to be 1.6 megacycles, $G=1$ and $g=1/2$. Figure 9(a) shows this signal, $1+1/2\sin(\omega t)$.

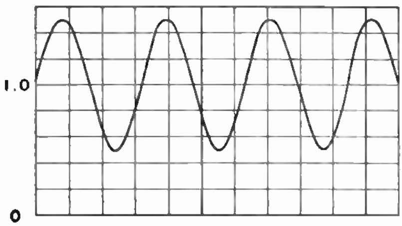


Fig. 9(a) — Signal out of green camera tube; $f = 1.6$ megacycles, $G + g \cdot \sin(\omega t) = 1 + 1/2 \sin(\omega t)$.

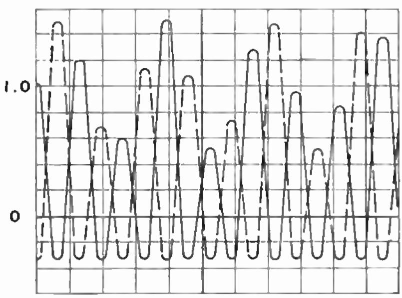


Fig. 9(b) — Signal to transmitter modulator. Also the signal on the kinescope grid of a conventional black-and-white receiver.

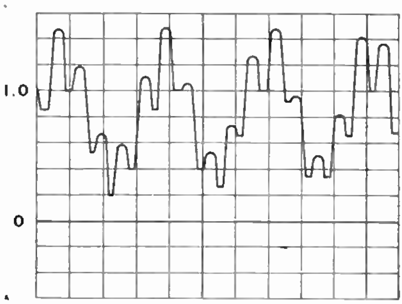


Fig. 9(c)—Combined light intensity of two successive scans of the same line on a black-and-white receiver.

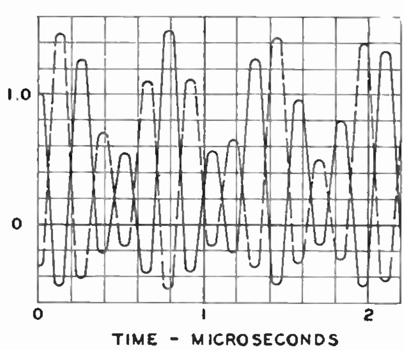


Fig. 9(d) — Signal on the green kinescope grid of a color television receiver.

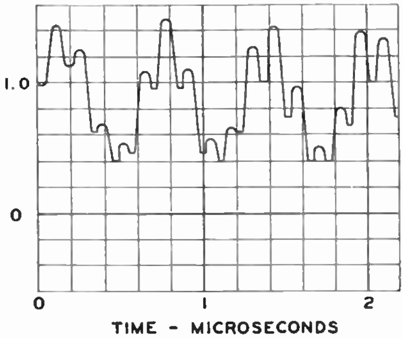


Fig. 9(e)—Combined light intensity of two successive scans of the same line on the green kinescope of a color television receiver.

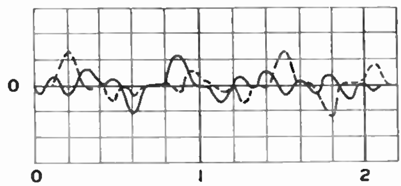


Fig. 9(f) — Cross talk voltage on red kinescope grid.

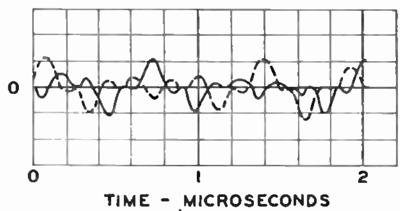


Fig. 9(g) — Cross talk voltage on blue kinescope grid

The signal from the green camera tube is sampled so that the signal at Adder No. 1 in Figure 1 is

$$[G+g\cdot\sin(\omega t)]\cdot\frac{1}{3}[1+2\cos(\omega_o t)]$$

$$=\frac{g}{3}[1+2\cos(\omega_o t)]+\frac{g}{3}\sin(\omega t)-\frac{g}{3}\sin(\omega_o-\omega)t+\frac{g}{3}\sin(\omega_o+\omega)t. \quad (16)$$

Since $f > f_B - f_o$, $f_o + f > f_B$ and the last term in (16) is lost in going through the final filter before the transmitter in Figure 1. (This filter is not necessarily a physical reality, but serves the purpose of specifying the upper limit of frequencies that may be transmitted. It is likely that the upper frequency restriction will be imposed by the receiver rather than the transmitter.) Hence the signal at the modulator is

$$\frac{G}{3}[1+2\cos(\omega_o t)]+\frac{g}{3}\sin(\omega t)-\frac{g}{3}\sin(\omega_o-\omega)t. \quad (17)$$

Inspection of (17) shows that the loss of one of the terms with a coefficient $g/3$ has made it impossible for (17) to reproduce the desired variation in correct amplitude. This condition may be corrected by altering the response characteristics of the low-pass filters preceding the sampler in the transmitter of Figure 1. The filters should have a response in the region $f_B - f_o < f < f_A$ which is 1.5 times the gain in the region $0 < f < f_B - f_o$. Under this new condition, (17) becomes

$$\frac{G}{3}\left[1+2\cos(\omega_o t)\right] + \frac{g}{2}\sin(\omega t) - \frac{g}{2}\sin(\omega_o - \omega)t. \qquad (18)$$

The condition where $G = 1$ and $g = 1/2$, computed from (18), is shown in Figure 9(b) for two successive line scans. These curves apply to the modulator signal in the transmitter and to the signal on the kinescope grid of a black-and-white receiver. Figure 9(c) shows the effective light intensity due to two scans of the same line on a black-and-white receiver.

Equation (18) may also be regarded as the signal into the sampler of the color receiver of Figure 3. The previously used expedient of multiplying by the generalized sampling function may now be resorted to just as was done in obtaining Equation (14). The result of sampling (18) is

$$\frac{G}{9}\left[1+2\cos(\omega_o t)\right]\left[1+2\cos(\omega_o t+\theta) + 2\cos(2\omega_o t+2\theta) + \cdots\right]$$
$$+\frac{1}{3}\left[\frac{g}{2}\sin(\omega t) - \frac{g}{2}\sin(\omega_o - \omega)t\right]\left[1+2\cos(\omega_o t+\theta)\right.$$
$$\left.+ 2\cos(2\omega_o t+2\theta) + \cdots\right]$$
$$= \frac{G}{9}\left[1+2\cos(\omega_0 t)\right]\left[1+2\cos(\omega_0 t+\theta) + 2\cos(2\omega_0 t+2\theta) + \cdots\right]$$
$$+\frac{g}{6}\left[\sin(\omega t) + \sin(\omega t+\theta)\right.$$
$$+ \sin([\omega_o+\omega]t+\theta) + \sin([\omega_o+\omega]t+2\theta)$$
$$\left.- \sin([\omega_o-\omega]t+\theta) - \sin(\omega_o-\omega)t + \cdots\right]. \qquad (19)$$

Now to find the signal on the green kinescope, let $\theta = 0$ in (19) and find

$$\frac{1}{3}\left[G+g\cdot\sin(\omega t)\right]\left[1+2\cos(\omega_o t)\right]. \qquad (20)$$

The condition where $G = 1$ and $g = 1/2$, computed from (20), is shown in Figure 9(d) which depicts the voltage on the grid of the green kinescope for two successive line scans. Figure 9(e) shows the effective light intensity due to two scans of the same line on the green kinescope.

The signal on the grid of the red kinescope (due to color cross talk) is found by setting $\theta = 240$ degrees in Equation (19), with the result

$$\frac{g}{6}\sin(\omega t - 60°)\,[1 + 2\cos(\omega_o t - 120°)] \text{ (on red kinescope)}, \qquad (21)$$

and setting $\theta = 120$ degrees gives

$$\frac{g}{6}\sin(\omega t + 60°)\,[1 + 2\cos(\omega_o t + 120°)] \text{ (on blue kinescope)}. \qquad (22)$$

The voltage on the grid of the red kinescope due to the erroneous sampling of the green signal is shown in Figure 9(f) as computed from (21), while Figure 9(g) shows the signal on the grid of the blue kinescope.

These equations show that cross talk up to fifty per cent is possible in the region where the frequency is greater than $f_B - f_0$ and less than f_A, or in the example, when the frequency lies between 0.3 megacycle and 2.0 megacycles.

At first glance, this degree of cross talk might seem intolerable. In the case shown in Figure 9, it is likely that non-linearity of the light-output versus grid-voltage characteristic of the kinescopes would make the cross talk of negligible importance. In the converse case, if the average intensity of the red tube were high, the erroneous voltage of Figure 9(f) might be enhanced to a point where cross talk produced undesirable effects. While this cross talk has not appeared to be a serious problem in the dot-sequential color television system, means for eliminating the effect will be described later in this report.

d. $G + g \cdot \sin(\omega t)$ *where* $f_A < f < f_B$

In this case, the green area is varying so that the electrical signal is made up of a dc component and an ac component of frequency f where $f_A < f < f_B$. This frequency region may be noted on Figure 2, and for illustrative purposes lies between 2.0 and 4.1 megacycles.

The signal from the green camera tube is $G + g \cdot \sin(\omega t)$. For purposes of illustration, f has been chosen as 3.4 megacycles, $G = 1$ and $g = 1/2$. Figure 10(a) shows this signal $1 + 1/2 \sin(\omega t)$.

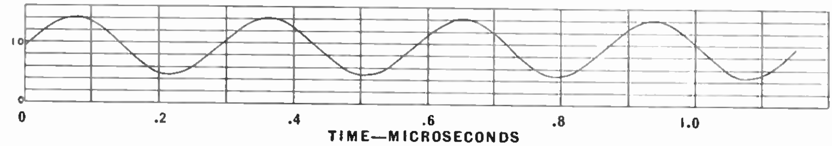


Fig. 10(a)—Signal out of green camera tube; $f = 3.4$ megacycles, $G + g \cdot \sin(\omega t) = 1 + 1/2 \sin(\omega t)$.

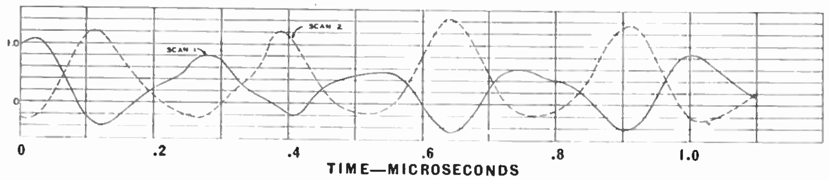


Fig. 10(b)—Signal on the green kinescope grid of a color television receiver.

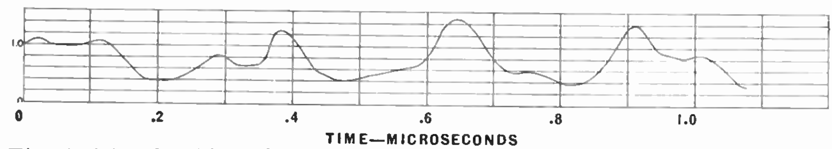


Fig. 10(c)—Combined light intensity of two successive scans of the same line on the green kinescope of a color television receiver.

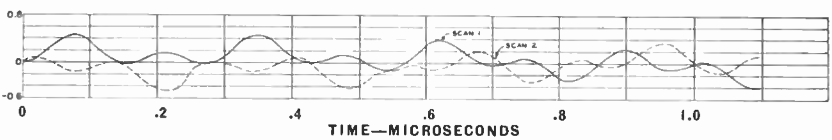


Fig. 10(d)—Voltage on the red kinescope grid.

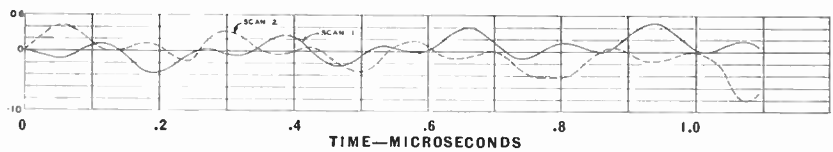


Fig 10(e)—Voltage on the blue kinescope grid.

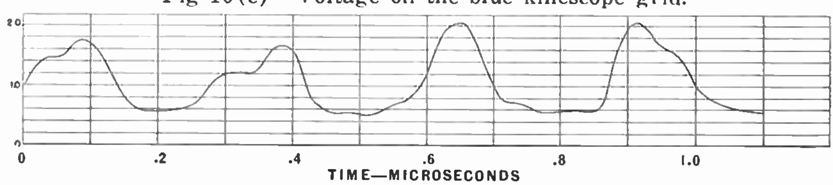


Fig 10(f)—Combined light intensity of two successive scans of the same line, obtained by adding light intensities of the green, red and blue tubes.

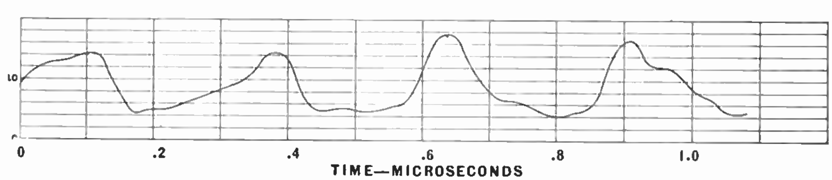


Fig. 10(g)—Combined light intensity of two successive scans of the same line, obtained by adding light intensities of the green and red tubes.

The dc signal G goes through the transmitter sampler, but since the ac term is of a frequency lying in the region committed to "mixed-highs," this latter signal goes through Adder No. 2 and the appropriate band-pass filter into Adder No. 1 of Figure 1. Hence the signal into the transmitter modulator is

$$\frac{G}{3}\left[1+2\cos(\omega_o t)\right] + g \cdot \sin(\omega t). \tag{23}$$

Equation (23) also applies to the voltage on the kinescope grid of a black-and-white receiver. The background term is sampled while the mixed-high signal, unsampled, is superimposed to supply fine detail.

The signal out of the second detector of a color receiver also has the form of Equation (23). Sampling in the receiver results in a signal on the grid of the green kinescope of the form

$$\frac{1}{3}\left[G+g \cdot \sin(\omega t)\right]\left[1+2\cos(\omega_o t)\right]. \tag{24}$$

A plot of this equation is shown by Figure 10(b), while Figure 10(c) shows the combined light intensity on a single line of the green tube for two successive scans of the same line. This latter plot shows the effect of the beat between the high-frequency component and the sampling frequency, so that Figure 10(c) is not a very faithful reproduction of Figure 10(a).

The output of the red sampler (the voltage on the grid of the red kinescope) is

$$\frac{g}{3}\sin(\omega t)\left[1+2\cos(\omega_o t-120°)\right], \tag{25}$$

while the voltage on the grid of the blue kinescope is

$$\frac{g}{3}\sin(\omega t)\left[1+2\cos(\omega_o t+120°)\right]. \tag{26}$$

The voltage on the red kinescope is shown in Figure 10(d), while the voltage on the blue kinescope is given by Figure 10(e).

It is obvious that the high-frequency signal mixing in this region is one hundred per cent, since the philosophy of the principle of "mixed-highs" has already been accepted, and the high frequency components of the three camera tubes have been deliberately combined at the transmitter so that these signals have completely lost color identity. Because of the inability of the eye to see color in the fine detail, it is possible to combine the positive values of Figures 10(d) and 10(e) with Figure 10(c) with the result shown in Figure 10(f), which is a more satisfactory representation of Figure 10(a). However, it is well known that the resolution of the eye is very poor in blue, so it seems more fair to combine only Figure 10(d) with Figure 10(c), with Figure 10(g) resulting. The improvement in reproduction of Figure 10(a) by Figure 10(g) is striking, particularly when it is recalled that the rate of the variations in Figure 10(a) is at a high frequency, and that the differences between Figure 10(a) and Figure 10(g) represent still higher frequency components, which are beyond the limits of ordinary resolution.

The Sampling Procedure Applied to Step Functions of Light Intensity

a. Response of conventional black-and-white television receivers

In the preceding pages, attention has been given to a color intensity change which produces an electrical signal consisting of a dc term and a single ac component. Such an analysis serves to demonstrate the detailed mechanism of the system. However, it is not a condition often encountered in producing an actual television image. Generally, it is more interesting to examine the action of the system near edges of objects in order to determine rise time, overshoot, and color cross talk.

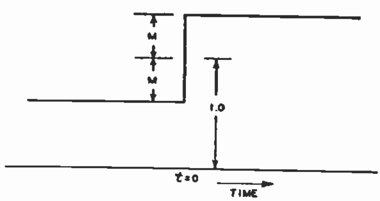


Fig. 11—Idealized step function of voltage which has a value of $1 - M$ for time less than zero and a value of $1 + M$ for time greater than zero.

For purpose of analysis, assume that the voltage coming from the green camera tube has the form shown in Figure 11, where the voltage has the value $1 - M$ for all values of time less than zero, and has the value of $1 + M$ for all times greater than zero. The function of Figure 11 may be produced exactly only when the associated circuits have unlimited frequency response. For this idealized condition, the signal from the green camera tube is given by the following Fourier Integral:

Green camera signal (G.C.S.)

$$= 1 + \frac{2M}{\pi} \int_{\beta=0}^{\beta=\infty} d\beta \int_{\omega=0}^{\omega=\infty}$$

where $\beta =$ an integration variable,

$\omega = 2\pi f$,

$f =$ a frequency component lying between zero and infinity,

$t =$ the instant of time at which the signal is to be evaluated.

Now assume that a circuit is imposed which has unity gain for all frequencies below f_B and zero response for all frequencies above f_B, and with no appreciable phase shift. Then Equation (27) becomes

$$\text{G.C.S.} = 1 + \frac{2M}{\pi} \int_{\beta=0}^{\beta=\infty} d\beta \int_{\omega=0}^{\omega=2\pi f_B} \sin(\omega\beta)\sin(\omega t) \cdot d\omega$$

$$= 1 + \frac{2M}{\pi} \cdot \text{Si}(2\pi f_B t), \qquad (28)$$

where $\text{Si}(x) =$ Integral sine of $x = \int_0^x \frac{\sin u}{u} \cdot du$, a well-known and tabulated function.

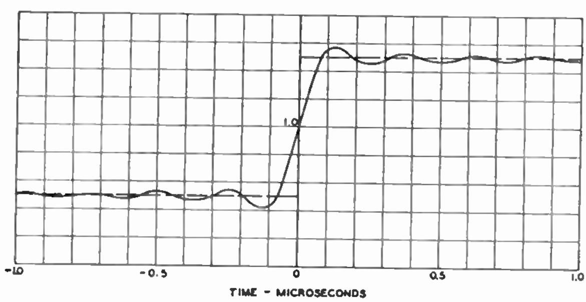


Fig. 12—Response to the step function of Figure 11 ($M = 1/2$) for a circuit which has unity gain for all frequencies below 4.1 megacycles, zero response above 4.1 megacycles, and linear phase shift.

Figure 12 shows a plot of Equation (28) with $M = 1/2$ and with the upper frequency limit f_B equal to 4.1 megacycles. It should be noted that Figure 12 and not Figure 11 should be used in judging the response when other circuit factors have been added.

To find the signal at the transmitter modulator, use may be made of previously developed material to operate upon the term $\sin(\omega t)$ in Equation (28). Three frequency regions must be considered.

First, when $0 < f < f_B - f_o$, Equation (13) teaches that $\sin(\omega t)$ becomes $\frac{\sin(\omega t)}{3}[1+2\cos(\omega_o t)]$ at the transmitter modulator.

Secondly, when $f_B - f_o < f < f_A$, Equation (18) shows that $\sin(\omega t)$ becomes $1/2\sin(\omega t) - 1/2\sin(\omega_o - \omega)t$ at the transmitter modulator.

Then also in the region where $f_A < f < f_B$, Equation (23) shows that $\sin(\omega t)$ is unchanged.

Equation (12) tells that the unit value in Equations (27) and (28) becomes simply 1/3 $[1+2\cos(\omega_0 t)]$.

When these operations are performed on Equation (27), the signal into the transmitter modulator (T.M.S.) is

$$\text{T.M.S.} = \frac{1}{3}[1+2\cos(\omega_o t)]$$

$$+\frac{2M}{3\pi}[1+2\cos(\omega_o t)]\int_{\beta=0}^{\beta=\infty} d\beta \int_{\omega=0}^{\omega=2\pi(f_B-f_o)} \sin(\omega\beta)\sin(\omega t)\,d\omega$$

$$+\frac{M}{\pi}\int_{\beta=0}^{\beta=\infty} d\beta \int_{\omega=2\pi(f_B-f_o)}^{\omega=2\pi f_A} \sin(\omega\beta)\sin(\omega t)\,d\omega$$

$$-\frac{M}{\pi}\int_{\beta=0}^{\beta=\infty} d\beta \int_{\omega=2\pi(f_B-f_o)}^{\omega=2\pi f_A} \sin(\omega\beta)\sin(\omega_o-\omega)t \cdot d\omega$$

$$+\frac{2M}{\pi}\int_{\beta=0}^{\beta=\infty} d\beta \int_{\omega=2\pi f_A}^{\omega=2\pi f_B} \sin(\omega\beta)\sin(\omega t)\,d\omega. \quad (29)$$

Integration and combination of terms yields

$$\text{T.M.S.} = \tfrac{1}{3}\{1 \pm \cos(\omega_0 t)\}\left\{1+\frac{2M}{\pi}\text{Si}\,[2\pi(f_B-f_0)t]\right\}$$

$$+\frac{M}{\pi}\{2\text{Si}(2\pi f_B t) - \text{Si}(2\pi f_A t) - \text{Si}\,2\pi(f_B-f_0)t\}$$

$$\pm\frac{M}{\pi}\cos(\omega_0 t)\,\{\text{Si}(2\pi f_A t) - \text{Si}[2\pi(f_B-f_0)t]\}$$

$$\pm\frac{M}{\pi}\sin(\omega_0 t)\,\{\text{Ci}[2\pi(f_B-f_0)t] - \text{Ci}(2\pi f_A t)\}. \quad (30)$$

where $\text{Ci}(x)$ = Integral cosine of $x = -\int_x^{\infty}\frac{\cos u}{u}\,du$. Where the $\pm$ signs appear in Equation (30), the plus sign applies to the first scan of the line and the minus sign applies to the second scan of the same line.

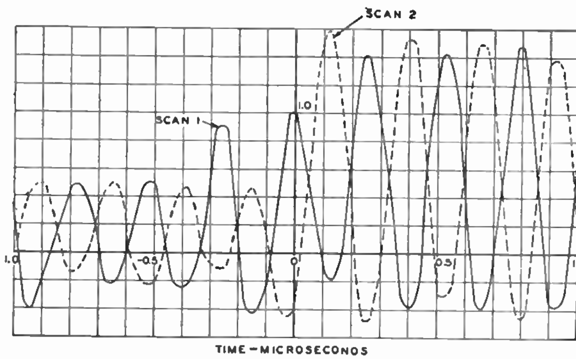


Fig. 13 — Transmitter modulator signal for two successive scans of the same line. Also the signal on the kinescope grid of a conventional black-and-white receiver.

It should be noted that Equation (30) is also the voltage appearing on the kinescope grid in a black-and-white receiver. Figure 13 is a plot of Equation (30) for the following conditions:

$$M = \tfrac{1}{2},$$

$$f_B = 4.1 \text{ megacycles},$$

$$f_0 = 3.8 \text{ megacycles},$$

$$f_A = 2.0 \text{ megacycles},$$

$$f_B - f_0 = 0.3 \text{ megacycle},$$

and may be considered as the transmitter modulator signal for two successive scans of the same line, as well as the signal on the kinescope grid of a conventional black-and-white receiver. Figure 14 shows the combined light intensity of two successive scans of the same line on a black-and-white receiver. This latter figure, constructed graphically from Figure 13, shows close agreement with Figure 12.

b. Response of a color television receiver

To find the signal on the kinescopes of a color television receiver,

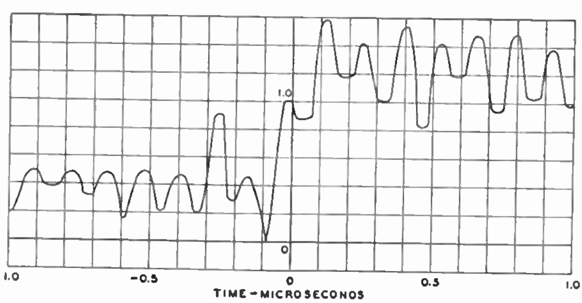


Fig. 14 — Combined light intensity of two successive scans of the same line on a black-and-white receiver.

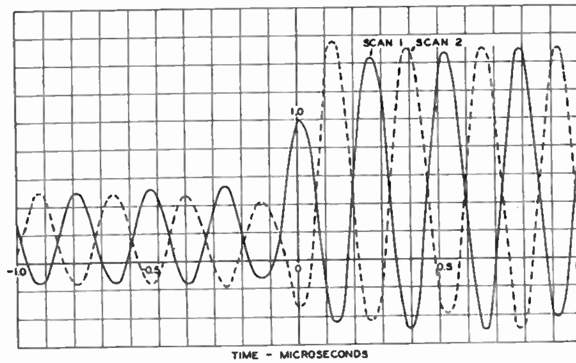


Fig. 15 — Signal on the green kinescope grid of a color television receiver.

the operating procedure on the $\sin(\omega t)$ term of Equation (27) is determined by referring to Equations (15), (20), and (24). These equations show that $\sin(\omega t)$ is converted to $\frac{\sin(\omega t)}{3}[1+2\cos(\omega_0 t)]$ in all frequency regions up to f_R so the signal on the green kinescope (G.K.S.) is simply

$$\text{G.K.S.} = \tfrac{1}{3}\,[1 \pm 2\cos(\omega_0 t)] \left[1 + \frac{2M}{\pi}\,\mathrm{Si}(2\pi f_R t) \right], \qquad (31)$$

where the $\pm$ signs apply to the first and second scans of the same line.

Comparison of Equations (28) and (31) shows that the signal on the green kinescope is a perfect reproduction of the green camera signal multiplied by the sampling function. Figure 15 shows a plot of Equation (31), the signal on the green kinescope grid of a color television receiver, for the same frequency restrictions used in the previous calculations, while Figure 16 shows the combined light intensity of two successive scans of the same line on the green kinescope of a color television receiver.

The cross talk may be deduced by using Equations (21) and (25)

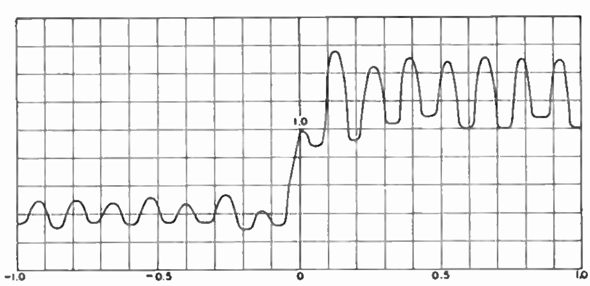


Fig. 16 — Combined light intensity of two successive scans of the same line on the green kinescope of a color television receiver.

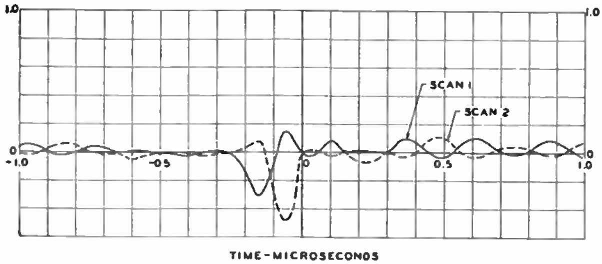


Fig. 17 — Cross-talk voltage on the red kinescope grid for two successive scans of the same line.

for the red tube cross talk and Equations (22) and (26) for the blue tube cross talk in conjunction with Equation (27). Then the red kinescope signal (R.K.S.) is

$$\text{R.K.S.} = \frac{2M}{\pi} \{1 \pm 2\cos(\omega_0 t - 120°)\} \{⅓ \,\text{Si}(2\pi f_B t) - ¼ \,\text{Si}(2\pi f_A t)$$

$$- \frac{1}{12} \text{Si}[2\pi(f_B - f_0)t] + \frac{\sqrt{3}}{12} \text{Ci}[2\pi(f_B - f_0)t] - \frac{\sqrt{3}}{12} \text{Ci}(2\pi f_A t)\}, \quad (32)$$

and the blue kinescope signal (B.K.S.) is

$$\text{B.K.S.} = \frac{2M}{\pi} \{1 \pm 2\cos(\omega_0 t + 120°)\} \{⅓ \,\text{Si}(2\pi f_B t) - ¼ \,\text{Si}(2\pi f_A t)$$

$$- \frac{1}{12} \text{Si}[2\pi(\text{f}_B - f_0)t] - \frac{\sqrt{3}}{12} \text{Ci}[2\pi(f_B - f_0)t] + \frac{\sqrt{3}}{12} \text{Ci}(2\pi f_A t)\}. \quad (33)$$

Figure 17 shows the cross-talk voltage on the red kinescope grid for two successive scans of the same line, while Figure 18 displays the combined light intensity of two successive line scans of the same line on the red kinescope.

Figures 19 and 20 show corresponding effects for the blue kinescope.

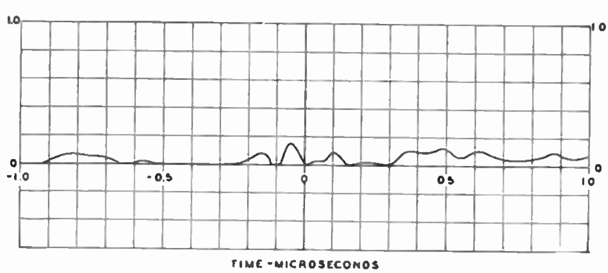


Fig. 18—Combined light intensity of two successive line scans of the same line on the red kinescope.

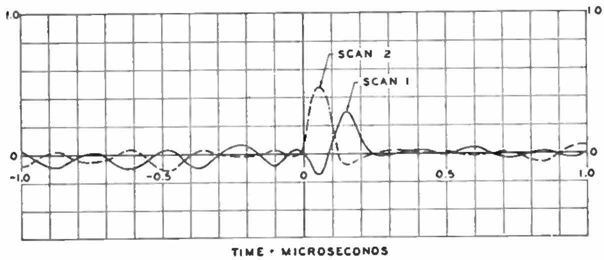


Fig. 19 — Cross-talk voltage on the blue kinescope grid for two successive line scans of the same line on the blue kinescope.

CROSS-TALK ELIMINATION BETWEEN THE GREEN AND RED CHANNELS WHEN $f_B - f_0 < f < f_A$

It was observed in Section III, where the frequency of the signal component lies between $f_B - f_0$ and f_A, that the color cross-talk terms may be up to fifty per cent of the desired terms. While it has not yet been clearly established that it is necessary to reduce or eliminate this cross talk, rather simple circuit expedients are possible to completely eliminate the cross talk. It should be remembered that the response in the mixed-high region has not been considered to be cross talk, since the crossing of signals in this region has been regarded as entirely legitimate.

a. Simple modification of the transmitter sampler

As a first step in describing a number of possibilities, a simple modification of the transmitter sampler may first be considered. Figure 21 shows the part of Figure 1 which has been changed somewhat. The mixed-high circuits have not been changed and are not shown. The low-pass filters in the red and the green channels, as before, pass frequencies up to f_A, but now have unity gain up to $f_B - f_0$ and have a gain of 2.0 from this frequency up to f_A. The low-pass filter in the blue channel may cut-off at $f_B - f_0$, since the eye is very poor in resolving power in the blue.

A band-pass filter and phase shifter connect the output of the low-pass filter in the green channel to the input of the red sampler. This

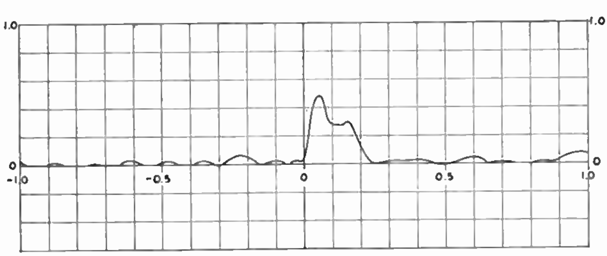


Fig. 20—Combined light intensity of two successive line scans of the same line on the blue kinescope.

channel passes frequencies between $f_B - f_0$ and f_A, the region where it is desired to eliminate cross talk. The signal from the green channel to the red sampler position is made one half of the signal going to the green sampler position. In addition, all the frequency components are advanced 120 degrees in phase in passing through the circuit. This latter condition is quite easily brought about by a double modulating and filtering process. The corresponding element going from the red channel to the green sampling position has the same characteristics except that the phase of the components is retarded by 120 degrees.

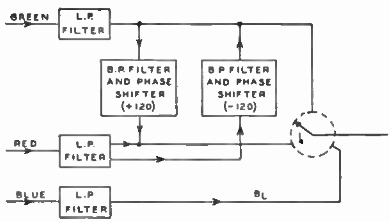


Fig. 21 — Circuit modifications at the transmitter to eliminate cross talk between the red and green channels.

With the signal $g \cdot \sin(\omega t)$ coming from the green camera, the signal going into the green sampler position at the transmitter is $2g \cdot \sin(\omega t)$. This signal is sampled by the function $\frac{1}{3}[1 + 2\cos(\omega_0 t)]$ and becomes $\frac{2g}{3} \cdot \sin(\omega t) - \frac{2g}{3} \sin(\omega_0 - \omega)t$. The signal into the red sampler (from the green channel through the phase shifter) is $g \cdot \sin(\omega t + 120°)$. This signal is sampled by the function $\frac{1}{3}[1 + 2\cos(\omega_0 t - 120°)]$ yielding a signal out of the red sampler of $\frac{g}{3}\sin(\omega t + 120°) - \frac{g}{3}\sin[(\omega_0 - \omega)t - 240°]$. The total signal into the modulator is

$$\frac{g}{\sqrt{3}}\{\sin(\omega t + 30°) - \sin[(\omega_0 - \omega)t + 30°]\}. \qquad (34)$$

When Equation (34) is sampled at the color receiver by the green sampler, using the sampling function

$$\frac{1}{3}[1 + 2\cos(\omega_0 t) + 2\cos(2\omega_0 t) + ---],$$

the signal on the green kinescope grid becomes

$$\frac{g \sin(\omega t)}{3}[1 + 2\cos(\omega_0 t)].$$

However, when Equation (34) is sampled by the red sampler at the receiver, using the sampling function

$$\frac{1}{3}\left[1+2\cos(\omega_o t-120°)+2\cos(\omega_o t-240°)+---\right],$$

the signal on the red kinescope grid becomes identically zero.

Thus a method of completely eliminating the cross talk between the red and the green channels in the frequency region above $f_B - f_o$ and below f_A has been displayed.

b. Addition of a low-pass filter to the color receiver

The additions of Figure 21 may be added to the transmitter without a single change in the receiver of Figure 3. If color receivers of this type were in operation in the field, the changes in the transmitter shown in Figure 21 could be made without altering a single receiver. The immediate effect would be an elimination of cross talk in the region in question between the green and red channels. The cross talk of red and green into the blue channel would be unchanged, but because of the high-frequency nature would probably be of no consequence. Cross talk of the blue into red or green would be eliminated by restricting the components of the blue signal to frequencies less than $f_B - f_o$ by means of the low-pass filter in the blue channel preceding the transmitter sampler.

As another experiment to investigate the matter of reduction of cross talk, a low-pass filter could be inserted in the video amplifier circuit leading to the blue kinescope in Figure 3. This filter would also remove the f_o sampling component in the blue channel.

Before proceeding with an examination of other circuit details, it may prove interesting to see what has happened to the step function response for the receiver and transmitter condition described in this section.

The desired response of the signal at the green kinescope has remained unchanged and is given by Equation (31) and by Figures 15 and 16. The cross-talk conditions have changed, however. For instance, in the red channel, the only signal mixing components are those that have been placed there deliberately by the use of mixed highs. The signal on the red kinescope grid due to the step function in the green channel is

$$\text{R.K.S.} = \frac{2M}{3\pi}\left[1\pm 2\cos(\omega_0 t-120°)\right]\left[\text{Si}(2\pi f_B t) - \text{Si}(2\pi f_A t)\right]. \qquad (35)$$

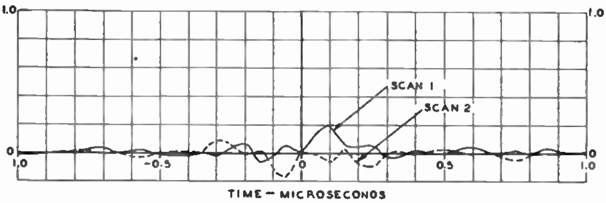


Fig. 22 — Signal on grid of red kinescope due to step function in the green channel and the use of mixed highs.

Figure 22 shows the signal corresponding to (35) for two scans of the same line, with $M = 1/2$, $f_B = 4.1$ megacycles, $f_A = 2.0$ megacycles, and $f_o = 3.8$ megacycles. Figure 23 shows the combined light intensities on the red tube for two scans of the same line, and Figure 24 has been constructed by adding Figure 23 to Figure 16, since the contribution from the red tube came entirely from the use of the mixed-highs.

c. Increased resolution in the blue channel

In the previous example, a method of eliminating cross talk between the red and green channels has been displayed, but the resolution in the blue channel to $f_B - f_0$ (0.3 megacycle in the numerical example) has been restricted. If it should prove desirable to follow the above path of exploration and it became evident that greater resolution were desired in the blue channel, the resolution could be doubled by a simple sampling or interrupting method with dot interlacing. By this method, the resolution could be increased to $2(f_B - f_o)$, or 0.6 megacycle in the example.

Suppose that a sampler with a very broad pulse but sampling at a rate of twice the frequency $f_B - f_o$ is incorporated in the blue channel and this sampler is followed by a low-pass filter which cuts off at one half the sampling frequency. Also a simple dot interlace is introduced. Let f_s be the sampling frequency. Then suppose the signal from the blue camera tube is $B + b \cdot \sin(\omega t)$. The function $1 + \cos(\omega_s t)$ will be used for sampling. When the frequency f is less than $f_s/2$, the signal out of the sampler and the low-pass filter is simply $B + b \cdot \sin(\omega t)$. This signal at the receiver is again sampled, this time by the function $\dfrac{1 \pm \cos(\omega_s t)}{2}$, giving

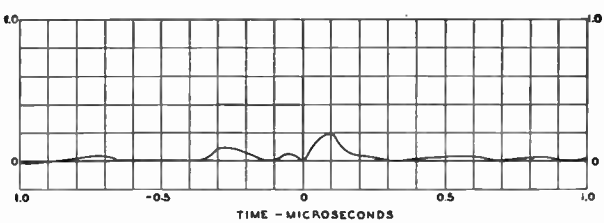


Fig. 23—Combined light intensities on the red tube.

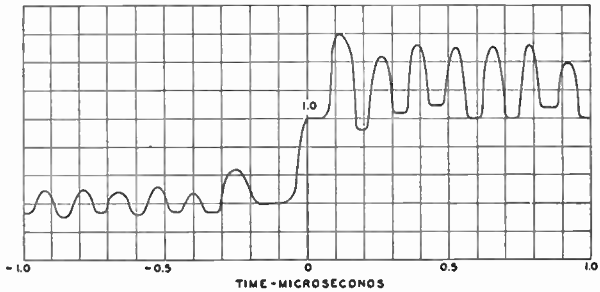


Fig. 24—Combined light intensities from red and green kinescopes.

$$[B + b \sin(\omega t)]\left[\frac{1 \pm \cos(\omega_s t)}{2}\right].$$

Figure 25 has been prepared, using

$$B = 1,$$
$$b = 1/2,$$
$$f = 0.1 \text{ megacycle},$$
$$f_s = 0.6 \text{ megacycle}.$$

Figure 25(a) shows the original function, while Figure 25(b) shows the effective light intensities for two scans. Figure 25(c) shows the sums of the light intensities for the two scans of Figure 25(b).

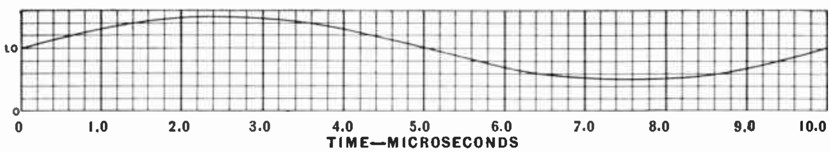


Fig. 25(a)—Signal out of the blue camera tube, $1 + 1/2 \sin(\omega t)$, with a frequency of 0.1 megacycle.

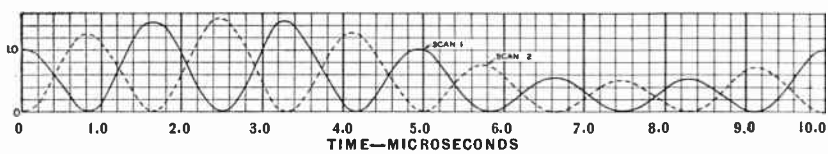


Fig 25(b)—Effective light intensities for two scans on the same line of the blue tube. Sampling frequency is 0.6 megacycle.

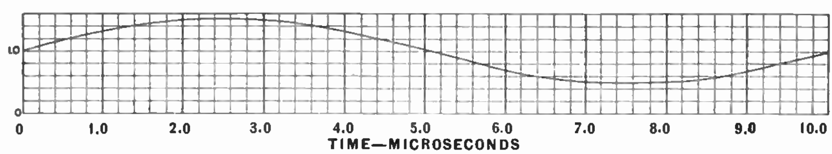


Fig. 25(c)—Sums of the light intensities for the two scans of Figure 25(b).

When the frequency f is greater than $f_s/2$, the response of the preceding circuits must be doubled. Hence the signal arriving at the sampler will be $B + 2b \sin(\omega t)$. After sampling at the transmitter by the function $1 \pm \cos(\omega_s t)$, the signal at the receiver second detector is $B \mp b \sin(\omega_s - \omega)t$. The second sampling at the receiver by the function $\dfrac{1 \pm \cos(\omega_s t)}{2}$ yields

$$\frac{B\ [1 \pm \cos(\omega_s t)]}{2} + \frac{b}{2} \sin(\omega t) \mp \frac{b}{2} \sin(\omega_s - \omega)t.$$

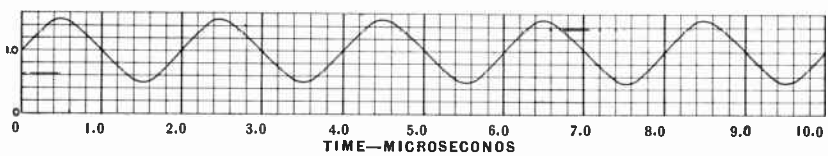


Fig. 26(a)—Signal out of the blue camera tube, $1 + 1/2 \sin(\omega t)$, with a frequency of 0.5 megacycle.

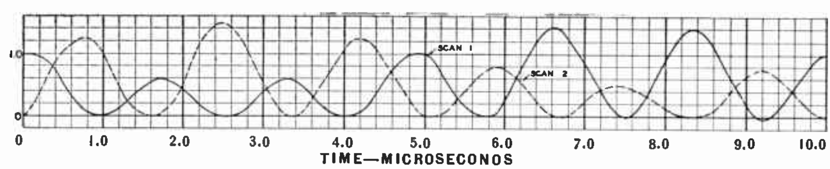


Fig. 26(b)—Effective light intensities for two scans on the same line of the blue tube. Sampling frequency is 0.6 megacycle.

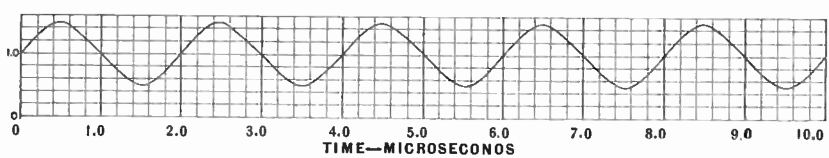


Fig. 26(c)—Sums of the light intensities for the two scans of Figure 26(b).

Figure 26 has been prepared, using

$$B = 1,$$
$$b = 1/2,$$
$$f = 0.5 \text{ megacycle},$$
$$f_s = 0.6 \text{ megacycle}.$$

Figure 26(a) shows the original function, while Figure 26(b) shows the effective light intensities for two scans. Figure 26(c) shows the sums of the light intensities for the two scans of Figure 26(b).

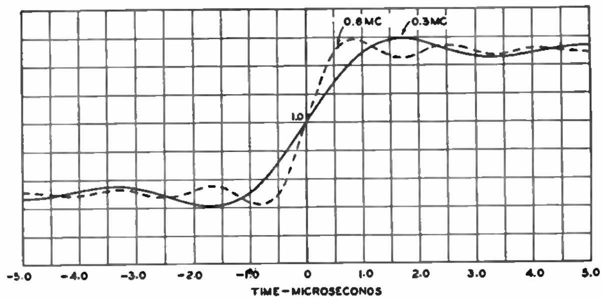


Fig. 27 — Change in step function for two frequency bands, one limited to 0.3 megacycle and the other limited to 0.6 megacycle.

This procedure illustrates the use of dot interlacing to obtain 0.6-megacycle resolution with a channel width of 0.3 megacycle.

The high-frequency sampling has been omitted from consideration in the above analysis. The signals from the blue channel are, of course, sampled at frequency f_o just as the red and green signals, but the filter at the receiver removes all traces of this sampling on the blue tube.

Figure 27 shows the change in a step function for two cases: first, where the frequency band is limited to 0.3 megacycle, and second, where the band is restricted to 0.6 megacycle. The increased steepness due to the wider band is apparent.

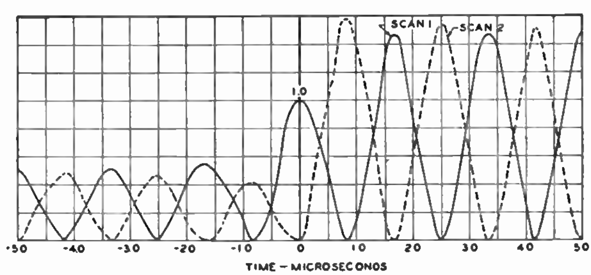
The step function response on the grid of the blue kinescope is given by

$$\text{B.K.S.} = \frac{[1 \pm \cos(\omega_s t)]}{2} \left[1 + \frac{2M}{\pi} \cdot \text{Si}(\omega_s t) \right]$$

$$\pm \frac{M}{\pi} \sin(\omega_s t) \left[\text{Ci}\left(\frac{\omega_s t}{2} \right) - \text{Ci}(\omega_s t) \right]. \qquad (36)$$

Figure 28 shows Equation (36) plotted for two line scans, where $M = 1/2$ and f_s is 0.6 megacycle. Figure 29 shows the addition of light intensities from Figure 28. It may be seen that Figure 29 is an exact reproduction of the dotted curve of Figure 27.

Fig. 28—Step function response on grid of blue kinescope tube with dot-interlacing and sampling.



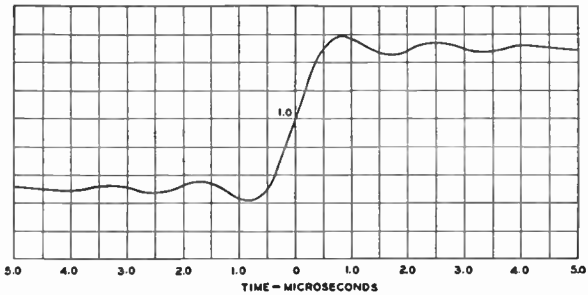


Fig. 29—Addition of light intensities on the blue tube, obtained by adding the curves of Figure 28.

CONCLUSION

Cross talk as a function of the width of the sampling pulse at both the transmitter and the receiver has been examined and limits established for reasonable cross talk. It is shown that narrow sampling at the receiver is more important than narrow sampling at the transmitter.

The sampling procedure was examined for large areas of color with a sinusoidal variation of the color. It was shown that the frequency passband was divided into three regions, the lower in which no cross talk existed, a middle region where fifty per cent cross talk was possible, and an upper region where signal mixing was expected because of the adoption of mixed highs. For the various cases, the role of dot interlacing was explained. In addition, the action on conventional black-and-white receivers as well as on color receivers was examined.

The sampling procedure was examined as it applied to step functions of light intensity. The response of a black-and-white receiver was examined and the desired and undesired responses of a color receiver were displayed.

A method of cross-talk elimination in the middle region is described. This method might be applied as an experiment in three parts. First, a simple cross-coupling and phase-shifting network is applied to the transmitter sampler. This circuit eliminates the cross talk between the red and the green channels in the middle region. No change is necessary at the receiver to take this first step. As a second improvement, a low-pass filter might be added in the blue channel at the receiver to knock out cross talk from the red and green into the blue channel. This step restricts the definition of the blue channel. A third step is suggested which doubles the resolution of the blue channel by a sampling and interlacing procedure.

The analysis and display of curves show that the sampling process in the dot-sequential color television system, together with the use of mixed highs provides a good uncoupling of the color channels together with full resolution equivalent to black-and-white transmission in the same channel.

The construction leading to Figure 10(g) emphasizes that the output of the sampler is the *product* of input signal and the gating function. This fact, together with the principle of mixed highs, produces full detail limited only by total bandwidth available. A study of Equation (31) shows that the rise time of the envelope is determined by the highest frequency passed in the mixed-highs circuit at the transmitter.

Throughout this report, the signal under consideration originated from a single primary color. If an area is a mixture of colors, the analysis may be carried out on the basis of the superposition of the individual responses to the three primary colors. Where, in the mixture of colors, the two stronger primaries are nearly equal in intensity, the variation due to the sampling frequency shown in Figures 7(c) and 8(c) virtually disappears, particularly on a standard black-and-white receiver.

During November, 1949, the sampling frequency of the dot-sequential color television system used experimentally in Washington was reduced from 3.8 to approximately 3.6 megacycles. Many of the calculations contained in this report were already completed at that time and were made on the basis of a sampling frequency of 3.8 megacycles. Rather than repeat the many laborious computations for the slight change in sampling frequency, the remainder of the calculations were continued at a sampling frequency of 3.8 megacycles. No very major change would have been apparent in the plotted results. The region free of cross talk in the simplest form of the system $(0 < f < f_B - f_0)$ would have been extended from 0.3 megacycle to 0.5 megacycle.