TELEVISION

Volume IV

(1942 - 1946)

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TELEVISION, Vol. IV

PREFACE

TELEVISION, Vol. IV, covering the period 1942-1946, is the fourth volume on **television** in the RCA REVIEW Technical Book Series. The first volume was published in 1936 followed by TELEVISION, Vol. II in 1937. TELEVISION, Vol. III, long-delayed by wartime security restrictions, appeared in early 1947.

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The large number of excellent papers on the subject of television has made necessary a very stringent selection process. All the available material can not be included in full form. A number of papers are, therefore, presented herein in summary form only; it has been necessary to omit others entirely. Suitably balanced presentation of the various phases of television was the major criterion in deciding which papers to include in full and which in summary. The presentation of a paper in summary form (or the non-inclusion of any particular paper) is not intended to indicate any deficiency in technical accuracy, literary merit, or importance.

* * *

The papers in this volume are presented in six sections: pickup, transmission, reception, color television, military television, and general. As a source of reference material, the Appendix to this book includes a television bibliography covering the period 1929-1946.

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RCA REVIEW gratefully acknowledges the courtesy of The Institute of Radio Engineers, the McGraw-Hill Book Company, and the Society of Motion Picture Engineers in granting to RCA REVIEW permission to republish material which has appeared in their publications.

The appreciation of RCA REVIEW is extended to all authors whose papers appear herein, and particularly to those whose papers are being printed in this book without prior publication.

* * *

TELEVISION, Vol. IV, like its predecessors, is being published for scientists, engineers and others interested in television, with the sincere hope that the material here assembled may help to speed developments and advance the position of television among companion arts and services.

RCA Laboratories
Princeton, N. J.
December 31, 1946

The Manager, RCA REVIEW

TELEVISION

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INTRODUCTION

BY

DAVID SARNOFF

President, Radio Corporation of America

SINCE the publication of earlier volumes on **television** in this RCA REVIEW Technical Book Series, the art has made notable advances. It has moved ahead from its status as a promising experiment to its present role as a proved medium of entertainment, education and news. Aided by research accelerated by the demands of war, the development of television has reached a point that otherwise might not have been expected at least until 1950. New tubes and new circuits have greatly improved transmitters and receivers and the growing interest currently being exhibited in television by the public is proof that television is being accepted rapidly as a new art and a promising new service.

For more than 20 years I have repeatedly expressed my confidence in the future of television. I have not hesitated to forecast that in due time it will become a greater industry and a greater art than sound broadcasting. And I believe that during this expansion of television, broadcasting will continue as a great industry, for the two forms of communication will supplement each other. There are natural reasons why sight and sound should be united to form a combination that will be far more effective than either medium alone, in imparting information, entertainment, culture and understanding of our life and our Government to all classes.

Television will not reach its full stature overnight. Well-founded scientific developments do not progress in that manner. History has proved that it takes about five years for any cycle in radio to translate itself into practical reality. The vacuum tube did not immediately supersede the spark in international communications nor did the superheterodyne become an overnight successor to the crystal receiver. After research has shown the way, a multitude of problems arise in the development of suitable programs and merchandising of the product that must be solved before the new service becomes universally acceptable. Television does not differ from other technical inventions in this respect.

An estimate at this time of the ultimate effect of television and the social and scientific consequences which will flow from its introduction would be impractical. But we do know from our experience that inventions which give us new powers over natural forces have had far-reaching effects on the human race. I need only mention Watts and his steam engine; Nobel and gunpowder; Morse's telegraph; Bell's telephone; Marconi's wireless and the Wright Brothers' aeroplane to support this view. These brain-children have produced and still are producing far-reaching effects on the family, on government, education, industrial production, yes, even on the habits and the beliefs of people. Moreover, as history shows, there is a cumulative gain as one development leads to another. This is exemplified in the case of the gas engine and jet propulsion, the single engined biplane and the stratoliner, and now in radio and television.

Because of these derivative results of inventions, the full social effects of a development such as television should be weighed on a general basis only. For instance, sociologists have called attention to the growing decentralization of industry. There are good reasons to believe that television may hasten this trend by simplifying the remote control of industrial operations, and by expanding the entertainment and cultural advantages which the video art will make available in small communities. As leisure time is increased along with greater technological efficiency, it is likely that television will be depended upon to make our hours of ease more profitable and satisfying.

Such a possibility creates the picture of a population which increasingly may center its interest more and more within the home. In such a setting, television will become a vital element in daily living. To people with more conveniences and with more time in which to enjoy them television may well become their principal source of entertainment, education and news.

Television, too, may enable man to keep pace with his thoughts. The human being has been created with a mind that can encompass the whole world in the fraction of a second; yet his physical senses lag woefully behind. With his feet he can walk only a limited distance; with his hands he can touch only what is within reach. With his unaided eyes he cannot see beyond the horizon, and his ears are useful only at short range.

These natural restrictions will vanish with the spread of television. In the years of its fulfilled destiny, television will bring the boundaries of the earth itself within the useful limits of Man's several senses. When this day arrives, there may come also a new philosophy and a finer, broader understanding between all peoples, whatever their nationalities or wherever they may live. If the airplane is named as the invention which annihilates distance, television can be said to annihilate both distance and time.

As this is written, television promises to make big strides in taking its place alongside the older arts, in many instances giving them new and modern import. Science has equipped the new industry for its expanded role. New supersensitive cameras, that make it possible to pick up scenes under low-lighting conditions; improved kinescope picture tubes that provide the viewer with steadier, brighter and clearer images on his receiver; mobile truck units which can speed to remote points and transmit pictures back to the main television transmitter for rebroadcasting; and new antennas that increase the area covered by television stations. To these advances, add the more skillful studio technique developed by broadcasters and a well-rounded, solid foundation for the new art is assured.

Leaders in education are becoming increasingly interested in television as an invaluable supplement to present school curricula. Through the medium of television, the skill and knowledge of the best teachers in the land can easily raise the educational level of the "little red schoolhouse" to the standing of leading schools in the larger cities.

In religion, also, the impact of television will be powerful and effective. Nationwide television will bring the services of the great cathedrals into the homes of the most remote residents. Viewers not only will hear the minister's words and the music but will see the preacher face to face as he delivers his sermon, and observe as though present, the solemn ceremonies at the altar.

Any discussion of television, however brief, should include some mention of its by-products. Some of the fields in which television devices may bring about important advances are in marine and aerial navigation, by permitting vision in fog and darkness through the use of infrared rays; in metallurgical, chemical, physical and biological research; in certain manufacturing processes as substitutes for human vision or for control purposes; in department stores for advertising and display; and in numerous other fields where a substitute for the eye may be useful.

The new flickerless, all-electronic color television system demonstrated at RCA Laboratories on October 30, 1946, and again on January 29, 1947, is the most recent positive contribution to the television leadership of our country. In the near future it is expected that outdoor scenes will be televised in color followed in 1948 by electronic color television on large-size theater screens.

The realization of this universal system of all-electronic color television, accomplished without the outmoded rotating disk or any other moving part, is as far reaching as was the creation of the RCA all-electronic television system which supplanted the mechanical disks used in black-and-white television when it first began.

With interest in television increasing more rapidly now than at any time since its introduction on a commercial basis in 1939, it is fitting that this Volume should be issued to continue the permanent history of research and achievement inaugurated by the preceding Volumes in the series. Reference to these records will provide those interested in this subject with useful information.

David Sarnoff

CONTEMPORARY PROBLEMS IN TELEVISION SOUND*†

BY

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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 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

Fig. 1—For good pictures, television cameras require most of the space. Sound equipment must operate in what remains.

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 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.

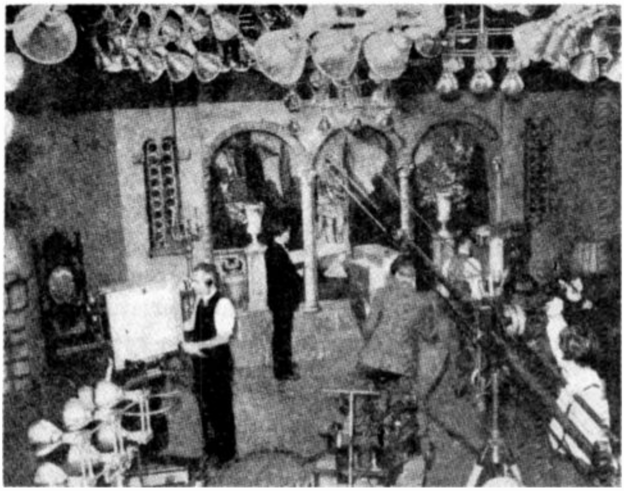

Fig. 2—Efficient utilization of floor space is a necessity in television.

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 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.

Fig. 5—Making close-ups and long shots simultaneously introduce problems of acoustic realism.

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.

THE FOCUSING VIEW-FINDER PROBLEM IN TELEVISION CAMERA*†

BY

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Summary—The technical excellence of a television program may frequently depend on the characteristics of the view finder used in the television camera. Conditions peculiar to television make it desirable that television-camera view finders be of the focusing type. The requirements of an ideal view finder of this type are discussed. During the past ten years a number of view-finder arrangements have been investigated in connection with the development of television cameras. Several of these are described and their relative merits indicated.

INTRODUCTION

ONE of the most essential elements in a television camera is the view finder. On its characteristics may depend the technical excellence of the television program. The desirability of minimizing operating personnel and the necessity for keeping a camera in practically continuous operation during television programs of one or two hours make it necessary that the view finder be of the focusing type. Such a view finder not only provides a view of the scene which is included in the field of the camera but also indicates when the lens is properly focused on the desired scene.

During the past ten years a number of focusing view finders were investigated to determine their suitability for use in television cameras. Brief mention of some of these arrangements has already been made in the technical literature on television equipment. Practical operating experience with several view finders both in the studio and outdoors has established certain requirements which an ideal view finder should meet. It is the purpose of this paper to discuss these requirements; to describe briefly several of the view-finder arrangements which have been investigated, and to indicate their relative merits.

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REQUIREMENTS OF AN IDEAL FOCUSING VIEW FINDER FOR TELEVISION CAMERAS

The requirements of an ideal view finder may be stated as follows:

1. It should at all times accurately indicate when the camera is in focus on the desired scene or object.

2. It should not only define that portion of the scene which is being converted into the television image but also should reproduce a sufficient portion of the scene outside the camera field so that the cameraman will know in advance what the television picture will include if he pans the camera in any direction.

3. It should provide an erect image which is correct left to right and of sufficient size and brightness to minimize eyestrain.

4. It should not unduly complicate the procedure of interchanging camera lenses or pickup tubes.

5. For portable pickup work the view finder should not contribute substantially to the size and weight of the camera.

It will be noted that the first three of these requirements deal with performance whereas the last two are concerned primarily with operating convenience.

In order to appreciate the significance of these requirements it is of interest to discuss them in connection with the two general groups of view finders into which the several individual view finders are subsequently classified. For the purpose of this discussion the first group will consist of those view finders which derive the view-finder image either directly or indirectly from the camera lens. The second group includes those which make use of a separate optical system for producing the view-finder image.

REQUIREMENT NUMBER 1

Requirement 1 specifies that the view finder should at all times accurately indicate when the camera is "in focus" on the desired scene or object. Practical operating experience has shown that in respect to this requirement it is desirable that the cameraman be aware of a degradation in picture detail due to improper focus before the loss in resolution is apparent to the television audience. The view finders in group 1 have several limitations with respect to this requirement. When the scene which is being televised is sufficiently illuminated so that the camera lens can be stopped down to provide greater depth of focus the view finders in this group do not provide an accurate focus indication since the view-finder image has the same depth of focus as the camera image. In other words, no apparent change in detail is observed by the cameraman as the lens is moved back and forth through an appreciable range. This limitation may not be particularly apparent to the television audience from the standpoint of picture detail but is likely to be disturbing for another reason. Under this condition the cameraman has a tendency to move the camera lens back and forth to determine by approximation the center of the range over which no effect on picture detail is observed and thus establish the "in-focus" position of the lens. As the lens is moved back and forth, the area included in the television image changes in such a manner that the sides of the picture appear to move in and out; an effect which is disturbing to most observers.

Another result of this inaccurate focus indication is encountered when the camera is used under conditions where the illumination may vary suddenly through a fairly wide range. Such conditions are frequently encountered in outdoor pickup of sporting events or spot news, etc. If the lens is stopped down and the camera is inaccurately focused on a scene in bright sunlight and the sun subsequently goes behind a cloud, making it necessary to increase the lens aperture, the camera will be out of focus. The focusing readjustment which is then required would have been avoided if the view finder had met requirement 1.

The view finders in group number 2 can all be made to meet requirement 1 provided they are constructed with sufficient mechanical rigidity to maintain, at all times, the proper alignment between the optical systems for the view finder and pickup tube.

REQUIREMENT NUMBER 2

Requirement 2 states that the view finder should always provide an image of and accurately define that portion of the scene which is being converted into the television picture and should also provide a view of at least a small part of the scene on each side of the television-picture area. Unless the first part of this requirement is fulfilled the cameraman may not know, for example, whether or not an individual's head is in the picture. The second half of this requirement gives the cameraman an indication of what will be included in the picture if he pans the camera in any direction. The need for this information may depend to some extent on whether the camera is being used in the studio or outdoors. From one standpoint, there is less need for this additional view-finder-image area in the studio because studio programs are usually rehearsed several times. On the other hand, in studios several sets are frequently used in a limited space so that a camera can be changed quickly from one scene to another. This makes it necessary for the cameraman to know what is included in a small area outside the field of his camera so that he does not inadvertently include an edge of an undesired set in the picture. If the view finder does not provide an image of this additional area it is frequently necessary for the cameraman to move his head sufficiently so that he can look along one side of the camera to determine the effect of panning the camera in a desired direction. Not only is this inconvenient but when the cameraman looks around the camera at the brighter scene and then again looks at the image in the view finder it is necessary for his eyes to readjust themselves to the difference in the light intensity. In outdoor pickup work such as sporting events, where the action is unpredictable, if the cameraman looks around one side of the camera he may lose the action altogether before he has time to again look into the view finder.

In general, the view finders in group 1 do not meet requirement 2 since the view-finder image which they provide is obtained from the camera lens and covers the same picture area as the television image.

The view finders in group 2 make use of a separate optical system and, therefore, can be made to provide a view of some of the scene around the area which is converted into the television image. Such view finders are, of course, provided with hairlines on the viewing screen or some other expedient which indicates the actual area of the scene which is included in the television picture. It is essential that the view finders in this group be provided with some means which will correct for parallax between the two optical systems.

REQUIREMENT NUMBER 3

The ideal view-finder requirement 3 is met if the view finder provides an erect image which is correct left to right and of sufficient size and brightness to minimize eyestrain. A difference of opinion may exist as to the necessity of having the view-finder image erect and correct left to right. If the cameraman has received considerable training with cameras providing images which are inverted and reversed left to right, such a view finder is undoubtedly satisfactory. He will then have developed the proper co-ordination between the image he sees in the view finder and the direction in which he must move the camera to produce a desired effect. On the other hand, in a new field, such as television, where it will be necessary to start with relatively untrained personnel, it is felt that the corrected view-finder image will be more satisfactory.

With respect to the other stipulations in requirement 3, a view-finder image at least 3 by 4 inches at a viewing distance of 12 inches has been considered to be satisfactory. The image should be as bright as possible. No difficulty has yet been encountered from having the view-finder image too bright. The ability of a specific view-finder arrangement to meet requirement 3 is basically determined by the amount of light which it supplies to produce the optical image since, if sufficient light is available, an optical system can be used to increase the image size and reverse it in either or both directions.

The problem of providing sufficient light to produce a satisfactory view-finder image is becoming more difficult as the sensitivity of camera pickup tubes is increased. This limitation may ultimately make it necessary to resort to a highly complicated view-finder arrangement which will be described later.

REQUIREMENT NUMBER 4

This requirement is concerned with the effect of the view finder on the ease of interchanging either pickup tubes or lenses. Since emergencies may arise which make it necessary to change pickup tubes and since it is frequently desirable to change to a different focal-length camera lens, it is essential that these changes be made in the shortest time and with the least inconvenience.

This requirement is met to the greatest extent by the view finders in group 1 since they derive the view-finder image from the camera lens. The view finders in group 2, which use a separate optical system for producing the view-finder image, all contain some element which must be adjusted to provide satisfactory optical alignment between the two optical systems when pickup tubes are changed. Up to the present time it has been impracticable to manufacture pickup tubes with sufficiently close tolerance on the position of the mosaics and other elements of the tubes to make them optically interchangeable. Some adjustment, therefore, is necessary so that the view-finder image and the image on the pickup tube are "in focus" simultaneously. It is possible to shift the position of the pickup tube in a camera to obtain satisfactory optical alignment between the two optical systems. The size and weight of the pickup tube with its deflecting yoke, however, make it much more practical to move a ground-glass screen or some other element in the view finder to provide the necessary alignment between the two optical systems.

No serious complications are encountered in interchanging lenses of different focal lengths in cameras employing the view finders in group 1 since only the camera lens is changed.

The additional lens required with the dual-lens view-finder arrangements in group 2 make the problem of interchanging lenses somewhat more difficult. This is particularly true where the lenses are large and heavy such as those having focal lengths of 20 inches or more and apertures of the order of *f*/4.5.

REQUIREMENT NUMBER 5

Requirement 5 is based on the desirability of keeping the size and weight of television cameras for portable pickup work at a minimum. Studio cameras are usually semipermanently mounted on large dollies similar to those used in motion-picture work and the size and weight of the television camera for studio work is, therefore, not a primary consideration. Portable television cameras, however, are used on conventional tripods and are set up and subsequently taken down at each pickup location. It is, therefore, desirable to keep the size and weight of cameras for portable pickup work at a minimum. In some cases a sacrifice in view-finder performance has been made to permit a reduction in the size and weight of the camera. Some portable cameras which employ one of the more complicated view finders are constructed so that the camera can be separated into two units. This construction not only makes the camera more portable but also makes it possible to mount the two parts separately on the tripod.

Since the view finders in group 1 require less parts, occupy less space, and contribute less weight, they are more acceptable from the standpoint of requirement 5 than those in group 2.

DESCRIPTION OF INDIVIDUAL VIEW FINDERS

The following is a list of the view finders which will be described.

1. Mirror arrangement for observing the optical image on the mosaic of the pickup tube.
2. Semisilvered mirror arrangement for utilizing the camera lens to produce an optical image on a ground-glass viewing screen.
3. Kinescope or electronic view finder.
4. Kinescope or electronic view finder with remote focusing control.
5. Split-image view finder as used in the Contax and similar cameras.
6. Duplicate-lens view finder as used in the Rolliflex camera.
7. Combination duplicate lens and kinescope view-finder.

The first four view finders in this list derive the view-finder image either directly or indirectly from the camera lens and are those which were previously classified as the group 1 view finders. View finders 5, 6, and 7 are the group 2 view finders and obtain the view-finder image from a separate optical system. For the sake of simplicity, the diagrams which will be used to illustrate the several view finders will not show any means either for magnifying the optical image or for correcting it in the vertical and horizontal directions. It is apparent, that if sufficient light is available, lens and mirror arrangements can be used to accomplish any of these results. The means for correcting for parallax is likewise omitted from the diagrams of the group 2 view finders. Although the iconoscope is shown as the pickup tube in each of the diagrams it is obvious that the orthicon or any other type of pickup tube may be used.

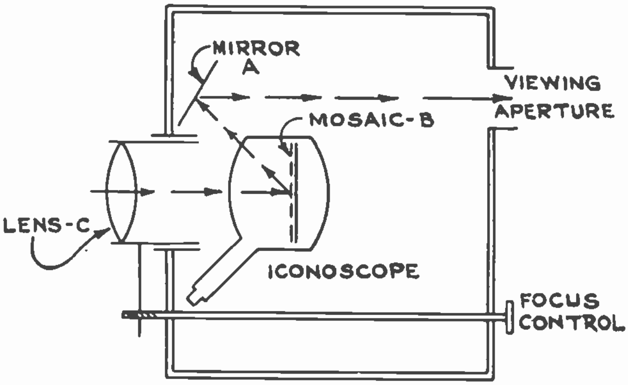

Fig. 1—Mirror arrangement for viewing the optical image on the mosaic of the pickup tube.

MIRROR ARRANGEMENT FOR VIEWING THE OPTICAL IMAGE ON THE MOSAIC OF THE PICKUP TUBE

The original iconoscope camera view-finder arrangement is illustrated by the diagram in Figure 1. With this view finder the cameraman, through the use of mirror *A*, observes on the mosaic *B* the optical image which is produced by the camera lens *C*. The shape of the glass envelope of the pickup tube is usually such that only a portion of the image on the mosaic can be observed through the use of this system. The mosaics of the more recent pickup tubes have very poor light-reflecting properties and the optical image produced on the mosaic is, therefore, unsatisfactory from the brightness standpoint. The chief advantages of this arrangement is its simplicity. It requires a minimum of equipment since it makes use of only the camera lens and does not employ a separate viewing screen. No special adjustments are necessary when changing either the camera lens or the pickup tube. It has, however, all the limitations previously mentioned in connection with the group 1 view finders.

SEMISILVERED MIRROR ARRANGEMENT UTILIZING THE CAMERA LENS TO PRODUCE AN OPTICAL IMAGE ON A GROUND-GLASS VIEWING SCREEN

The diagram in Figure 2 illustrates the view finder system, which makes use of a semisilvered mirror *A* to reflect some of the light transmitted by the lens *B*. This light is again reflected by the mirror *C* to produce an optical image on the ground-glass viewing screen *D*. In the experimental work on this arrangement, mirrors were used in which the reflected light varied from 15 to 40 per cent. Since the total light reflected from the front-surfaced mirror *A* is a comparatively small percentage of the light passing through the mirror, the light reflected from the back surface of the mirror may be a fairly large percentage of the total reflected light. It is, therefore, necessary to use either a very thin mirror or else have the back surface of the mirror coated with a nonreflecting film; otherwise the light reflected from the back surface produces an image which is sufficiently displaced from the front-surface image to reduce the effective resolution of the view finder to a point where it is definitely unsatisfactory.

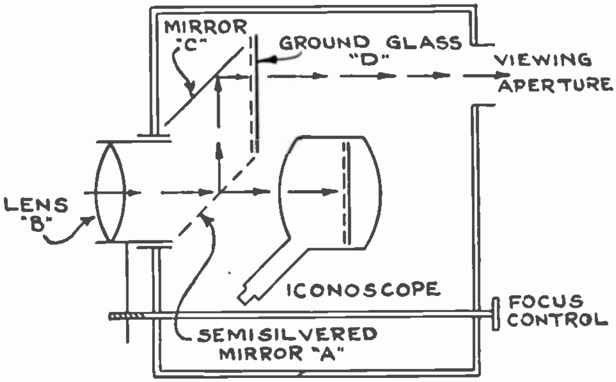

Fig. 2—Semisilvered-mirror view-finder arrangement.

The chief advantage to be found in this view finder likewise lies in its relative simplicity. With respect to the arrangement shown in Figure 1, it has the advantages of giving a somewhat brighter image and also will provide a view of the scene whose area is greater than that included in the field of the camera.

The most serious disadvantage of this view-finder arrangement is that it robs light from the mosaic of the pickup tube and therefore decreases the effective light sensitivity of the system. Although it meets requirement 2 it has the other limitations of the group 1 view finders. Since a separate ground-glass viewing screen is used with this arrangement it is necessary to adjust the position of the viewing screen when changing pickup tubes so that the viewing screen is the same distance from the optical center of the lens as the mosaic of the pickup tube. This view-finder arrangement also imposes a limitation on the shortness of the focal length of the camera lens which can be used.

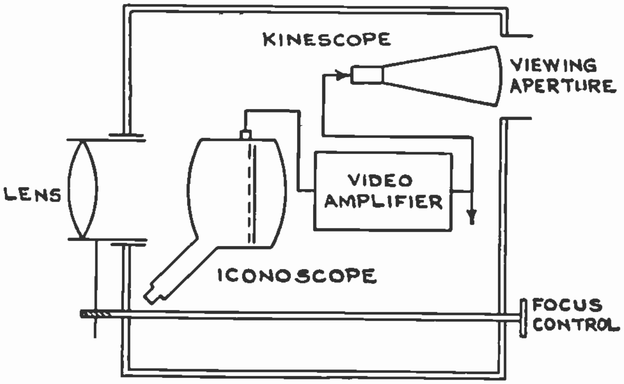

Fig. 3—Kinescope or electronic view finder.

KINESCOPE OR ELECTRONIC VIEW FINDER

This view-finder arrangement is obtained by incorporating in the camera a kinescope on which is reproduced the television image. It is illustrated by the diagram in Figure 3.

The chief advantage of this view-finder system is that the relative brightness of the view-finder image does not diminish as the sensitivity of the pickup tube is increased. The brightness of the Kinescope view-finder image is determined primarily by the characteristics of the kinescope which is used and the operating voltages which are employed. It, like the view-finder arrangements illustrated in Figures 1 and 2, does not necessitate any view-finder adjustments when either pickup tubes or camera lens are interchanged and no correction for parallax is required.

In addition to the several limitations discussed in connection with the group 1 view finders the kinescope type of view finder has the further restriction that the sharpness of the view-finder image is dependent on the resolution of that portion of the television system which it includes. It is, therefore, necessary that satisfactory electrical focus of the kinescope be maintained for this view finder to function satisfactorily. The space required in a camera to house this type of view finder is relatively large. The several thousand volts which are used as anode supply for the kinescope present a problem in providing a satisfactory camera cable. If this camera-cable problem is avoided by incorporating a voltage-supply unit in the camera a corresponding increase in the size and weight of the camera results.

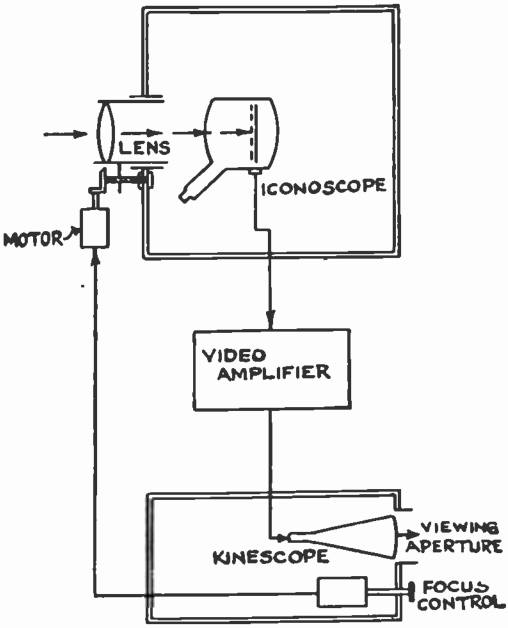

Fig. 4—Kinescope or electronic view finder with remote focusing control.

KINESCOPE OR ELECTRONIC VIEW FINDER WITH REMOTE-FOCUSING CONTROL

In the kinescope view-finder arrangement just described, a television monitoring unit with its kinescope is in effect moved from its normal location so that it can be associated directly with the focusing control in the camera. In the remote-control form of the kinescope view finder the physical location of the parts is reversed and a remote camera-focusing control is provided that can be used at the normal location of the television monitoring unit. The diagram in Figure 4 illustrates this arrangement. As indicated in the diagram, the remote control of focus is accomplished through the use of Selsyn motors.

The chief advantage of this view-finder system lies in the fact that it permits a camera which is small in size and light in weight. This is especially desirable in portable pickup work. It makes possible a camera which is particularly suitable for locations which are inaccessible to a cameraman. It also provides the advantages which have been discussed in connection with the previous kinescope view finder. With the remote-focusing arrangement the only view-finder equipment which must be housed in the camera is the small Selsyn motor. A wire-frame view finder mounted on the side of the camera is used by the cameraman to keep the camera trained on the desired scene. The focusing is done by a control operator at the monitoring unit.

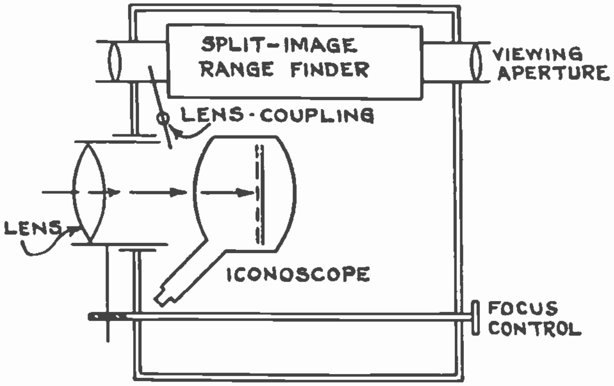

Fig. 5—Split-image view finder.

In addition to the deficiencies of the kinescope view finder illustrated by Figure 3, this arrangement has the further limitation that a fairly high degree of co-ordination is required between the man who is panning the camera and the man at the remote point who is operating the focusing control. When the focusing and panning are done by the same individual he subconsciously starts to adjust the focusing control in the proper direction to correct for any change in distance between the camera and the desired scene.

SPLIT-IMAGE VIEW FINDER AS USED IN THE CONTAX AND SIMILAR CAMERAS

The diagram in Figure 5 illustrates this type of view finder. It utilizes an optical system which is actuated by the focusing control simultaneously with the camera lens and produces two optical images which are accurately superimposed when the camera lens is in focus on a desired object or scene. The two images are displaced with respect to each other when the focusing control has not been properly adjusted. In a view finder of this type which was investigated the condition of focus could be accurately determined only in a small area in the center of the picture. Another limitation of this particular view finder was that when using long focal-length lenses the actual size of an object in the view finder remained the same as when a short focal-length lens was used. A hairline indicator was provided to indicate the smaller field covered by the longer focal-length lens. An adjustment is required with this type of view finder when interchanging pickup tubes so that the optical system of the view finder is adjusted to compensate for variations in the position of the mosaic in different pickup tubes.

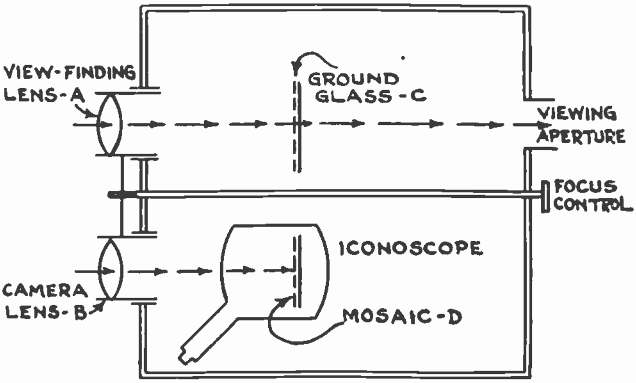

Fig. 6—Duplicate-lens view finder.

DUPLICATE-LENS VIEW FINDER AS USED IN THE ROLLIFLEX CAMERA

As shown in Figure 6, an auxiliary lens *A*, which has the same focal length as the camera lens *B*, is used to produce on the ground glass *C* an optical image which corresponds to the optical image on the mosaic *D* of the pickup tube. The position of the ground glass *C*, with respect to the optical center of the lens *A*, must always correspond to the position of the mosaic *D* with respect to the lens *B*. The two lenses must be matched accurately for focal length. To facilitate interchanging lenses of different focal lengths each pair of lenses are usually assembled on a single mounting plate. This view-finder system provides an image of a portion of the area outside that covered by the field of the camera. A view finder of this type provides a very accurate indication of focus under all conditions since the view-finder lens can be kept wide open when the camera lens is stopped down. Since a fast lens is normally used to provide the view-finder image the brightness of this image has been relatively satisfactory. The increased sensitivity of pickup tubes, however, is causing the image brightness obtained from this view-finder system to decrease to the point where it no longer will be satisfactory. Some system for parallax correction is required with this type of view finder. The amount of correction which is necessary is generally determined by the maximum diameter of the lenses supplied with the camera.

The inability of this view finder to meet the ideal view-finder requirements is found in connection with requirements 4 and 5. Since a separate lens is used to produce an optical image on a ground-glass screen, the position of this screen must be adjusted to correspond to that of the pickup-tube mosaic whenever pickup tubes are interchanged. Since the longer focal-length lenses (20-inch, *f*/4.5 lenses are frequently used) are large and heavy, the additional lens required for this view finder not only makes the problem of interchanging lenses more difficult but materially increases the over-all size and weight of the camera.

COMBINATION DUPLICATE-LENS AND KINESCOPE VIEW FINDER

It has previously been pointed out that as the sensitivity of the television pickup tube is increased a corresponding decrease occurs in the relative brightness of the image in an optical view finder. At present, when the maximum sensitivity of the orthicon pickup tube is utilized, the image brightness obtained from an optical view finder, such as the duplicate-lens arrangement previously described, is on the verge of being unsatisfactory. With the kinescope type of view finder any increase in the sensitivity of the pickup tube is automatically compensated insofar as the brightness of the view-finder image is concerned. The types of kinescope view finders which have been described, however, do not meet performance requirements 1 and 2. If a further improvement is made in the sensitivity of television pickup tubes, it may be necessary to use a view finder of the type illustrated in Figure 7.

In this diagram it will be noted that two pickup tubes are used with a pair of duplicate lenses. Associated with the camera pickup tube are the normal television amplifier and deflection circuits. The amplifier used with the view-finder pickup tube is designed to pass a wider frequency band than is normally required by the television system. The increase in resolution, which the wider frequency band permits, enables this view finder to provide a more accurate indication of focus than could be obtained from the previous kinescope view finders. Since a separate view-finder lens is employed it can always be used at its maximum aperture even though the camera lens is stopped down, and thus provide at all times an accurate indication of the proper focus adjustment. The deflection circuits for the view-finder pickup tube are so arranged that a slightly greater area of the scene is scanned than is the case with the camera pickup tube.

The deficiencies of this view finder from the standpoint of the ideal view finder are in respect to the requirements 4 and 5 which deal primarily with operating convenience. With reference to requirement 4, when pickup tubes are interchanged, the position of one of the pickup tubes must be adjusted so that the mosaics of the two tubes are the same distance from their respective lenses. The electrical focus of both the view-finder pickup tube and kinescope must be kept in proper adjustment for this view finder to function satisfactorily. The extra equipment required for this type of view finder materially increases the size and weight of a television camera.

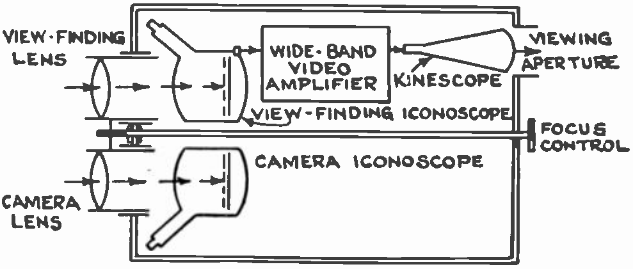

Fig. 7—Combination duplicate-lens and kinescope view finder.

COMPARISON OF THE INDIVIDUAL VIEW FINDERS

Table I shows the ideal view-finder requirements that are met by the several view finders which have been described. The wording used in the table for each of the requirements is such that a "yes" in the column beneath a given view finder indicates that it meets the specified requirements.

CONCLUSIONS

It is apparent that none of the view finders which have been de-

TABLE I

View Finders	No. 1 Mirror Arrangement for Observing the Optical Image on the Mosaic of the Pickup Tube	No. 2 Semisilvered Mirror Arrangement Utilizing Camera Lens to Produce Image on Viewing Screen	No. 3 Kinescope or Electronic View Finder	No. 4 Kinescope View Finder with Remote-Focus Control	No. 5 Split-Image View Finder	No. 6 Duplicate-Lens View Finder	No. 7 Combination Duplicate-Lens and Kinescope View Finder
Ideal View-Finder Requirements							
No. 1 Provides accurate indication of focus under all conditions.	No	No	No	No	Questionable	Yes	Yes
No. 2 Accurately defines scene in television picture and gives view of some additional area	No	Yes	No	No	Yes	Yes	Yes
No. 3 Provides image of satisfactory size and brightness	No	No	Yes	Yes	Questionable	Yes at Present	Yes
No. 4 No special adjustments required when pickup tubes or lenses are interchanged	Yes	No	Yes	Yes	No	No	No
No. 5 Does not require a serious increase in the size and weight of the camera	Yes	Yes	No	Yes	Yes	No	No

scribed meet all the requirements of an ideal view finder. The relative importance of some of the requirements is determined to a considerable extent by whether the camera is intended for studio or outdoor pickup work. In general, the duplicate-lens type of view finder has given the most satisfactory results. If it is desired to keep the size and weight of the camera as near the minimum as possible, the kinescope view finder with remote-focusing control is a practical arrangement. A substantial increase in the sensitivity of television pickup tubes will result in more consideration being given to the several types of kinescope view finders.

In this discussion no reference has been made to the relative cost of the various view-finder arrangements. For the time being, at least, the cost of television pickup equipment has been considered to be of secondary importance to performance and operating convenience.

ACKNOWLEDGMENT

The writer wishes to acknowledge the individual and cooperative efforts of numerous Radio Corporation of America and National Broadcasting Company engineers who have contributed to the solution of the view-finder problem.

ELECTRON BOMBARDMENT IN TELEVISION TUBES*†

BY

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Summary—A detailed analysis of actions occurring in an Iconoscope when an elemental area of the mosaic is bombarded by the scanning electron beam under conditions varying from dark to light. The sticking effect, important in projection kinescopes, is explained.

IN AN ordinary vacuum tube there are two main effects—the controlled unidirectional flow of electrons from cathode to plate, and the bombardment of the plate by these electrons. The controlled unidirectional conduction is in these tubes the desired effect, and the electron bombardment is generally undesirable because it heats the collecting electrode and results in energy losses.

While modern all-electronic television makes use of a great number of ordinary radio vacuum tubes, its actual functioning depends mainly on cathode-ray tubes at the transmitting and receiving ends of the television system. In television cathode-ray tubes the same two main effects exist, but with an important difference—the electron bombardment is utilized, while the unidirectional conduction is incidental to the operation of the tubes. Electron bombardment of targets and the resulting secondary emission make possible the operation of both the Kinescope (receiving tube) and the Iconoscope (camera pickup tube).

There is another distinction between the ordinary vacuum tube and the television cathode-ray tube. In ordinary tubes the plate acts as both the target and the collector, whereas in television tubes the target is either an insulator or an insulated conductor, with the collection of electrons being done by another electrode. This collecting electrode is usually the second anode.

When a surface is bombarded by electrons of considerable velocity, the incident electrons may impart to the electrons near the surface sufficient energy to escape from the surface. The incident electrons are called the primary electrons, while the electrons leaving the sur-

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* Decimal Classification: R583.6 x R138.3

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face are called the secondary electrons. The number of secondary electrons emitted for each primary electron, and the velocity of the secondaries, vary with the velocity of the primary electrons and with the chemical nature and physical condition of the surface.

BOMBARDMENT RESEARCH TUBE

In Figure 1 a typical tube for studying electron bombardment of a given surface is shown. The target is located at the center of a metallic sphere and is kept at a desired potential with respect to the cathode, thereby assuring a definite velocity of the bombarding primary electrons. The electron beam is produced in a conventional electron gun and enters the sphere through a small hole in its wall. Provisions are made for varying the potential of the sphere with respect to the target, as well as for reading the currents to the target, the sphere and the beam current.

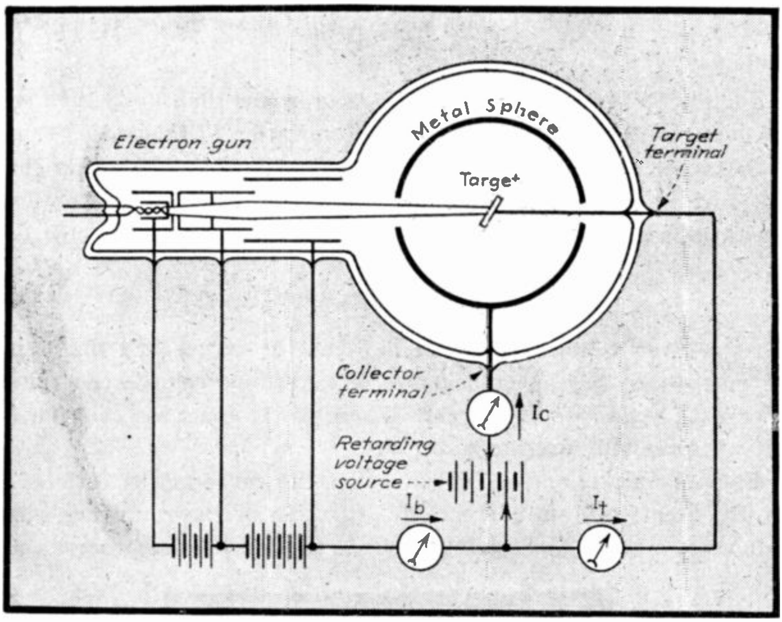


Fig. 1—Essential features of special cathode-ray tube developed for studying electron bombardment of target surfaces.

For a conducting target, such as pure nickel, the target current I_t is equal to the beam (primary) current I_b minus the collector current I_c ($I_t = I_b - I_c$). A negative value for target current I_t indicates that the ratio of secondary current to primary current (I_s/I_p) is greater

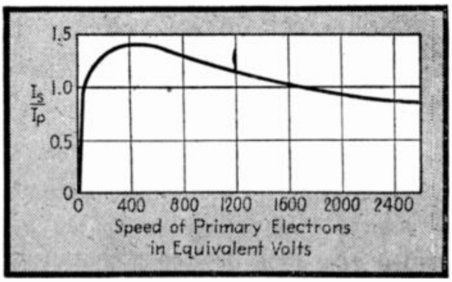


Fig. 2—Ratio of secondary electrons to primary electrons as plotted against speed of primary electrons when using a pure nickel target in the tube of Fig. 1.

than unity. When I_t is positive, the ratio is less than unity, and when I_t is zero the ratio is unity. A curve of variation of this ratio for pure nickel target as a function of velocity of primary electrons is shown in Figure 2. (For a contaminated metal surface the secondary emission ratio is generally greater than for the clean surface shown.)

EQUILIBRIUM POTENTIAL OF TARGET

At a given voltage difference between the target and the cathode one may apply increasing negative or retarding voltages on the collector with respect to the target. When this is done, the collector current will gradually drop to zero.

For pure nickel and a 500-volt beam, the curve of the ratio of collector current to beam current as a function of the retarding voltage is shown in Figure 3. At point P, where the curve goes through a ratio

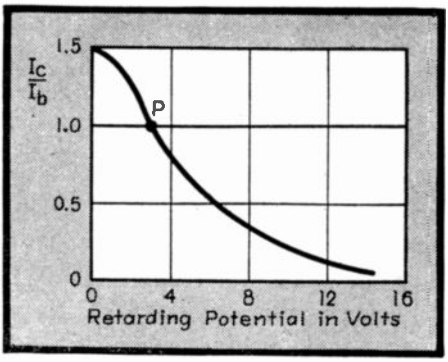


Fig. 3—Ratio of collector current to beam current as plotted against negative retarding potential (collector negative with respect to target) when using a pure nickel target and a fixed primary electron speed of 500 equivalent volts in the cathode-ray tube of Fig. 1.

value of unity, the target current I_t is equal to zero. This is an important point on the curve. Since there is no current flowing to the target, the lead to the target may be cut under the conditions at P without disturbing either the electrode potentials or the currents to them.

Since the potential of the target is not changed when the target lead is cut under the conditions at P, the target potential is still equivalent to the beam velocity. The collecting sphere, however, is at —3 volts with respect to the target. A conclusion follows: an *insulated* nickel target will assume a potential of 3 volts higher than the collector for 500-volt primary electrons. (The collector has to be at 497 volts.) Or, generally speaking, when the secondary emission ratio is higher than unity, an insulated metal target will assume a potential of a few volts positive with respect to the collector. This potential is called the equilibrium potential of an insulated target under electron bombardment. The velocity of the primary electrons will of course be equivalent to the sum of the collector voltage plus the voltage between the insulated target and the collector.

STICKING EFFECT

Now let the nickel target float and increase the collector potential to 1700 volts. The secondary emission ratio of pure nickel at this voltage is unity and the target will float to the same potential as the collector, so that the difference of potential between the two is zero.

If the collector potential is further increased, the insulated target will stay at the same potential with respect to the cathode. In other words, it will be getting more and more negative with respect to the collector. This phenomena is called the "sticking effect" in television vernacular. To observe how it happens, make a metallic connection between the target and the collector and raise both to 2000 volts with respect to the cathode. The secondary emission ratio at 2000 volts is nine-tenths, and while meter I_b will measure the current flowing to the target and collector together, nine-tenths of it will flow to the collector and one-tenth to the target.

If the target is now disconnected and left floating, more electrons will be arriving at it than departing from it, charging it negatively with respect to the collector. At the same time the arriving primary electrons slow down to the velocity equivalent to the actual voltage between the target and the cathode.

With primary electrons slower than 2000 volts, the secondary emission ratio increases, and finally, when the target is at exactly 1700 volts positive with respect to the cathode, and 300 volts negative with respect to the collector, the ratio becomes unity. In other words, the target here "sticks" at an equilibrium potential of 1700 volts, and any increase in the voltage on the collector will not increase the potential of the floating target.

The two cases just described play a very important part in the performance of television cathode-ray tubes and have to be clearly understood before an analysis of their performance can be undertaken. Sticking is especially important in projection kinescopes when it is desired to use high voltages on the second anode to get more light. Before using these high voltages, such as 20 to 70 thousand volts, one must make certain that the luminescent material does not "stick" below the value chosen.

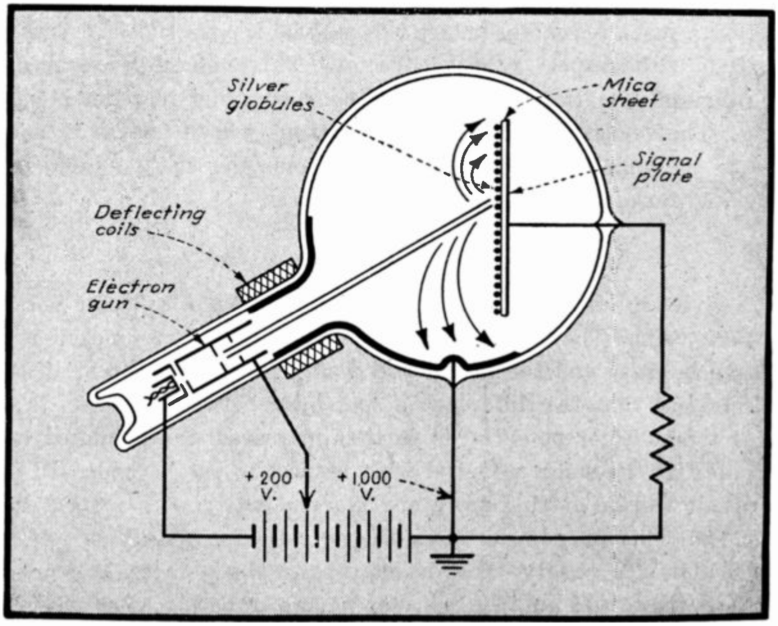


Fig. 4—Essential features of a standard Iconoscope for electronic television cameras.

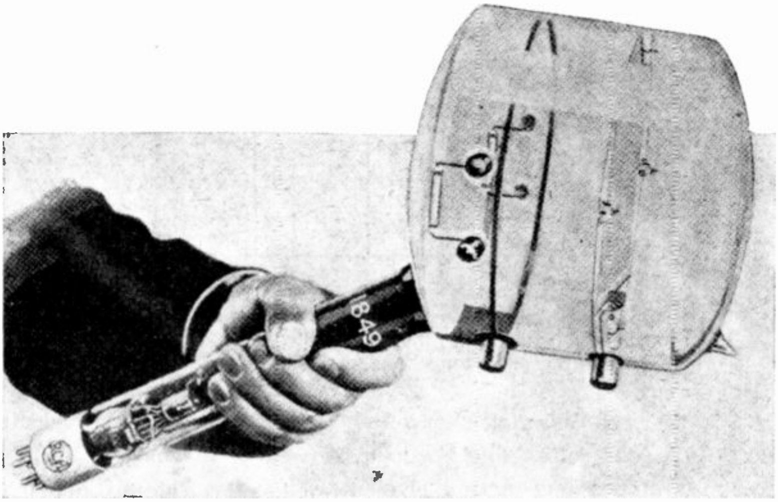
MOSAIC IS TARGET IN ICONOSCOPE

The action of electron bombardment in a standard Iconoscope is somewhat similar to the case of bombarding an insulated metal target with electrons having a velocity at which the secondary emission ratio of the target is greater than unity.

Figure 4 shows the arrangement of the essential parts in a standard Iconoscope. Generated by a conventional electron gun, an electron beam of approximately 1000-volt velocity enters a nearly equipotential space in the bulb portion of the tube, where it strikes the photosensitive mosaic. The mosaic consists of a multiplicity of minute silver globules, oxidized and caesiated (in other words, photo-sensitized), uniformly distributed on a 9 × 12-cm sheet of mica only 0.0025 cm thick. The back side of the mica sheet is platinized to form a capacitor between the globules and the platinum coating, having a value of 122 $\mu\mu$f per sq cm.

The secondary emission characteristics of oxidized and caesiated silver vary greatly, depending on the condition and method of preparation of the surface. These variations however are restricted to the maximum value of the secondary emission, while the relative velocity or energy distribution of secondaries changes little.

The Iconoscope—a television pickup tube.



A typical energy distribution curve of secondary emission of photosensitized silver for 1000-volt primaries is shown in Figure 5. As may be seen from the curve, the secondary emission characteristics of a complex surface differ from those of pure metals. In the case under consideration two important characteristic features attract attention at once. The first is the fact that the secondary emission ratio reaches a very high value of 5.1, compared with slightly more than one for pure nickel.

The second feature is that, with the collector at zero potential with respect to the target, not all the electrons from the target are collected. Apparently either there are some electrons hidden in the crevices of the surface, or some electrons are emitted with "insufficient" velocities to reach the collector. They may be drawn to the collector by applying positive potentials to the collector with respect to the target. The second feature, while interesting, is of little importance to us since the collector is seldom positive with respect to the mosaic. Our interest lies, therefore, in the portion of the curve to the right of the line of zero collector potential.

SCANNING ACTION

In actual Iconoscopes, the primary beam may be used with electron velocity between 500 and 2000 volts. Usually however, the velocity is 1000 equivalent volts. The beam is focused at the mosaic to an area of approximately one picture element. Therefore, at any one instant the area under electron bombardment is that of one picture element. The scanning spot, however, moves along the mosaic at a very high speed, bombarding a point on the mosaic for only 1.28×10^{-7} sec. for every frame of a 441-line picture.

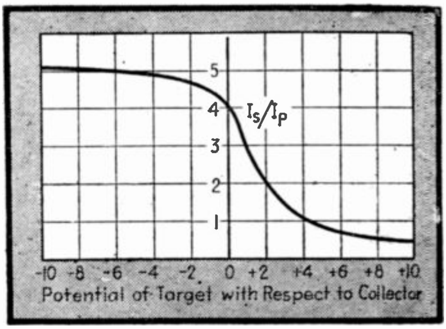


Fig. 5—Secondary emission characteristic of photosensitized silver.

If light falls on a portion of the mosaic, photoelectric emission takes place, and by losing some electrons that portion of the mosaic acquires a positive charge. Suppose a portion of the mosaic is charged to one volt positive, and consider what happens when this mosaic is scanned, first the part of it having no charge, then the boundary between the dark, the uncharged mosaic and the positively charged lighted mosaic, and finally when the lighted area is scanned.

The scanning spot may be considered as a square brush having an area of one picture element. The charges on the globules or sub-elementary capacitances under the spot are instantaneously equalized, so that one may talk of the potential of the spot and the charge on the spot. The charge on the spot is affected by the charging current, which is equal to the difference between the secondary current and the primary or beam current.

The charging current characteristic shown in Figure 6 is easily derived by subtracting the beam current (unity) from the secondary emission characteristic. In the interval from 0 to 2 volts the charging characteristic is represented very closely by the straight line: $I_{CH} = (\alpha - \beta V)\ I_b = (3 - 1V)\ I_b$, where V is the potential of the target with respect to the collector.

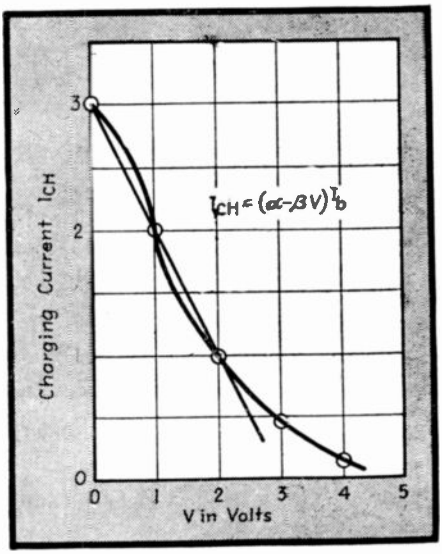


Fig. 6—Charging current characteristic of an Iconoscope mosaic.

SCANNING A DARK AREA

Now suppose the spot is moving in a normal scanning manne
the uncharged portion of the mosaic. If a steady state has been
the current to the signal plate, the potential of the movin
the potential of elements already scanned are all steady
The charging current is therefore

$$I_{CH} = (\alpha - \beta V)\ I_b$$

The mosaic elements after scanning are all charged to the same potential V. Since the capacitance charged per second by the spot is nNC, the charge left by the spot on the mosaic is $nNCV$ coulombs per second, where n is the number of elements per frame, N is frames per second and C is the capacitance of a picture element. Coulombs per second is current in amperes; therefore, the two currents must be equal:

$$I\ (\alpha - \beta V)\ I_b = nNCV \qquad (2)$$

Solving this for V gives

$$V = \frac{\alpha I_b}{nNC - \beta I_b} = \frac{\alpha}{\beta} \cdot \frac{1}{\dfrac{nNC}{\beta I_b} + 1}$$

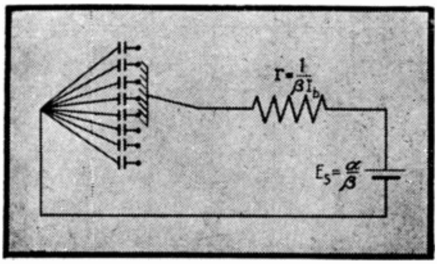


Fig. 7—Equivalent circuit of a dark mosaic undergoing scanning.

Now letting $1/nNC = R$, letting $1/\beta\, I_b = r$ and letting $\alpha/\beta = E_s$ gives

$$V = E_s \frac{R_1}{r + R_1} \qquad (3)$$

The solution in Equation (3) is an exact equivalent of a simple charge of capacitors by a commutating brush in the arrangement shown in Figure 7. The brush, covering a number of capacitors of a total capacitance C and commutating nNC farads per second, charges them through a resistance $1/\beta\, I_b$ from a battery of α/β volts.

SCANNING A BOUNDARY

If at time $t = 0$ the leading edge of a rectangular scanning spot or brush reaches the boundary between dark and lighted portions of the mosaic, a transient condition will prevail until some later time, when a new and different steady state will be reached. As shown in Figure 8, before the leading edge of the spot reached the lighted portion, the current to the mosaic and voltage to which its elements are charged are both steady and of values given. Thus, before $t = 0$, $V = E_s R_1/(r + R_1)$ and $I_0 = E_s R_1/(r + R_1)\, r$.

After $t \underset{>}{=} 0$ the charge coming under the spot is $nNCE_0$ coulombs per second, the charge left over on the mosaic after scanning is $nNCV$ coulombs per second, and the secondary emission charging current is $(E_s - V)/r$. These quantities should satisfy the following equation at all times:

$$VC = \int_0^t \left(nNCE_0 + \frac{E_s - V}{r} - nNCV \right) dt \tag{4}$$

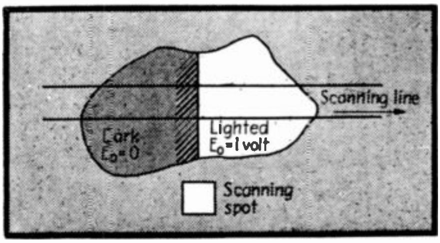


Fig. 8—Scanning from a dark area into a lighted area.

Equation (4) reduces to a linear differential equation which when solved by conventional methods gives

$$\frac{dV}{dt} + \frac{r + R_1}{rR_1C} V = \frac{E_0 r + E_s R_1}{R_1 r C}$$

$$V = E_s \frac{R_1}{r + R_1} + E_0 \frac{r}{r + R_1} \left(1 - e^{-\frac{r + R_1}{rR_1C} t} \right) \tag{5}$$

At $t = \infty$ a new steady state is reached, expressed by

$$V = E_s \frac{R_1}{r + R_1} + E_0 \frac{r}{r + R_1} \tag{6}$$

Scanning Currents and Potentials

The transient charging current while scanning the boundary is given by

$$I_0 = I_{CH} = (\alpha - \beta V)\, I_b = (E_s - V)/r$$

$$= I_0 \frac{1}{r + R_1}\left[E_s - E_0\left(1 - e^{\frac{-r+R_1}{rR_1C}t}\right)\right] \qquad (7)$$

For values of constants encountered in practice, a plot of the charging current *I* is given in Figure 9. A spot which instantaneously equalized all the charges under it has been assumed. That such a condition actually occurs, there is little doubt. However, authorities disagree as to the values of electron densities at which the surface under bombardment begins to be thoroughly conductive.

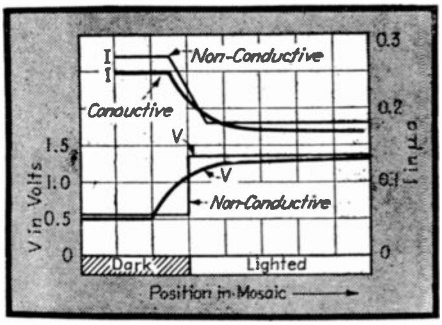


Fig. 9—Transient charging current *I* and mosaic potential *V* during scanning.

If there is no surface conductivity under the electron spot, the nonconductive charging current curve in Figure 9 applies. It may be seen that it is of small importance whether there is, or is not, conductivity under the spot. Except for a small difference in the duration of the transient, the two discharge processes differ very little one from another. The magnitudes of current changes are between 8 and 9 hundredths of a microampere.

The potentials *V* left over on the mosaic by the scanning spot are also almost equal in the two cases, as is shown by the lower curves of Figure 9. A thing to note, however, is that it is hard to expect a potential change on the globules from before to after scanning, greater than 6 tenths of a volt. For a most efficiently activated mosaic this value may rise to one volt.

MOSAIC CAPACITANCE

So far in this study of charging action of the electron bombardment some broad simplifying assumptions were used. This action was studied under the assumption that the mosaic plate is infinite in area, but has constants per unit area of the actual mosaic. As a result expressions were obtained for the video-frequency currents, time constants, potentials, of the mosaic before and after scanning, etc. An infinite area, however, means an infinite capacitance capable of absorbing any charge without a change in voltage. Actually, the iconoscope mosaic has large but finite capacitance, which may be considered infinite only for the upper region of the video frequencies.

At low video frequencies the mosaic capacitance plays a very important part, as is substantiated by the experimental evidence shown in Figure 10. When scanning a mosaic in the dark, the iconoscope output current is not zero. In fact, its peak-to-peak value is of the order of 5×10^{-8} ampere. In the reproduced picture this spurious or dark spot current would produce a great distortion, if it were not compensated for and balanced out.

Now, if a light pattern of horizontal bars is thrown on the mosaic, a square wave of current is generated by the bombardment. The square wave is superimposed on the dark-spot signal. The oscillograms show the wave shape of one field of television signal with its darkspot signal (vertical dark spot). Besides the vertical dark spot there is a horizontal dark spot in the signal, but in these oscillograms, the line frequency and all higher frequencies were filtered out.

SPURIOUS OUTPUT SIGNALS

The spurious signal of the iconoscope is a result of scanning of the mosaic by the bombarding beam. Essentially, the Iconoscope is an a-c device, since its output current flows to its signal plate which is a terminal of a 0.013-μf capacitor. The secondary emission current at a steady-state condition therefore has to average out to a value equal to the beam current, while the bombarding electrons are knocking out five times their number from the mosaic. The excess electrons return to other parts of the mosaic and charge it in their turn. They charge it in a nonuniform manner, contributing to the vertical and horizontal spurious signals.

In general, with the mosaic in darkness at the start of the scan of a frame (or a field, rather) more electrons flow to the collector or

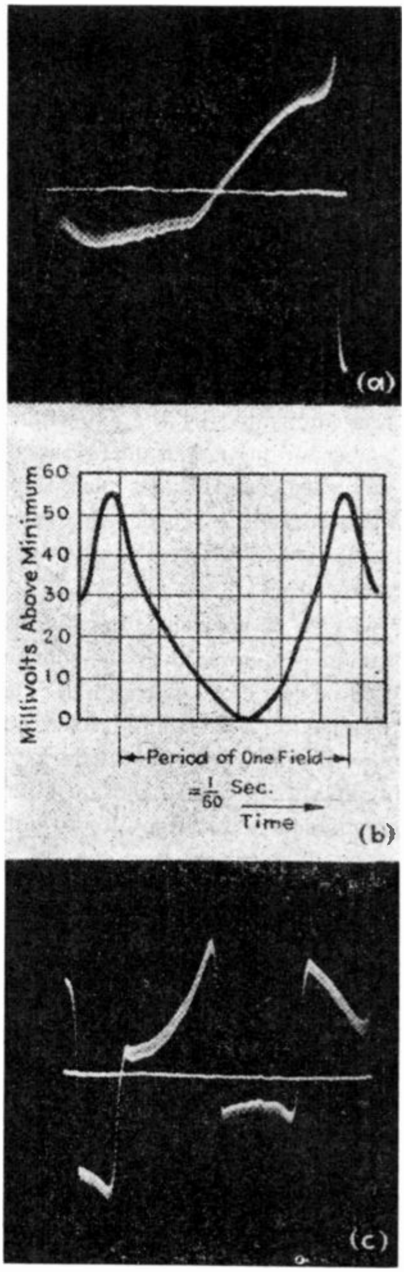


Fig. 10—Oscillograms and curve showing effect of mosaic capacitance. (a)—Signal plate current with mosaic in darkness; (b)—signal plate potential with mosaic in darkness; (c)—signal plate current with mosaic illuminated by 120-cycle square wave pattern.

second anode than are supplied by the beam to the mosaic. The current flows down from the signal plate, charging it negatively. Somewhere in the middle of the scan the flow of electrons to the mosaic becomes equal to the flow to the collector, and the current to the signal plate becomes zero. Then it reverses direction and charges the signal plate in a positive direction, while the flow of electrons to the mosaic is larger than to the collector.

EQUIVALENT GENERATED EMF

Since the Iconoscope is a generator of electrical signals, one may inquire whether it can be represented as a source of an emf generated in some sort of a passive network. Since its output occurs as a current between an output terminal and ground one may inquire as to the value of equivalent emf generated and the magnitude and sense of its equivalent impedance, reactance and resistance. These have been investigated by the old reliable experimental method used in determining the emf and the internal resistance of a battery. The device is loaded with resistances of various values and the output current is read by a suitable meter. A set of simple simultaneous algebraic equations results which when solved yields the desired values.

EFFECT OF BAR PATTERN

Assume that a pattern of alternating black and white bars is thrown onto a mosaic of a normally operating Iconoscope. The resultant signal is then composed of a 180-cycle square wave plus a spurious signal. The voltage output across a normal coupling resistance is observed, then resistors of different values are inserted in series with the normal coupling resistor. The resultant oscillograms are shown in Figure 11. Apparently more than the capacitance of a single picture element is active in the Iconoscope at low frequencies—about one-tenth of the mosaic area in fact, while the emf generated by the Iconoscope is 0.3 to 0.5 v.

CONCLUSIONS

As we have just shown, the fact that the Iconoscope is utilizing electron bombardment and secondary emission does not mean that it is in a class of devices which are foreign to communications engineers. It is a generator of electrical signals, having an internal impedance and a definite electromotive force. Its characteristics are readily measured and used in the design of television systems; while certain of its

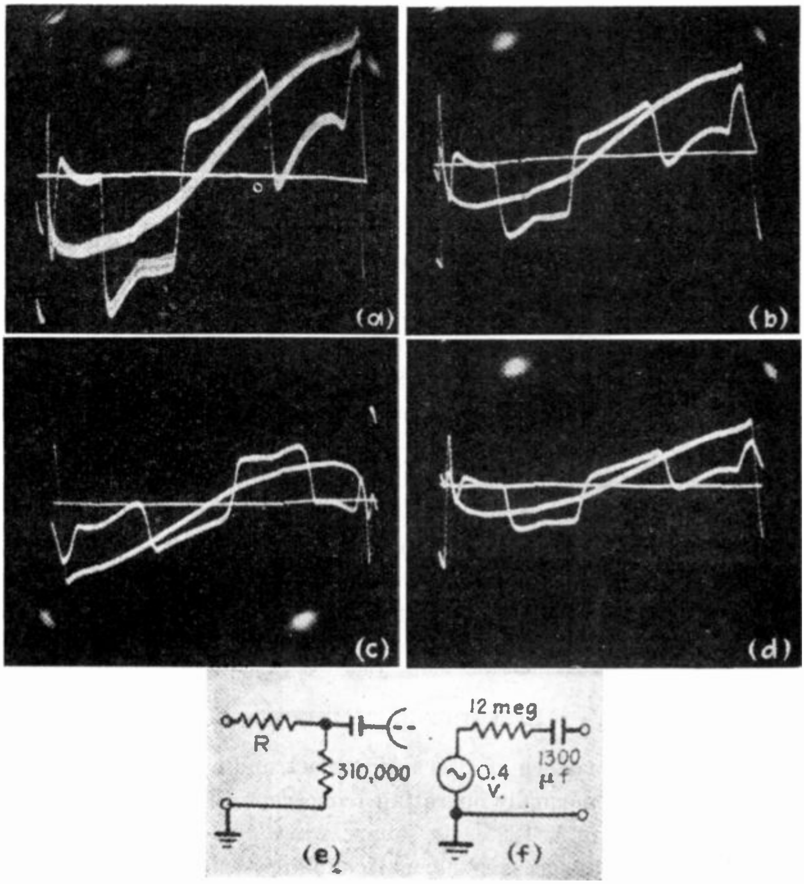


Fig. 11—Internal impedance measurements on an Iconoscope having a beam current of 0.1 μa and a 180-cycle square-wave output signal. Impedance values *R* in Iconoscope circuit (c) for the four oscillograms are as follows: (a)—o ohms; (b)—6 megohms; (c)—12 megohms; (d) 18 megohms. The equivalent circuit of 180 cycles with no backlighting is shown at (f).

actions do not make it an ideal generator of a video signal, it has made possible modern high-definition television.*

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* For a list of references on the subject the reader is referred to the extensive bibliography in "Television," by Zworykin and Morton, John Wiley & Sons, New York, 1940.

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.

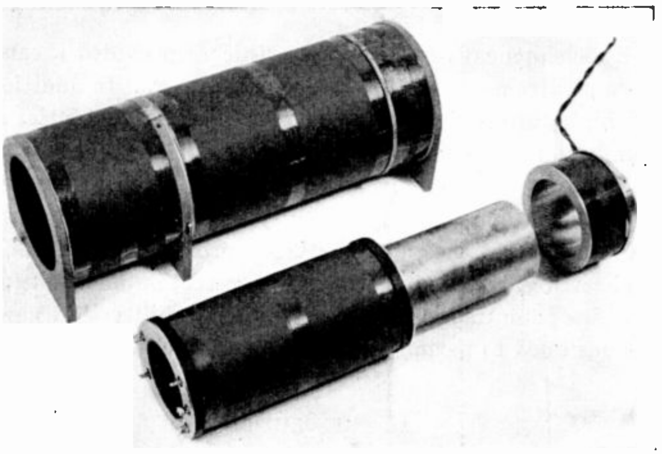

Fig. 1—Focusing, Deflection, and Alignment Coil Assemblies.

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 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.

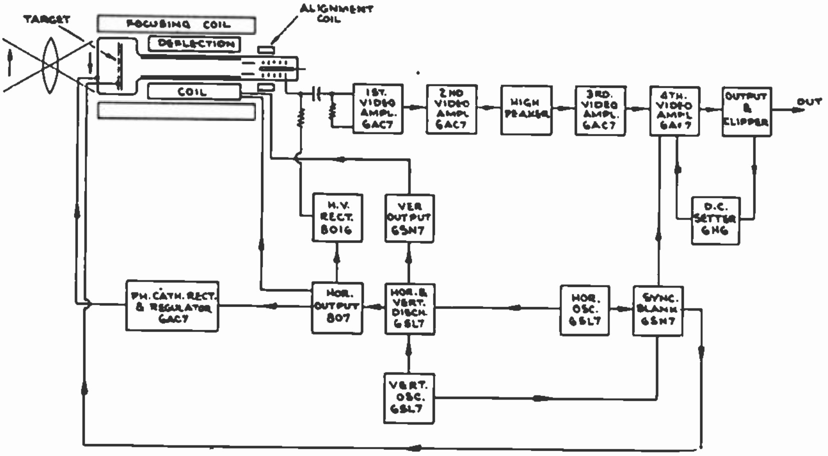

Fig. 2—Camera Block Diagram.

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[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 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.

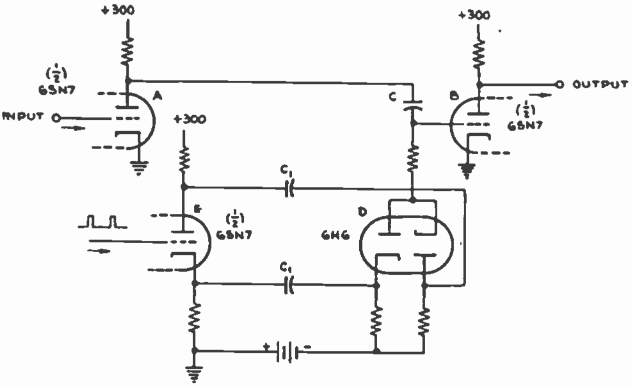

Fig. 3—Direct Current Setting Circuit.

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[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 to the control grid of the pentode rectifier and degenerating any change that might occur.

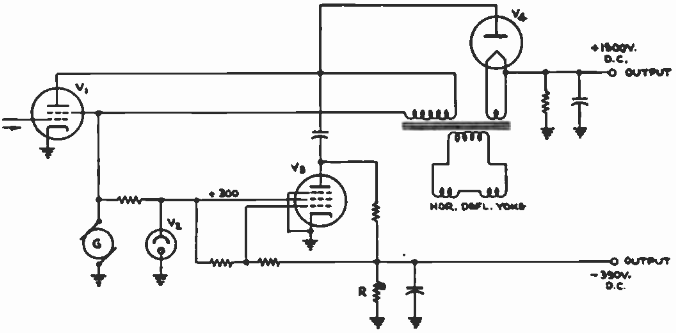

Fig. 4—High Voltage Power Supply and Regulator

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.

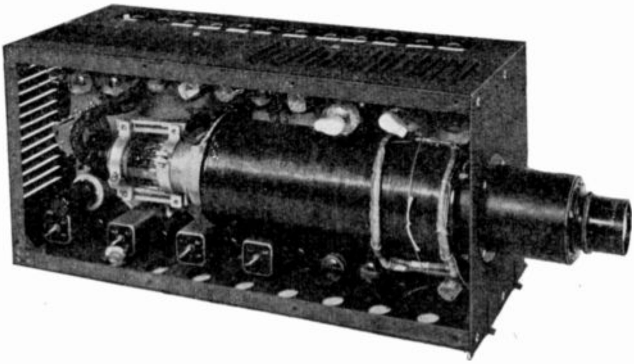

Fig. 5—Top View of Camera Chassis

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.

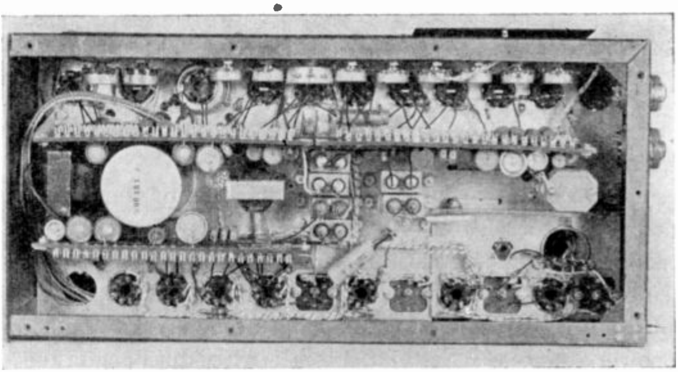

Fig. 6—Bottom View of Camera Chassis

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.

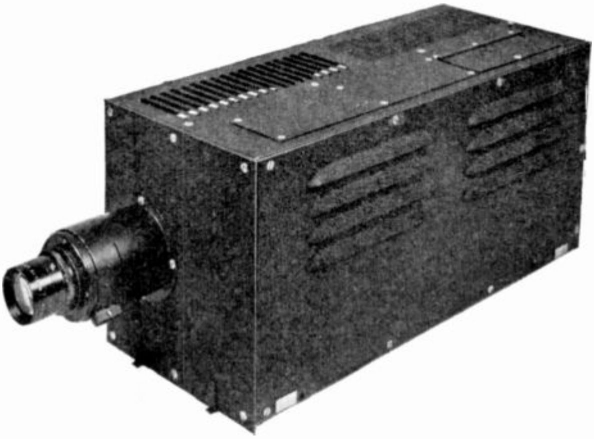

Fig. 7—External View of Camera Assembly with Lens

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

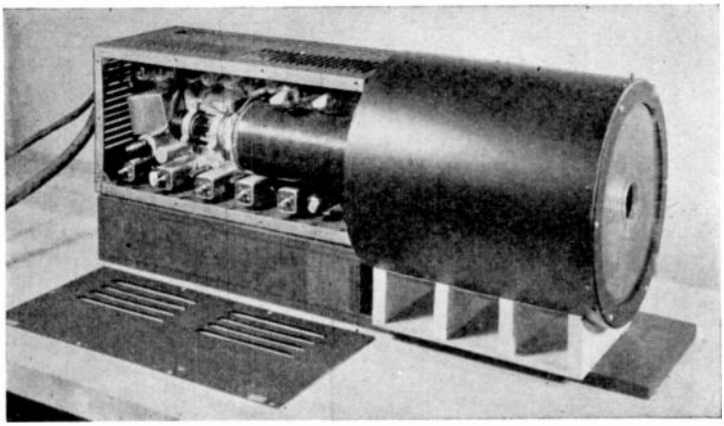

Fig. 8—Camera Assembly with Reflective Optical System

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

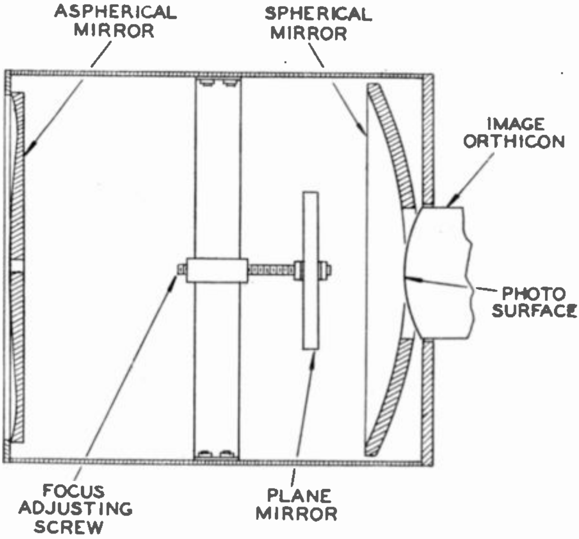

Fig. 9—Construction of the Reflective Optical System

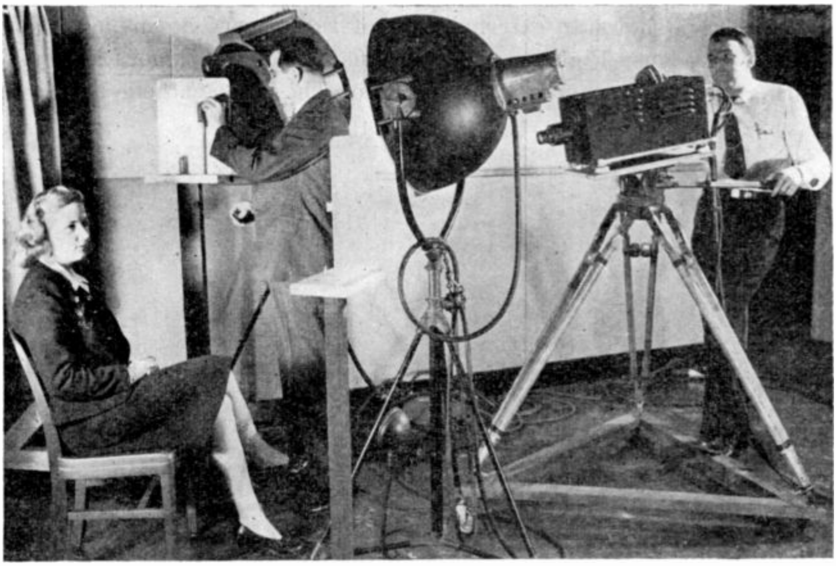

Fig. 10—Camera Demonstration Setup

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.

FIELD TELEVISION*†

BY

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Summary—A resume is given of the history of NBC Field Television. Operations. The four periods of this history, corresponding to four major types of pickup equipment, are outlined and the scope of activities possible during each period is described. Special attention is paid to the fourth period, just now beginning, which is characterized by a greatly widened scope of potential field programs made possible by the new Image Orthicon camera. Some of the characteristics of this new camera, as they affect field operation, are discussed and experience in its use is described.

A recapitulation of NBC television programs reveals that 40% of the program hours broadcast between the opening date of the public service, April 30, 1939, and December 31, 1945 were originated by remote pickup. A total of 1167 program hours were devoted to field events during that period although, because of the war, no activity was recorded for the 16½ months between the middle of May 1942 and the first of October 1943. Television programs originating outside the studio have always been considered to hold an important place in a well-rounded program service, and in television development work early attention was given to providing facilities for originating such programs. Progress in this line of work has been reported from time to time in the technical literature#; therefore the present report will not give the history of these developments in detail. It is intended rather to give an overall picture of the present status with some review of the past work which has lead up to the present state of the art.

Field operations for NBC television broadcasting is divided into four rather distinct periods, determined principally by the characteristics and capabilities of the pickup equipment available during each period. It is true that much progress was continuously being made at all stages in the development of this service on matters of technique and operating procedures, but it is, nevertheless, also true that the characteristics of the pickup equipment were the major factors in determining the scope of field television operation. The four periods referred to above may be characterized as follows in terms of the pickup equipment available in each:

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 # See Various Footnotes.

1. Iconoscope Studio Type Equipment permanently mounted in large vehicle.
2. Orthicon Pickup Equipment permanently mounted in large vehicle.
3. Transportable Suitcase Type Pickup Equipment.
4. Equipment Employing the Image Orthicon.

The fourth period is the one which we are just entering and it gives promise of surpassing by far all previous records with regard to the wide variety of events which will become available for television broadcasting.

In addition to the pickup equipment employed in field operation, it is, of course, necessary to provide a suitable means for transmitting the television signal back to the main studios or to the broadcast transmitter. This may be done by either radio relay circuits or wire lines. Both means have been used successfully in the past and there is every indication that we will see the continued use of both means for at least some time to come. Progress made in microwave radio frequency equipment for a variety of uses during the war will undoubtedly lead to substantially improved radio frequency links for this television relay application in the near future. The present paper will be concerned principally with the pickup equipment and the program limitations imposed by it rather than with the problem of the relay link. Each of the four periods of NBC field television operation will now be considered in more detail.

ICONOSCOPE STUDIO TYPE EQUIPMENT PERMANENTLY MOUNTED IN LARGE VEHICLE

Some indication of the rapid development that has taken place in television pickup equipment may be obtained by examining the facilities available for field operation at the beginning of the NBC television public service on April 30, 1939. A large van type of vehicle was necessary to house and transport the studio type equipment—cameras, camera and power cable, microphone cables, interconnection cables, and the large variety of accessories required to do a proper job on a field set-up. Despite the size of these vehicles (a second vehicle of about the same size housed the permanently mounted radio relay transmitter), they were relatively efficient and mobile. (See Fig. 1) Their size and weight did prevent their reaching marshy or sandy locations, and some areas where parking was at a premium were difficult to reach. They were, however, capable of moving normally in city traffic and on the open road. The great disadvantage from an operating standpoint lay in the fact that the equipment was, of necessity, permanently mounted in the vehicles. The cameras were equipped with 250 feet of camera cable and this accordingly was the radius of action from the vehicle housing the control equipment.

The original equipment installed in this large vehicle employed an iconoscope camera and studio type rack-mounted amplifiers, control equipment and synchronizing signal generator. At the time this equipment was built, the iconoscope was (and in some respects still is) the most satisfactory type of direct television pickup tube available. With medium and high levels of illumination, it produced highly satisfactory pictures and for day-time out-door field television operation where the incident illumination did not fall below several hundred foot candles, the results obtained were generally good. In addition to the fact that the pictures obtained under conditions of very low light levels were degraded by excessive amounts of "dark spot" signal, edge flare, and other defects, it was soon found that the iconoscope had additional limitations in field operation which were not serious in its use in the studio. As regards the use of iconoscope cameras, studio operation has, in effect, a threefold advantage over field operation. In the studio, the lighting is under control of the operator and may be modified to suit the needs of the scene being televised; scenery, back-drops and drapes may be employed having light reflection characteristics suitable for

Fig. 1—Telemobile Units in Rockefeller Plaza, New York City, 1939.

the camera; and finally, in studio operation rehearsal will usually have given the operator a knowledge of the shading problem to be encountered as the program progresses. On the other hand, in field operation —especially outdoors—the lighting is generally not subject to any control, and may fluctuate over a wide range; the scene being televised in many instances may have an unfavorable background or direction of the lighting may be unfavorable; and usually the event being televised is spontaneous and unrehearsed so that the operator has little warning of the changes to be encountered in shading. During this period of the NBC field television operation, the fact that peak field program hours occurred between the months of June and September was not because most field events suitable for television pickup occurred during those months, but rather because most of the *outdoor* events of this kind occurred at that time. The field program curve tapered off in November during the football season and took a sharp drop as the gridiron season ended. During the winter months, an occasional outdoor program was originated and a few indoor programs were attempted using added illumination for the benefit of the iconoscope camera. Many events which would have made good television programs could not be transmitted because of the amount of light required by the iconoscope camera.

Fig. 2—Ceremonies at Opening of New York World's Fair, April 30, 1939. (Television Camera on Platform at Extreme Right.)

Despite the limitations imposed by the equipment in use, several hundred interesting and timely television programs were presented originating at locations as far as 28 miles from the main studios. Although the iconoscope suffers from relatively low sensitivity in comparison with other types of pickup tubes, when an adequate light level is available, the overall quality of the picture obtained with an iconoscope camera is probably superior to that from any other type of pickup device so far used in the program service.

The outstanding program originated with this first field pickup equipment was the inauguration of the television public service on April 30, 1939 when the late President Roosevelt was televised during the opening ceremonies of the New York World's Fair. (See Fig. 2) A program more typical of field operation with this equipment, however, was the pickups of tennis matches at the Westchester Country Club at Rye, New York during the summer of 1939.

ORTHICON PICKUP EQUIPMENT

A major revision in the field pickup equipment was made during the month of September 1939 when a new camera employing the orthicon type of pickup tube was substituted for one iconoscope camera in the mobile unit. For some time thereafter, the unit was operated with a combination of one orthicon camera and one iconoscope camera. The importance of this change in equipment lay in the relatively greater sensitivity of the orthicon tube. Although a direct comparison in sensitivity between the orthicon and the iconoscope is difficult because of their differing contrast characteristic, the orthicon has an effective sensitivity between 3 and 10 times that of the iconoscope for scenes of low incident illumination. In addition, it is essentially free from spurious signals, such as the "dark spot" in the iconoscope. With the orthicon camera, the scope of television field activity was immensely widened. Many events previously unavailable because of the lighting problem now could be satisfactorily televised. The second halves of football games played in the late Fall were now made entertaining as television program fare whereas with the iconoscope, particularly on overcast days, the pickup had been very unsatisfactory. Even with the orthicon camera, in late November and early December, the light condition near the end of the games sometimes was such that the picture quality was seriously degraded. In some instances, the light level dropped so low that floodlights were turned on for the benefit of the players and spectators and under these conditions the orthicon gave very satisfactory results. Another important advantage of the orthicon, in comparison with the iconoscope, is the smaller mosaic size. The area of the mosaic of the orthicon is only approximately one-fourth that of the iconoscope, and this permits the use of substantially smaller lenses to obtain the same angle of view. This is a fairly important item in field operation where a large assortment of lenses must be provided to obtain various viewing angles depending upon the available camera location and the type of coverage contemplated.

Perhaps the most important class of programs made available for the first time by use of the orthicon camera was the large class of indoor sporting events such as boxing, wrestling, ice hockey, basketball, indoor track, etc. Nearly all of these events are presented under conditions of lighting which are satisfactory for the orthicon camera but too low for useable pictures with the iconoscope camera. The importance to television broadcasting of the availability of events of this kind can hardly be over-emphasized. They have formed an important part of the total television program service ever since the first orthicon camera was placed in service. Figure 3 shows the orthicon camera in use for program pickups that are typical of the enlarged scope afforded television field operation by the use of this important development.

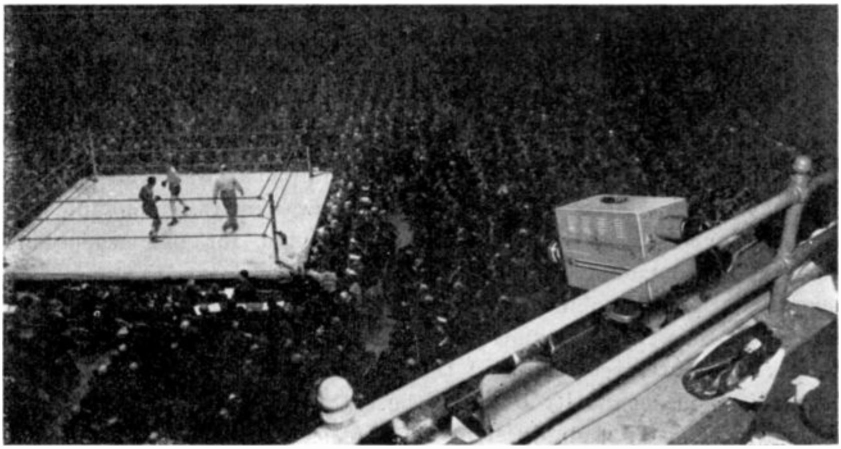

Fig. 3—Orthicon Camera in Use in Madison Square Garden, New York.

Although the orthicon has substantially greater sensitivity than the iconoscope and is essentially free from the spurious "dark spot" signal which plagues the iconoscope, it does have certain disadvantages compared to the earlier, less sensitive type of pickup tube. One of the principal disadvantages is its tendency to "charge up" when subjected to light intensity exceeding a certain threshold value. This phenomenon is due to the inability of the low velocity electron scanning beam to completely discharge areas of the mosaic subjected to high light intensities exceeding a certain threshold value determined by the beam current. This weakness produces especially annoying effects on indoor pickups when photographers' flashbulbs are set off in the field of vision of the camera. This invariably occurs during climactic episodes of the event being televised and frequently portions of the most exciting action are lost to the view of the television audience before stable operating conditions can be re-established in the orthicon pickup tube. Trouble is also encountered in televising outdoor events in bright sunlight when a portion of the scene is in shadow and the rest in sunlight. Changes in the contest being televised which necessitate sudden and rapid panning of the camera from dark shadow to bright sunlight frequently lead to this "charging up" or blocking effect. It is minimized by operation of a special control that temporarily raises the value of the scanning beam current to an abnormally high value in order to dissipate the excessive charge on the mosaic.

The orthicon has a contrast characteristic (frequently referred to loosely as "gamma" characteristic), which is linear over its entire useful operating range. The iconoscope, on the other hand, possesses a contrast characteristic which exhibits a substantial amount of saturation in the higher light ranges and is, therefore, roughly equivalent to a "gamma" of less than unity. This contrast characteristic of the iconoscope complements rather well the corresponding characteristic of the kinescope tube used in most receivers so that the overall contrast characteristic of the system is generally satisfactory. When using the orthicon, however, it is necessary to provide in the video amplifier chains associated with the camera a controllable amount of saturation which will reduce the equivalent "gamma" characteristic to a value more suitable for the kinescope. Generally speaking, the combination of orthicon and its "gamma" correction circuit does not produce quite as satisfactory an overall contrast characteristic as the iconoscope possesses, but it is hoped that future work will improve the "gamma" correction circuits. Despite the previously mentioned fact that the orthicon requires smaller size lenses than the iconoscope for a given angle of view, the sizes of lenses necessary to provide the desired range of camera angles are still substantial. The Type 1850 iconoscope requires a lens of 16″ focal length, when the camera is located 70 feet from a boxing ring, to create images of the contestants which will be large enough when reproduced on the average receiver to provide optimum coverage of a prizefight. A lens of longer focal length and smaller viewing angle makes it difficult for the camera operator to follow the fast-moving contestants about the ring and often results in one of the two fighters being out of the picture. The smaller mosaic in the orthicon tube allows the use of lenses which are approximately one-half the focal length of those used with the iconoscope camera for the same viewing angle. With the orthicon camera, boxing at Madison Square Garden is generally covered with a lens of 8″ focal length, the distance from the camera to the ring being approximately 70 feet. Even with the reduction in lens size made possible by use of the orthicon pickup tube, it has not been considered feasible to attempt the use of lens turrets on these cameras.

TRANSPORTABLE SUITCASE TYPE PICKUP EQUIPMENT

The wider field of television activity permitted by the orthicon camera was still restricted by the limitation due to the location of equipment permanently mounted in a vehicle. The camera cables contained four flexible coaxial cables and thirty-two other electrical conductors and were 1½″ in diameter. Storage and transportation difficulties, plus the need for compensating electrical networks to correct for pulse delay in very long cables were the factors which restricted the length of the camera cables to a practical value of approximately 250 feet. Shorter lengths of 50 feet each were carried for use where the longer lengths were not necessary. The 250 foot radius of activity of cameras was sufficient for most outside pickups when only two cameras were employed. Additional cameras, which would be those located at greater distances, could not readily be accommodated because of lack of space for additional control equipment in the vehicle. Field pickup service was restricted to coverage of events taking place below the fourth floors of buildings. Banquets, important meetings, interesting exhibits and panoramic scenes are among the potential programs which were unreachable with this equipment. The need for portable equipment which could be carried closer to the pickup scene became apparent early in field operations.

The development of the small iconoscope (Type 1848) transportable equipment[1] which was contained in eight boxes approximating the size of suitcases again greatly expanded the scope of the field service. The boxes, weighing approximately 65 lbs. each, were inter-connected on the scene of action to form a complete operating chain of television pickup equipment. This equipment could be used for television pickup either in a vehicle in which it was transported or removed from the vehicle and carried to the upper floors of a building or to other locations inaccessible to the vehicle. (See Fig. 4)

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[1] G. L. Beers, O. H. Schade and R. E. Shelby, "The RCA Portable Television Pickup Equipment", *Proc. I.R.E.*, Vol. 28, pp. 450-458, October, 1940.

The availability of this transportable type of television pickup equipment again opened up new sources of television programs previously unavailable, and also increased the ease of operation in some cases where television pickups had previously been made with equipment permanently mounted in a vehicle. This type of equipment was used in the Rainbow Room atop the RCA Building in Radio City to televise the New Year's Eve festivities there in 1940. When the Republican National Convention of 1940 at Philadelphia, Pennsylvania was televised and the signals transmitted via coaxial cable to New York for broadcasting, both the transportable type of pickup equipment and the equipment permanently mounted in the mobile unit truck were employed, thus providing a four-camera pickup. The four cameras were used to pick up both wide angle and close-up scenes inside the main Convention Hall, for studio type pickup in a small improvised interview studio and for pickups on the sidewalk in front of the Convention Hall. All sessions of the five-day convention were televised.[2,3]

Fig. 4—Complete Single Camera Iconoscope Type Transportable Pickup Equipment.

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[2] O. B. Hanson, "Televising a Political Convention", *RCA REVIEW*, Vol. V, No. 3, pp. 267-282, January, 1941.

[3] H. P. See, "Televising the National Political Conventions of 1940, *Jour. Soc. Mot. Pic. Eng.*, Vol. XXXVI, pp. 82-100, January-June, 1941.

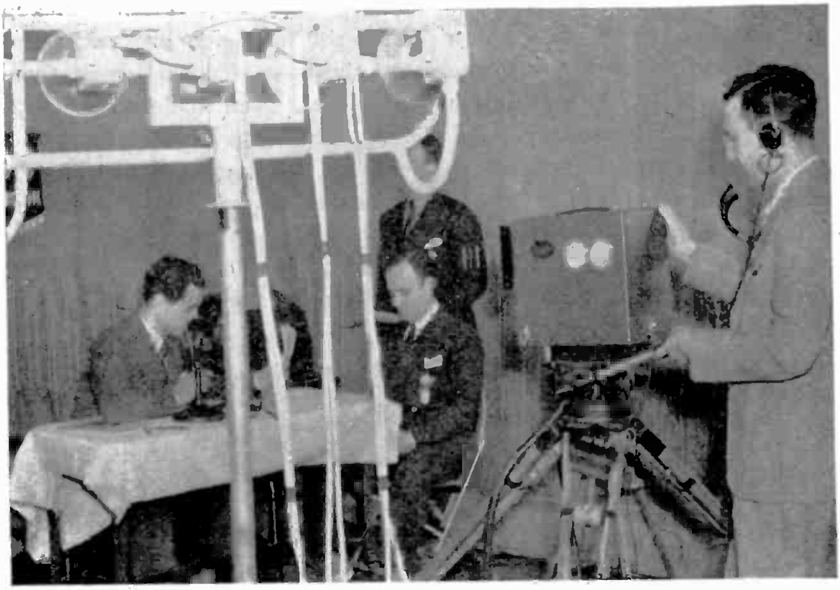

Fig. 5—Transportable Iconoscope Camera in Temporary Studio Set-up at GOP Convention, Philadelphia, 1940.

Fig. 6—Televising the GOP Convention, Philadelphia, 1940. (Orthicon Camera in Foreground, Iconoscope Camera on Right.)

The photographs of Figures 5 and 6 show scenes at the Philadelphia Convention pickup.

Transportable equipment employing orthicon cameras[4,5] became available shortly before the war and a two-camera system of this type has been used for the majority of all NBC field pickups since 1944.[6,7] Figure 7 shows the suitcase type equipment for the orthicon cameras situated on a movable table in a small control room at Madison Square Garden. This is the normal location for the control equipment during the televising of events in Madison Square Garden and its use here illustrates one of the advantages of the transportable type of equipment. Prior to its availability, when televising events in Madison Square Garden, the large mobile unit vehicle was parked at the curb outside the Garden and camera cables and power and communication cables had to be strung in place for each program.

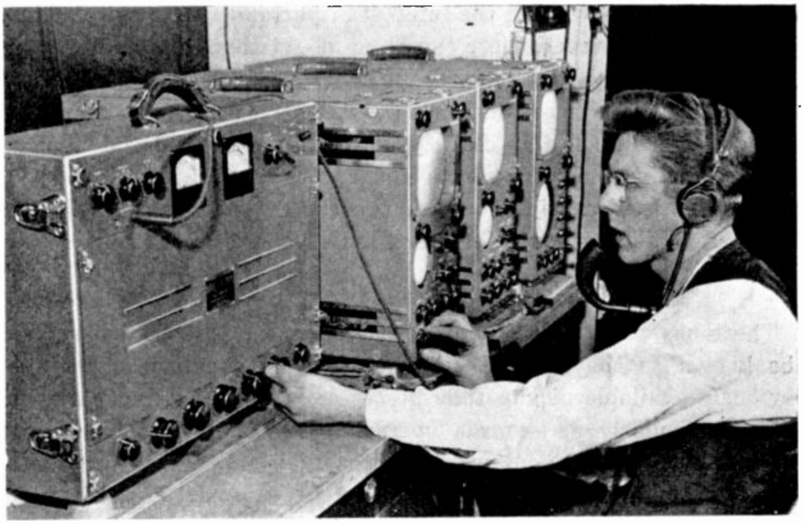

Fig. 7—Television Control Set-up at Madison Square Garden, 1944. (Audio Amplifier on Left, Transportable Orthicon Control Units on Right.)

Although the transportable type of equipment has several distinct advantages, as already indicated, there are some offsetting disadvan-

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[4] Albert Rose and Harley Iams, "The Orthicon, A Television Pickup Tube", *RCA REVIEW*, Vol. IV, No. 2, pp. 186-199, October, 1939.

[5] M. A. Trainer, "Orthicon Portable Television Equipment", *Proc. I. R. E.*, Vol. 30, pp. 15-19, January, 1942.

[6] R. E. Shelby, H. P. See, "NBC's Experience with Portable Television Broadcast Equipment", *Broadcast News*, No. 39, pp. 14-21, August, 1944.

[7] R. E. Shelby, H. P. See, "NBC and Madison Square Garden", *Television*, pp. 2-3, April, 1945.

tages. In comparison with the older style of rack mounted equipment, the suitcase type of equipment is highly condensed and the components crowded rather closely together. This means that maintenance work and trouble-shooting are sometimes more difficult than in the rack mounted type of equipment. The large number of interconnecting cables with their sockets and plugs increases the chance for contact failure and delays occasioned by loss or damage. The smaller size of the monitoring kinescopes and cathode-ray oscilloscopes is somewhat of a handicap in operations compared to the larger size screens used in the older equipment. Convenient accessibility of operating controls has had to be sacrified somewhat in the interest of portability. In spite of these disadvantages, however, there can be no doubt that this type of equipment provides a substantial net gain in television field operation. It is interesting to note in passing that the extensive amount of work done during the war on highly compact military type of television pickup equipment which had to operate under very rigorous conditions will undoubtedly lead to significant improvements in future designs of transportable pickup equipment for the television broadcasting service.

IMAGE ORTHICON EQUIPMENT

There has recently been announced a new type of television pickup tube known as the Image Orthicon which in comparison with any previously available pickup tube possesses rather startling characteristics, particularly as regards operation at extremely low levels of illumination. Development work on this tube had been started prior to the war but it received its greatest impetus as a part of the war-time research on military television. Technical design details and characteristics of the device have been given in a recent paper[8] by Dr. Albert Rose and these will not be repeated here. Very briefly, however, it consists essentially of an orthicon scanning tube with an image electron amplifier section in front of the mosaic and a multiple stage electron multiplier at the output. It is still smaller in size than the Type 1840 orthicon tube, having a photo-cathode area approximately one-quarter that of the Type 1840 orthicon and one-sixteenth that of the larger iconoscopes. The sensitivity of the tube is such that it will produce satisfactory pictures at illumination levels lower by a factor of about 100 than those required for the iconoscope. In addition, it is much less subject to the "charging up" effect which is so troublesome in the case of the Type 1840 orthicon.

The importance of this new tube in television field pickups is self-

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[8] Albert Rose, P. K. Weiner and H. B. Law, "The Image Orthicon. A Sensitive Television Pickup Tube", presented at the I.R.E. Winter Technical meeting, on Jan. 24, 1946.

evident. It removes all practical barriers with respect to operation at low light levels. It seems safe to state that any event which has illumination adequate for direct viewing by an audience can be televised satisfactorily with the Image Orthicon camera.

The smaller length and diameter of the new tube plus its increased sensitivity and small photo-cathode area result in a camera substantially smaller than the standard orthicon field camera. The standard iconoscope and orthicon cameras in use today are considered too large and heavy for efficient handling in the field. The comparatively large mosaic areas of the iconoscope and orthicon tubes require optical systems of appreciable proportions. In contrast, the optical systems used with the Image Orthicon are about the same as those used on standard thirty-five millimeter motion picture cameras. Whether future camera models incorporate the twin lens optical viewfinder, the kinescope viewfinder or some other view-finding device, the reduction in the size of lens with the introduction of this tube makes practical the use of a lens turret. The sizes and weights of lenses included in the complement for a Type 1840 tube are generally considered to be approximately half those necessary for a Type 1850 iconoscope. The Image Orthicon reduces this by approximately one-half again for the same viewing angles. The lack of a lens turret in some cases in the past has imposed a limitation upon the latitude of operations and programming.

There are characteristics of the Image Orthicon which, at least for the present, partially offset some of its advantages. Its signal-to-noise ratio is not as good as that of the iconoscope under conditions of strong illumination, although at low levels of illumination it continues to produce satisfactory pictures far below the levels at which the signals from the orthicon and iconoscope are completely submerged in noise. The lower signal-to-noise ratio of the new tube is generally not noticeable except on scenes which include relatively large dark areas. Test-chart resolution in excess of 500 lines has been obtained with the Image Orthicon, but it has not yet quite equalled the performance of the better iconoscopes in this respect. Models of the tube produced for military use and early samples available for tests in television broadcasting possess rather high infra-red sensitivity. This necessitates the use of optical filters to attenuate the infra-red light when televising most outdoor scenes in daylight—particularly scenes which include appreciable amounts of living green foliage.

The Image Orthicon is more sensitive to ambient temperature than other types of pickup tubes. It does not give maximum performance until it has warmed up to approximately 100° F., and when used outdoors in cold weather auxiliary heating units are sometimes needed. When the tube is too cold, its resolution may be impaired and retentivity of the electrical "image" on the target will be abnormal, producing excessive smearing in the reproduced picture whenever the camera is panned or when rapid motion occurs in the scene.

One important advantage which the Image Orthicon has over the Type 1840 orthicon is its ability to handle a very wide range of light values. If the various electrode potentials are properly set, a change in the scene from deep shadow to brilliant sunshine in outdoor pickups is readily accommodated without serious degradation in the transmitted picture and without the necessity for instantaneously coordinated readjustment of controls, as in the case of the Type 1840 orthicon. In one outdoor test, the Image Orthicon was adjusted for optimum performance with the iris on the pickup lens set for an opening of *f* 32. Without changing any other control, the iris setting was then changed to *f* 8, thus increasing the amount of light on the photocathode by a factor of 16. To the casual observer, at normal viewing distance, there was no appreciable change in the transmitted picture. In an indoor test at Madison Square Garden, it was found that a single suitably chosen setting of all controls would give acceptable results when the Image Orthicon camera was panned from the dimly-lighted outer fringes of the audience to the brilliantly-illuminated boxing ring at the center of the arena.

While it is true, as indicated above, that the Image Orthicon possesses great practical flexibility under varying conditions of illumination, it is also true that *peak performance* will be obtained only when the settings of the iris and other controls are proper for the brightness of the scene being televised. Close inspection of the transmitted picture shows appreciable loss of information in the highlights if the light image on the photocathode is excessively bright, and signal-to-noise ratio will be lowered by operating the beam current at a value greatly above that required to discharge the target. In preparing for a television pickup, after the Image Orthicon has warmed up to normal operating temperature, the entire scene to be televised should be explored with the camera to determine, to the extent possible, the upper and lower limits of reflected light to be expected from the scene during the program. Settings of the iris, beam current, and other controls should be made in the light of this test, and if wide fluctuations in brightness are to be encountered, plans should be made for readjusting beam current and iris opening at proper times during the program.

Television broadcasting experience with the Image Orthicon camera has included the following pickups:

1. *Herald-Tribune Forum, Grand Ballroom of Waldorf-Astoria Hotel.* Moderate illumination as normally used for audience.
2. *Navy League Dinner, Grand Ballroom of Waldorf-Astoria Hotel.* Moderate illumination as normally used for audience.
3. *Army-Navy Football Game.* The day was an unusually bright one for the season, and offered no test of sensitivity, but did afford a good comparison with orthicon cameras under conditions of high-level illumination. (See Fig. 8)

Fig. 8—Image Orthicon Camera Televising Army-Navy Football Game at Philadelphia Municipal Stadium, December first, 1945.

4. *Mayor-Elect O'Dwyer of New York City, from his campaign headquarters in the Commodore Hotel on Election Day.* Illumination provided by two 100 watt lamps in a frosted glass shade, and a 40 watt lamp for relieving shadow areas on faces.
5. *New Year's Eve Celebration in Times Square, New York City.* Illumination as normally present in Times Square at night—from electric signs, street lamps, theatre marquees, show windows, and automobile lights.
6. *Memorial Service at Lincoln Monument, Washington, D. C. on February 12, 1946.* A program originated jointly by stations WABD, WCBW and WNBT and transmitted to New York over the coaxial cable of A. T. and T. to inaugurate television service over this cable.

In addition to the foregoing on-the-air programs originated with the Image Orthicon camera, a number of successful test pickups have been made. These include the following:

1. *Baseball game at Polo Grounds, New York City.* The day was a bright one, and it was found that the minimum lens stop available—*f* 32—gave more than the optimum amount of light. Depth of focus was, naturally, no problem. A filter had to be used to reduce the infrared light reflected from the grass of the playing field. Most observers felt that the Image Orthicon camera had a net advantage over the Type 1840 orthicon camera, which was set up for comparison.

Fig. 9—Image Orthicon Camera (Foreground), in Comparative Test with Orthicon Camera, Televising Rodeo at Madison Square Garden.

2. *Strollers on the Mall at night in Central Park, New York City.* The only illumination was that provided by the normal light fixtures in the park.

3. *Scenes from the Rodeo in Madison Square Garden, New York City.* The dark tan-bark on the floor, and the relatively low light level employed for many events to avoid blinding the contestants, had made this a relatively unsatisfactory program for orthicon cameras. The Image Orthicon camera was able to do an excellent job on all events. (See Fig. 9)

4. *Television Pickup of Standard Sound Broadcast.* No special lighting of any kind was employed, and those in the studio did not even know that the test was being made.

5. ***"Stunt" Pickups in the Studio.*** Successful pickups were made using only the light of a pocket flashlight, or a single candle, or one match, and also in total darkness, using invisible infra-red illumination.

The importance of the Image Orthicon development to field television broadcasting is emphasized by statistics of operation which show that in the past more than one-half of the seven hundred odd field programs have originated indoors under artificial illumination—despite the fact that many potential indoor programs had to be passed up because cameras were not sensitive enough to give acceptable results with the illumination available. In the future, it is probable that the percentage of indoor programs will go even higher, now that there are virtually no technical limitations on the televising of such events. The economic advantages of the new camera are by no means insignificant, since the cost of providing special illumination for some events televised in the past has been a major item of expense. On the basis of experience to date, it seems safe to say that the Image Orthicon represents the greatest single advancement so far made in field television.

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.

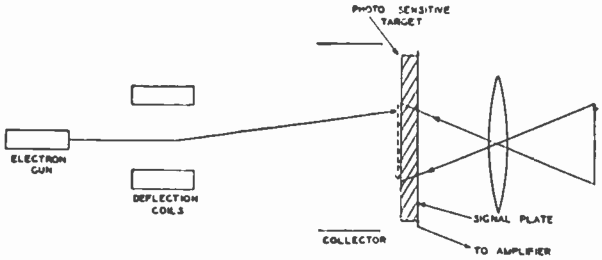

Fig. 1—Typical parts of storage type of pickup tube.

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

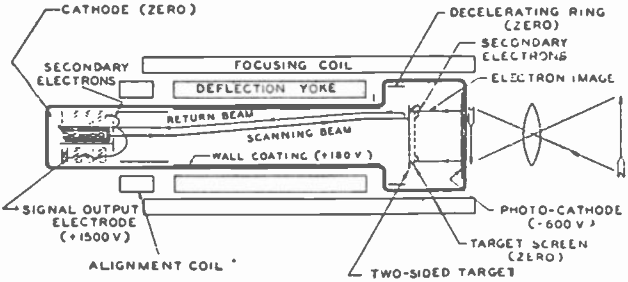

Fig. 2—Diagram of the image orthicon.

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.

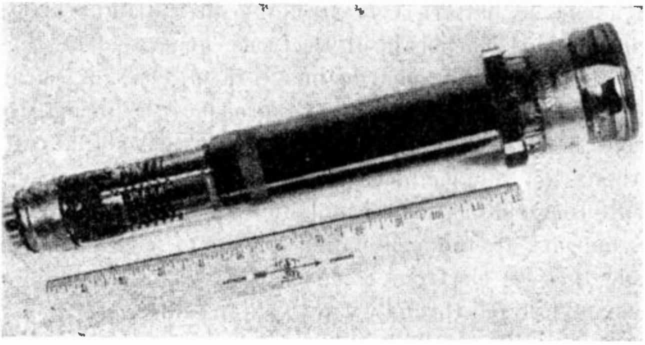

Fig. 3—The image orthicon.

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),

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[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 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.

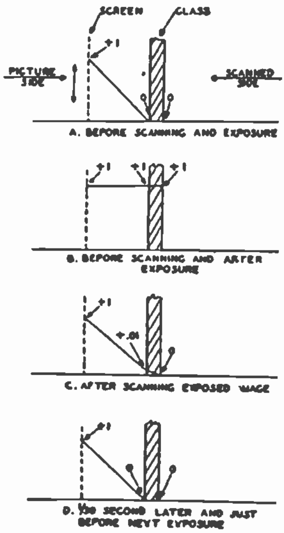

Fig. 4—Target potentials during a typical scanning cycle.

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

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[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

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[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 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.

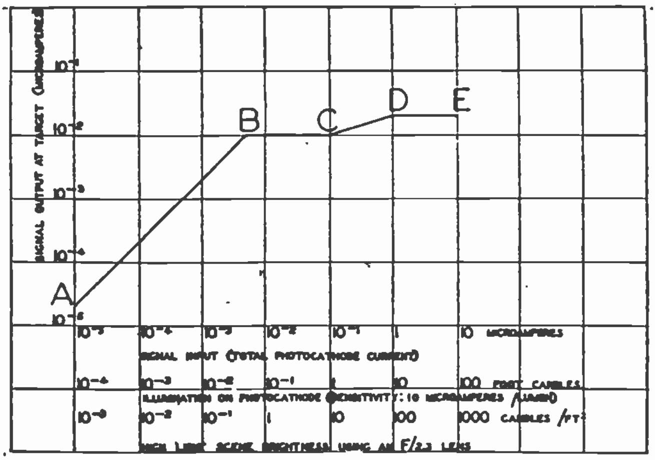
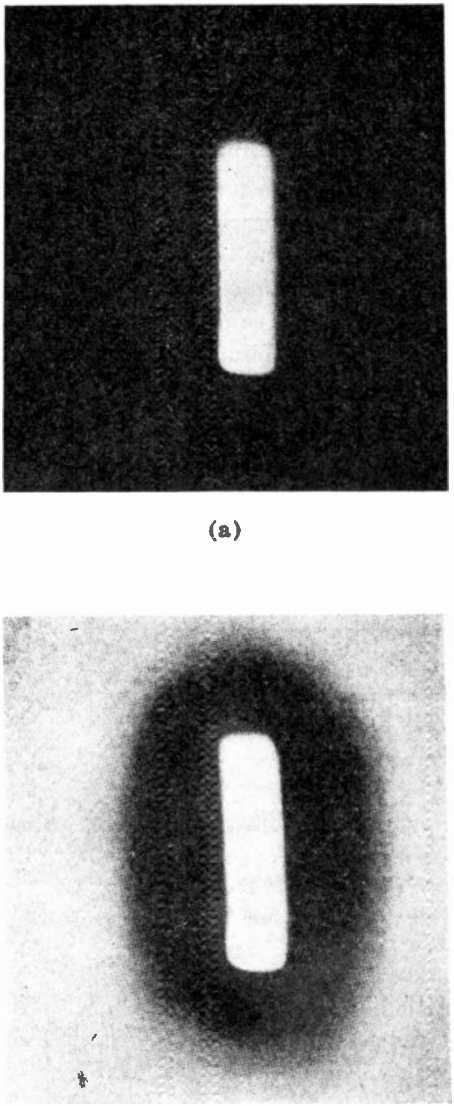


Fig. 5—Signal versus light characteristic.

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.

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[12] D. B. Langmuir, "Theoretical Limitations of Cathode-Ray Tubes," *Proc. I.R.E.*, vol. 25, pp. 977-991; August, 1937.

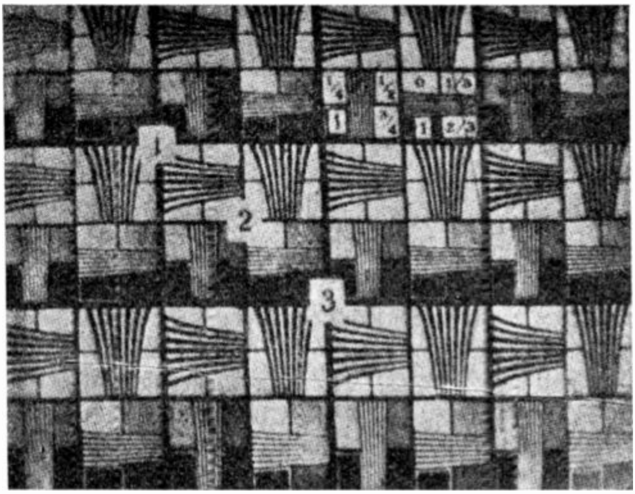


Fig. 7—Test pattern transmitted by image orthicon.

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.

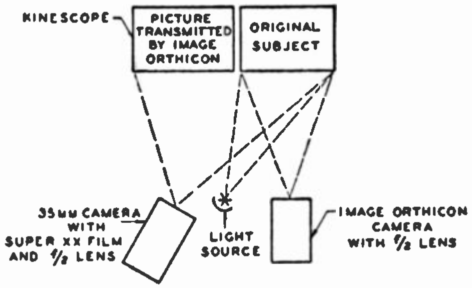
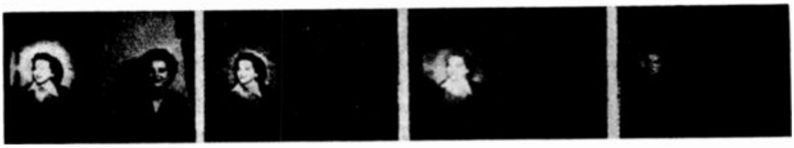


Fig. 9—Setup for comparing the sensitivities of image orthicon and photographic film.

individuals. Among these are H. B. DeVore,[13] L. E. Flory,[14] R. B. Janes,[15] H. A. Iams,[14] G. L. Krieger,[14] 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 UNIFIED APPROACH TO THE PERFORMANCE OF PHOTOGRAPHIC FILM, TELEVISION PICKUP TUBES, AND THE HUMAN EYE*†

By

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Summary—The picture pickup devices—film, television pickup tube, and eye—are subject ultimately to the same limitations in performance imposed by the discrete nature of light flux. The literature built up around each of these devices does not reflect a similar unity of terminology. The present paper is exploratory and attempts a unified treatment of the three devices in terms of an ideal device. The performance of the ideal device is governed by the relation

$$\text{scene brightness} = \text{constant}\ \frac{(\text{signal-to-noise ratio})^2}{\text{picture element area} \times \text{quantum efficiency}}$$

The three devices are shown to approximate this type of performance sufficiently well to use it as a guide in treating their common problems. Simple criteria are derived for characterizing the performance of any one device as well as for comparing the performance of different devices. For example, quantum efficiency is used to measure sensitivity; the signal-to-noise ratio, associated with a standard element area, is used to measure both resolution and half-tone discrimination. The half-tone discrimination of the eye governs the visibility of "noise" in the reproduced picture and, in particular, requires that pictures be photographed or picked up at increased scene brightness when the brightness of the reproduction is increased. The observation and interpretation of visual "noise" are discussed.

INTRODUCTION

THERE are three picture pickup devices that have separately been the subject of considerable investigation. These are the human eye, motion picture film, and television pickup tubes. For each of these, a large technical literature has been built up relatively independently of the others. The language, the units, the concepts, and the conclusions of the separate arts are not in a form that allows them to be readily compared. This situation is understandable in the early stages of the arts because the primary emphasis is then to get something—anything—that will transmit a usable picture. As the art progresses, however, interest shifts naturally to an examination of the

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* Decimal Classification: R583.11

† Presented May 10, 1946, at the SMPE Technical Conference in New York, N. Y. Reprinted from *Jour. Soc. Mot. Pic. Eng.*, October, 1946.

theoretical limits of expected improvements. Such an examination is especially significant because all three devices are subject ultimately to the same simple statistical limitations arising from the discrete nature of light flux. The time is opportune for the three devices to profit from a consideration of their problems in common terms.

Some illustrations will make the present situation clear. In films, graininess is a familiar concept. Its origin, control, and visual effects have been treated extensively and for a long time. In pickup tubes, signal-to-noise ratio is an ever-present consideration for getting pictures of good quality. For human vision, interest has frequently been centered on the minimum discernible contrast. There is good reason now to say that graininess, signal-to-noise ratio, and minimum discernible contrast are only three different names for the same property of a picture pickup device. Again: the limiting resolution of film is a standard and advertised characteristic; the frequency response curve of a television pickup tube is an important specification of the tube's performance; the minimum resolvable angle of the eye is a well-known figure and one which, perhaps, has received more than its just share of attention. It is obvious that in all three instances, an attempt has been made to count the number of separate picture elements.

A third illustration concerns sensitivity. There is little need to remind one of the variety and confusion of sensitivity scales that have been proposed for film. On the other hand, the sensitivity of a television pickup tube can, with reasonable adequacy, be defined by its microampere signal output per lumen input. The sensitivity of the eye has variously, and often with deliberate dramatic emphasis, been described in terms of the farthest distance at which one can still see a lighted candle; the order of magnitude of the faintest visible star; the number of lumens falling on the retina necessary for a visual sensation; and so on. Only recently have there been more fundamental attempts to measure the sensitivity of the eye in terms of its quantum efficiency.

These illustrations serve to show, first, that the basic properties of a picture pickup device—resolution, sensitivity, and contrast discrimination—are indeed of common concern to the eye, film, and pickup tube; and, second, that the specification of these properties has not enjoyed an appropriately common treatment.

The purpose of the present discussion is to explore the extent to which such a common or unified treatment is both possible and profitable.

The order of the discussion will be:

(1) The development of the properties of an ideal picture pickup device;

(2) The examination of eye, film, and pickup tube for the purpose of finding out how well they approximate ideal performance;

(3) A re-examination of a number of current problems in the light of (1) and (2).

It will become clear that the performance of an ideal device is completely specified by a single number, the quantum efficiency of its photo process, taken together with some simple optical relations; that the performance of eye, film, and some pickup tubes approach sufficiently close to ideal performance to suggest a unified approach to many of their current problems and that such an approach leads to simplifying concepts.

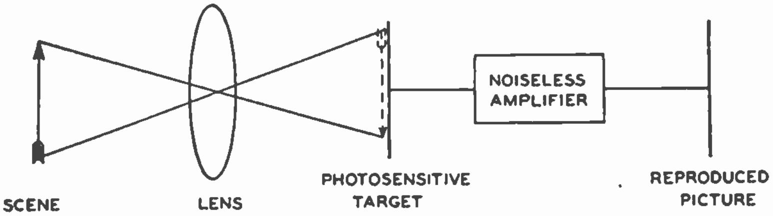

Fig. 1—Essential parts of a picture pickup system.

IDEAL PICTURE PICK-UP DEVICE

Figure 1 shows the essential parts of a system for picking up and reproducing a picture. Attention will be centered on the target of the pickup device, and, in particular, on one picture element of that target. A picture element is here taken to be an element of area of arbitrary size, not necessarily the smallest resolvable area. Let that element have a length of side h, and absorb an average number N, of quanta in the exposure time allowed. The absorption of each quantum will give rise to a separate event such as the release of an external photoelectron, or an internal photoelectron or the dissociation of a molecule. These are uncorrelated chance events. For this reason, the average number N has associated with it fluctuations whose root mean square magnitude is the square root of the average number. Thus, if N is taken to be the measure of the signal, $N^{1/2}$ is a measure of the smallest discernible difference in signal. In particular, the ratio

$$\frac{N}{N^{1/2}} = N^{1/2}$$

is the signal-to-noise ratio. We may write, therefore,

$$\text{Signal-to-noise ratio} \equiv R = N^{1/2} \tag{1}$$

and the geometric relation:

$$\text{Scene brightness} \equiv B = \text{constant}\,\frac{N}{h^2}. \tag{2}$$

Combination of Equations (1) and (2) yields:

$$B = \text{Constant}\,\frac{R^2}{h^2}. \tag{3}$$

Equation (3) is the characteristic equation for the performance of the ideal picture pickup device. It must be emphasized that Equation (3) is not concerned with the particular mechanism used to generate a picture so long as full use is made of all the absorbed quanta. For this reason, it is meaningful to inquire whether the performance of such diverse mechanisms as the eye, film, and pickup tubes can all be described by the same characteristic equation.

Equation (3) defines the scene brightness B required to transmit a picture having a signal-to-noise ratio R associated with picture elements of linear size h. It says that the scene brightness must be increased as the square of the signal-to-noise ratio demanded, and as the square of the number of lines in the picture, the number of lines being proportional to $1/h$.

The constant term on the right-hand side of Equation (3) contains, among other parameters, the quantum efficiency of the photo process. It is this quantum efficiency* alone which sets the performance range of the ideal pickup device. The complete constant term will be given later. For the moment, it will be useful to examine a plot of Equation (3).

Figure 2 is a plot of Equation (3) for several values of scene brightness. Figure 2 shows that the signal-to-noise ratio increases linearly with the size of picture element considered. In particular, there

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* If the term "ideal pickup device" were to receive its full emphasis, the quantum efficiency of the photo process should, of course, be taken to be 100 per cent. The emphasis here, however, is on the complete utilization of all absorbed quanta rather than on the absorption of all incident quanta.

is a smallest element which is determined by the smallest signal-to-noise ratio that can be observed. The smallest element would be called the limiting resolution. The smallest observable signal-to-noise ratio has often been taken to be unity. Actually, by virtue of its statistical origin, the smallest observable R is a function of how often one prefers to have his observations correct. This much is verifiable both from analysis and from the use of physical instruments as observers. For a human observer, tests[1] have been made which suggest a threshold value of R in the neighborhood of five. Whatever this threshold is, one may draw on Figure 2 a horizontal line whose intersections with B_1, B_2, and B_3 mark the limiting resolutions for the several scene brightnesses.

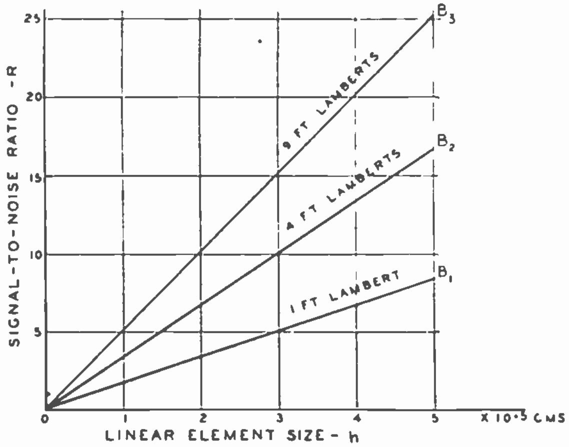

Fig. 2—Performance curves for an ideal picture pickup device.

The complete form of Equation (3) may be readily obtained[2] from well-known optical relations and is

$$B = 2.8\, f^2 \frac{R^2}{t h^2 \theta} 10^{-13} \text{ ft-L.} \qquad (4)$$

Here f = the f/value (numerical aperture) of the lens

t = exposure time (seconds)

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[1] Romer, W., and Selwyn, E. N., "An Instrument for the Measurement of Graininess", *Phot. Jour.*, 83, (1943), p. 17.

[2] Rose, A., "The Relative Sensitivities of Television Pickup Tubes, Photographic Film and the Human Eye", *Proc. I.R.E.*, 30, 6 (June 1942), p. 295.

θ = quantum yield of the photo process ($\theta = 1$ means 100 per cent quantum efficiency)

h = length of side of element (centimeters)

1 lumen = 1.3×10^{16} quanta per second (average for white light).

If one takes the hyperfocal distance as a measure of depth of field, the performance of the pickup device is completely specified by Equation (4) together with the relation[3]

$$\text{hyperfocal distance} = \frac{FD}{2h} \tag{5}$$

where F = focal length of lens

D = diameter of lens.

Complete specification means that one selects the desired values for the hyperfocal distance, exposure time, signal-to-noise ratio, angle of view, and size and number of picture elements, and from them computes the scene brightness required.

The scale factors for the curves of Figure 2 are based on Equation (4) with the choice of $f = 2$, $t = 1/30$, $\theta = 1$. These curves show what may be expected from an ideal device with 100 per cent quantum efficiency.

TELEVISION PICKUP TUBES

No operable pickup tube has yet been reported which completely fulfills the properties of the ideal pickup device. The effective exposure time of the image dissector,[4] or other nonstorage devices, is limited to a picture element time and such devices are correspondingly insensitive. The performance of the iconoscope[5] and orthicon[6] is limited by noise currents inherent in the primary photo process. The image noise currents inherents in the primary photo process. The image orthicon[7] (Figure 3) goes a long way toward removing this limitation

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[3] DeVore, H. B., and Iams, H. A., "Some Factors Affecting the Choice of Lenses for Television Cameras", *Proc. I.R.E.*, 28, 8 (Aug. 1940), p. 369.

[4] Farnsworth, P. T., "Television by Electron Image Scanning", *Jour. Frank. Inst.*, 218, 4 (Oct. 1934), p. 411.

[5] Zworykin, V. K., Morton, G. A., and Flory, L. E., "Theory and Performance of the Iconoscope", *Proc. I.R.E.*, 25, 8 (Aug. 1937), p. 1071.

[6] Rose, A., and Iams, H. A., "The Orthicon, a Television Pickup Tube", *RCA REVIEW*, 4, 2 (Oct. 1939), p. 186.

[7] Rose, A., Weimer, P. K., and Law, H. B., "The Image Orthicon, A Sensitive Television Pickup Tube", *Proc. I.R.E.*, 34, 7 (July 1946), p. 424.

in so far as the high light signal-to-noise ratio of its output is, within limits, determined by the signal-to-noise ratio in the primary photo process. It is handicapped, as are the other storage-type tubes, mainly by having as much noise in the low light portions of a picture as in the high lights. Equation (4) may, however, be used to describe the performance of the image orthicon if signal-to-noise ratio is interpreted to mean the signal-to-noise ratio in the high light portions of the picture.* The quantum yield of the primary photo process is about 0.01 and the noiseless amplifier to be compared with Figure 1 is its electron multiplier.

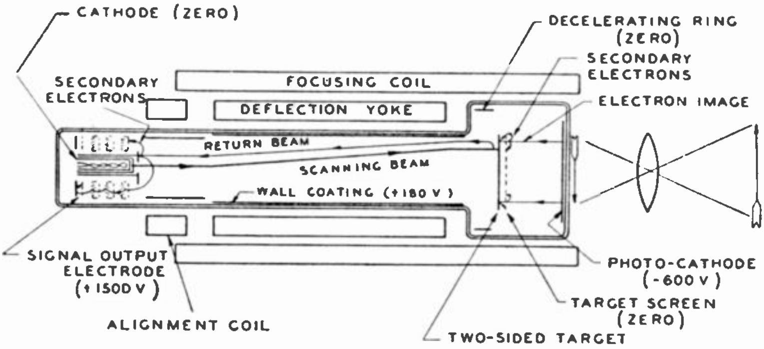

Fig. 3—Image orthicon (schematic).

PHOTOGRAPHIC FILM

One does not readily think of film as having a signal-to-noise ratio. Yet, the separate grains randomly situated in film are immediately comparable with the separate and randomly spaced electrons in the scanning beam of a television pickup tube. And, in fact, a number of recent objective measurements as well as analyses of graininess have led to the expression[8]

$$\Delta D \doteq \frac{\Delta T}{T} = \text{constant} \times a^{-1/2} \qquad (6)$$

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* The beam current used to scan the target must be sufficient to discharge the high light portions of the picture. Under these conditions, the signal-to-noise ratio inherent in the beam is approximately that of the high lights. The same beam current, however, scans the low lights and adds appreciable noise over and above the noise inherent in the low lights.

[8] For summary of literature, see: Jones, L. A., and Higgins, G. C., "The Relationship Between the Granularity and Graininess of Developed Photographic Materials", *Jour. Opt. Soc. Amer.*, 35, 7 (July 1945), p. 435.

where ΔD and ΔT are the r-m-s deviations in density and transmission, respectively, of an area a of film.** With the notation of Equations (1) and (2), $\frac{\Delta T}{T} = R^{-1}$ and $a^{-1/2} = h^{-1}$ and one may write for film

$$R = \text{constant} \times h. \tag{7}$$

The value of this constant is proportional to the reciprocal grain diameter. There is good evidence that, for the same type of photographic grain, the film speed is proportional to the grain area. The last two statements combined with Equation (7) give

$$B = \text{constant}\,\frac{R^2}{h^2}$$

just as for the ideal device [Equation (3)]. One can accordingly use Equation (4) to describe the performance of film with the understanding that the ratio R^2/h^2 is characteristic of film with a given average grain diameter and changes in R^2/h^2 are obtained by use of other films with different average grain diameters. The quantum yield is the reciprocal of the number of incident quanta required to make a grain developable* and from published statements[11] is in the neighborhood

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** Equation (6) obviously cannot hold for values of a in the neighborhood of and less than the grain size. Krevald and Scheffer[9] and Raudenbusch[10] have observed such departures and more recently Jones and Higgins[14] have reported them. The problem is further involved by a range of grain sizes in any one film.

* Strictly, this use of the term "quantum yield" is in accord with its normal definition only if a grain is made developable by the absorption of a single quantum. If more than one quantum needs to be absorbed for this purpose, the process still may be looked upon for noise computations as the equivalent of the absorption of one quantum because the noise arises mostly from the random distribution of grains rather than from the fluctuations in rate of absorption of light quanta.

[9] van Krevald, A., and Scheffer, J. C., "Graininess of Photographic Material in Objective and Absolute Measure", *Jour. Opt. Soc. Amer.*, 27, 3 (Mar. 1937), p. 100.

[10] Raudenbusch, H., "Measurements of Graininess and Resolution of Photographic Film", *Phys. Zeits.*, 42 15/16 (Aug. 1941), p. 208.

[11] Silberstein, L., and Trivelli, A., "Quantum Theory of Exposure Tested Extensively on Photographic Emulsions", *Jour. Opt. Soc. Amer.*, 35 2 (Feb. 1945), p. 93. (The writers of this paper avoid emphasizing the physical implications of their analysis. At the same time they do interpret their results to show that the intrinsic sensitivity of film is increased by longer development times. So far as other measurements have shown that the increase in speed resulting from long development time is paralleled by an increase in graininess, the present paper would argue that the intrinsic sensitivity is unchanged but that the developed grains are made larger by longer development time.)

of 0.001. The noiseless amplifier to be compared with Figure 1 is the complete development of a silver grain after only a few silver atoms have been formed by the action of the light.

HUMAN EYE

Equation (4) is not immediately applicable to the human eye because there is no way of directly measuring the signal-to-noise ratio that the brain perceives. It is necessary, therefore, to replace signal-to-noise ratio by its equivalent in terms of minimum discernible contrast in the test object viewed.** The signal-to-noise ratio R has already been referred to as a measure of the minimum discernible difference in signal. This allows one to write with reasonable assurance

$$\text{Minimum discernible contrast} \equiv C = \frac{\text{Const.}}{R} \times 100 \text{ per cent.} \qquad (8)$$

To get a value for the constant, let C take on its maximum value, *viz.*, 100 per cent. This defines the constant to be equal to the minimum perceivable value of R. As mentioned earlier, the measurements of Romer and Selwyn may be interpreted to give a value of about five. Unpublished measurements by O. H. Schade on television pictures yield a value of three. The determination of this constant is of considerable importance in estimating the quantum efficiency of the human eye and deserves more experimental work.† For the present it will be included as an undetermined constant. Substitution of Equation (8) in Equation (3) gives

$$B = \text{constant} \frac{1}{C^2 h^2} \qquad (9)$$

for the characteristic equation which the eye would satisfy if its performance were "ideal." Equation (9) may be rewritten with the minimum resolvable angle α in place of distance h to make it more readily comparable with published data. Thus,

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** Contrast is defined as $\frac{B_L - B_D}{B_L} \times 100$ per cent, where B_D is the brightness of a gray test object immersed in a white surrounding of brightness B_L.

† An interpretation of the experimental results of Jones and Higgins[11] in which the blending distances and signal-to-noise ratios were measured for the same films also leads to a value of about five.

$$B = \text{constant}\,\frac{1}{C^2\,\alpha^2}\cdot \qquad (10)$$

How well the performance of the eye matches Equation (10) may be seen from Figures 4 and 5. Figure 4 shows a plot[12] of C versus α^{-1} for a large range of scene brightnesses and, as expected from Equation (10), the data fall closely on 45-degree lines. Data in the immediate neighborhood of $\alpha = 1$ minute and $C = 2$ per cent are omitted because these represent limits to the performance of the eye set by other than statistical considerations. The smallest angle that the eye can resolve at high lights, for example, is set by the physical size of the retinal elements or cone structure. A more precise treatment would include, and be slightly modified by, the shape of the eye curve near its "cutoff" limits.

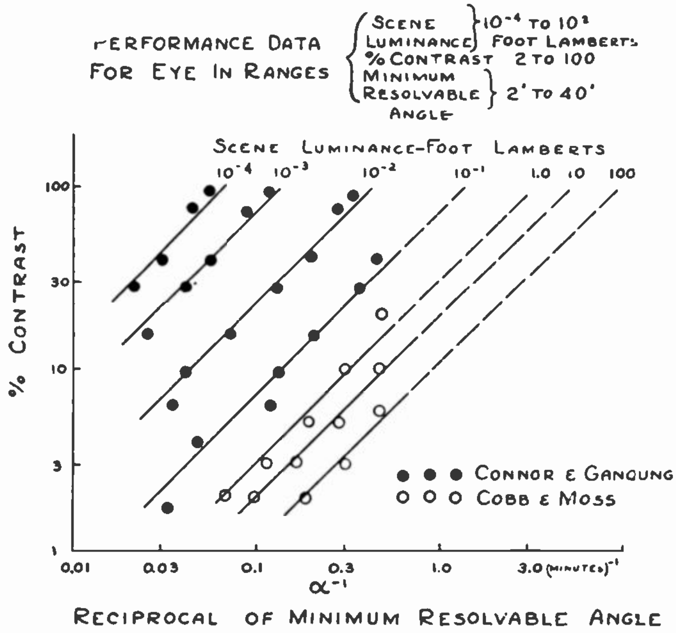

Fig. 4—Comparison of experimentally observed performance of the eye with ideal performance.

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[12] Connor, J. P., and Ganoung, R. E., "An Experimental Determination of Visual Thresholds at Low Values of Illumination", *Jour. Opt. Soc. Amer.*, 25, 9 (Sept. 1935), p. 287; Cobb, P. W., and Moss, F. K., "The Four Variables of Visual Threshold", *Jour. Frank. Inst.*, 205, 6 (June 1928), p. 831.

The complete characteristic equation for the eye, from Equations (10) and (4), is

$$B = 1.4 \frac{f^2k^2}{\alpha^2C^2t\theta} \times 10^{-2} \text{ ft-L} \tag{11}$$

where α is the angle in minutes of arc subtended by a picture element at the eye and k is the undetermined constant relating C and R. Figure 5 is a replot of the data in Figure 4. It is a more complete test of the characteristic Equation (11) and shows the small range[*] of the factor $k^2/t\theta$ from very low to very high lights as well as its actual value. At low lights the value of $k^2/t\theta$ is 2800. If one takes the exposure time t to be 0.2 sec, $k^2/\theta = 560$. It is known[13] that at threshold about 150 quanta (near 5300 Å) must be *incident* on the eye to generate a sensation. This corresponds to about 500 quanta if white light is

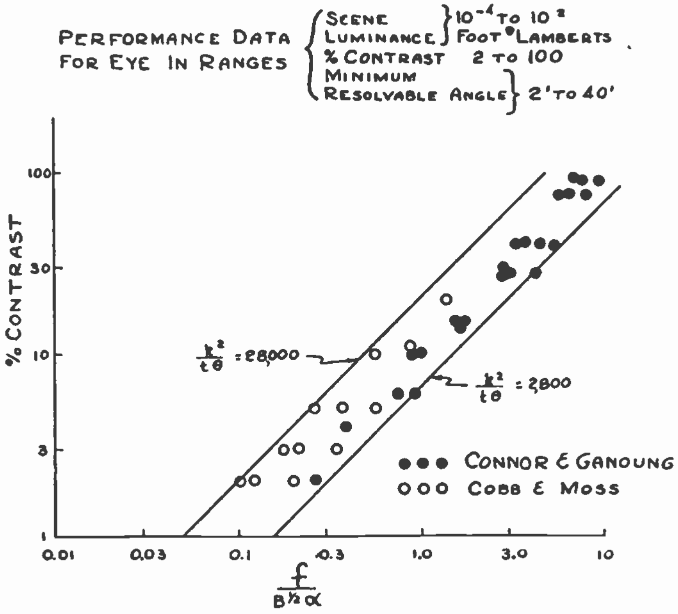

Fig. 5—Replot of data in Fig. 4.

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[*] If the full range of this factor is ascribed to the variation of quantum efficiency from low to high lights, one is presented in this approach with at most a ten-to-one variation in sensitivity of the eye from low to high lights as opposed to the usual statement that the dark adapted eye is 10^3 to 10^4 times as sensitive as the light adapted eye.

used. Various measurements and computations[13] of the number of quanta actually *used* in generating the sensation vary from 1 to 50, giving θ the range from 0.002 to 0.1 and k the range from 1 to 7. This range of k is to be compared with the independently obtained values of five from Romer and three from Schade.

All of the above discussion has been for the purpose of showing that the performance of the eye satisfies the same type of equation as that obtained for the ideal pickup device. The quantum efficiency, assuming $k = 5$, is about 5 per cent at low lights and about 0.5 per cent at high lights. The noiseless amplifier to be compared with Figure 1 may be some catalytic or triggering action induced by the absorption of quanta in the retina.

GENERAL DISCUSSION

The classes of picture pickup problems that have received frequent attention are:

(1) Specification of the performance of any one pickup device;
(2) Comparisons of the performance of two pickup devices of the same kind, or of different kinds;
(3) The setting of standards of performance for pickup devices that would "satisfy" the human eye.

The particular problems to be discussed here are intended only to be representative, rather than exhaustive.

SENSITIVITY

The simplest test for the relative "sensitivities" of two devices is accomplished by observing the lowest scene brightnesses at which they can still record a picture. This type of test is immediately subject to the questions: Was the exposure time the same for the two devices? What were the relative lens speeds used? What were the relative picture sizes? While these are obvious questions, there is no essential

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[13] References to number of quanta used by the eye for a threshold sensation:

1 quantum—DeVries, H., "The Quantum Character of Light and Its Bearing on Threshold of Vision, Differential Sensitivity and Visual Acuity of the Eye", *Physica*, 10, 7 (July 1943), p. 553.

2 quanta—van der Veldon, H. A., "The Number of Quanta Necessary for a Light Sensation for the Human Eye", *Physica*, 11, 3 (Mar. 1944), p. 179.

4 quanta—Hecht, S., "Quantum Relations of Vision", *Jour. Opt. Soc. Amer.*, 32, 1 (Jan. 1942), p. 42.

25 to 50 quanta—Brumberg, E. M., Vavilov, S. I., and Sverdlov, Z. M., "Visual Measurements of Quantum Fluctuations", *Jour. Phys.* (Russian), 7, 1 (1943), p. 1.

reason to pause here. The further questions of relative angles of view, numbers of picture elements and signal-to-noise ratios are of equal importance. In brief, the comparison of the sensitivities of two devices is not meaningful until the devices and their transmitted pictures are completely specified. But complete specification, as pointed out earlier, means that the quantum efficiency of the primary photo process is the only parameter that can vary the range of performance of an ideal device. And accordingly, the quantum efficiency is the measure of sensitivity. Not all devices, however, are ideal. For this reason, a more general figure of merit, based on Equation (4) is here proposed. The figure of merit is proportional to the reciprocal of the total light flux required to produce a picture of specified signal-to-noise ratio and resolution in a specified exposure time. The figure of merit is

$$\frac{f^2}{BA}$$

where f is the numerical aperture of the lens, B is the scene brightness, and A the area of target. If the performance of the device is ideal, the figure of merit becomes also a measure of its quantum efficiency.

It is recognized that the signal-to-noise ratio of a given picture is not a readily accessible parameter and that there is no general agreement on a measure of resolution. The evaluation of sensitivity, however, can be no more accurate than the knowledge of these parameters. It is of interest to apply the figure of merit to the interpretation of several familiar problems.

FILM SPEEDS

Consider the range of film speeds advertised. For the most part. these are films of the same quantum efficiency but different grain size* and, for the most part, the essential sensitivity performance of these films is the same. A simple example will make this clear. Two films, A and B, are rated at the relative speeds of one and four, respectively. Their quantum efficiencies are equal and the average grain area of B is four times that of A. Normally, one might say that B can pick up a scene with one-fourth the light required by A. While this statement is true, it is misleading. Suppose one wants the same resolution and depth of focus in both pictures. This would mean a film area of B

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* The relative speeds of Super XX and Eastman High Resolution plates are in the ratio of about 10^4 to 1. The relative grain areas are in the ratio of about 10^3 to 1.

four times as large as A to match resolutions and consequently a lens for B stopped to twice the numerical aperture (f/number) of the lens for A to match depth of focus. The result is that both films require the same scene brightness to transmit the same picture—a result which could have been anticipated from their equal quantum efficiencies or from their figures of merit evaluated for the same transmitted picture quality.

COMPARISON OF EYE AND FILM

An interesting application of the figure of merit is to the taking and viewing of motion pictures. For obvious reasons, the quality of the motion picture (signal-to-noise ratio and resolution) is aimed at equaling or exceeding the quality of picture which the eye can transmit at the brightness of the motion picture screen. For equal quality one can anticipate that the figure of merit for the eye would be at least a factor of twenty better than for film based on relative quantum efficiencies. But, in so far as film aims at better quality and attempts to compensate for some of its limitations by projecting pictures at a higher than unity gamma, an additional factor can be expected in favor of the eye.

Table 1 gives approximate values for f, B, and A to be associated with the camera that takes the pictures and the eye that views them. The area of target used for the eye is that area of retina occupied by the motion picture at a 4:1 viewing distance. The figure of merit for the eye is seen to be 250 times that for film.

Table 1

		B, Ft-L	A, In.^{2}	$\frac{f^2}{BA}$
Eye	2.5	10	0.03	20
Film	2	100	0.5	0.08

COMPARISON OF FILM AND TELEVISION PICKUP TUBES

Figure 6 shows the setup for comparing the low light performance of Super XX film and an image orthicon. 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-mm camera photographed simultaneously the original and reproduced pictures. Both cameras used $f/2$ lenses and an exposure time of

1/30 sec. Figure 7 shows the results. Only in the first exposure, at 2 foot-lamberts brightness of the subject, do both original and reproduced pictures appear. At 0.2 ft-L 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 ft-L which is the brightness of a white surface in full moonlight.

One interpretation of this test is that the image orthicon is 50 times as "sensitive" as Super XX film because it can transmit a picture with 1/50th of the light required by the film. The present paper argues against this interpretation and sets the factor at ten. This is based on the fact that the area of target (photo-cathode) used by the image orthicon was five times the area of the 35-mm film frame.

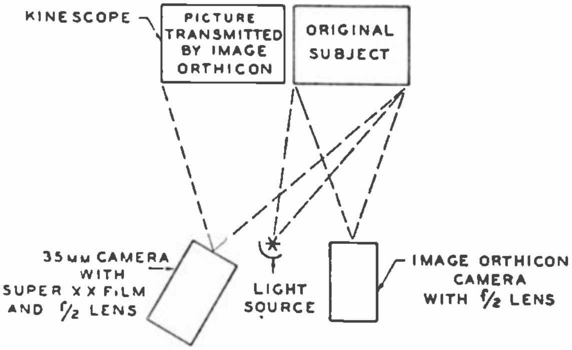

Fig. 6—Setup for comparison of low light performance of Super XX film and an image orthicon.

If the cameras were to be set up to transmit the same picture with the same angle of view and depth of focus, the lens on the image orthicon would have to be stopped $5^{1/2}$ times the numerical aperture of the lens for the 35-mm camera. The threshold scene brightnesses would then be in the ratio of 10:1.

GRAININESS AND SIGNAL-TO-NOISE RATIO

An excellent survey of the extensive history of the problem of specifying the graininess of film has recently appeared by Jones and Higgins.[8] In this paper and in a second one[14] they undertake to compare two general methods of measuring graininess. Method I, which they describe as a psychophysical measurement, records the distance

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[14] Jones, L. A., and Higgins, G. C., "Photographic Granularity and Graininess", *Jour. Opt. Soc. Amer.*, 36, 4 (Apr. 1946), p. 203.

from the observer at which the grainy film appears to blend into a uniform surface. (After introducing an observer for his special virtues as a measuring instrument, he is ushered part way out again by the device of normalizing his results with a standard test chart.) Method II is an objective measurement of the transmission or density fluctuations of the film using scanning apertures of various sizes. Broadly, Jones and Higgins argue (1) that the objective measurements should match the "blending distance" measurements in order to be considered valid; (2) that the two types of measurements do not match; and (3) that a major discrepancy is that the blending distance measurements tend to *decrease* at large densities while the objective measurements tend to *increase*.

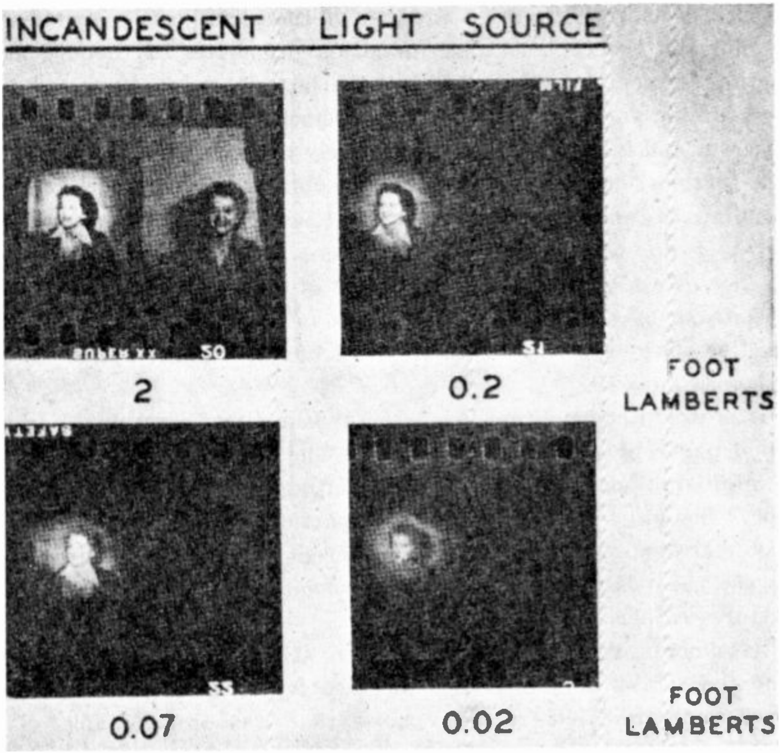

Fig. 7—Comparison of low light performance of Super XX film and an image orthicon. (Image orthicon picture is on the left of each frame.)

In contrast to the above, the present paper would argue that the two types of measurement, I (by the eye) and II (by a scanning aperture), should, so far as the eye and film satisfy the same physical equations derived for an ideal device, show good* agreement. A large part of the discrepancy noted above under (3) is removable when reference is made to Figure 4. Here it is seen that in the range of 0.1 to 10 ft-L the discrimination of the eye for small contrast differences varies by about five to one. This would correspond to a five-to-one ratio of blending distances for the same film viewed at a brightness of 10 and at a brightness of 0.1 ft-L. Because the visual observations of blending distance are made with a fixed source brightness attenuated by films of varying density, the resulting blending distance measurements are a product of the graininess properties of the film and the contrast discrimination properties of the eye as a function of scene brightness. When the latter term is separated out, the graininess versus density measurements by the two methods (observer and instrument) show relatively good agreement.

A further rough confirmation may be obtained by reference to some "blending distance" measurements of Lowry[15] in which a *constant viewing brightness* was preserved. These showed about a factor of two *increase* in graininess for a variety of films in the range of densities from 0.2 to 1.0. This increase is in good agreement with the objective (large aperture) measurements of Jones and Higgins[14] shown in their Figure 16.

It is worth commenting briefly on another item emphasized in the second paper by Jones and Higgins.[14] The concept of the "effective scanning area" used by the eye in evaluating graininess is introduced. This is thought to be a useful concept particularly because the results of objective measurements, using different scanning aperture sizes, suggest the possibility of matching visual observations with small apertures rather than large apertures.

Arguments, similar to the above, were at one time current in evaluating the "noise" in a television picture. It was often remarked that it was only the high-frequency noise that was objectionable. This would correspond, for example, to selectively emphasizing the observations of graininess of film obtained with small scanning apertures (either retinal or instrumental). It is a relatively simple experiment in a television system to increase the effective scanning aperture

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* A precise correlation between eye and instrument observations must, of course, take into account the detailed performance of eye and film near their limiting resolution—performance which both for eye and film is determined more by the finite size of its mosaic elements than by statistical fluctuations. The significance attached to precise visual observations, however, should be tempered by the known large spread of eye characteristics from individual to individual.

[15] Lowry, E. M., "An Instrument for the Measurement of Graininess of Photographic Materials", *Jour. Opt. Soc. Amer.*, 26, 1 (Jan. 1936), p. 65.

several fold, either by reduction of pass band or by defocusing the kinescope spot. Such aperture changes are accompanied by large changes in total noise power as viewed on the kinescope. Yet the effect on visibility of noise of cutting out the high-frequency noise components is small compared with the same changes in noise power distributed uniformly over the noise spectrum. This latter statement is borne out by the curves in Figure 9. In brief, the visibility, or annoyance, of noise must be assessed over the full range of picture element sizes from elements at the limiting resolution of the eye to the largest element, which is the picture itself considered as a unit.

RESOLUTION

The most frequently used, because it is the most easily observed, specification of resolution is the finest detail that a system can resolve. This is true for film, pickup tubes, the eye, and optical lenses. In general, this specification is satisfactory if it is appreciated that the limiting resolution itself has only narrow utility and that the limiting resolution is more an indirect measure of what detail is *well resolved* by the system. The "well-resolved" detail may be two to four times coarser than the finest detail. And in the judgment of picture quality, the eye attaches little weight to the picture elements in the neighborhood of limiting resolution.

One illustration of the confusion caused by the use of limiting resolution is the comparison frequently made between the resolution of motion picture film and of a television system. The *limiting resolution* of film is compared with the *"cutoff" resolution* of a television picture. The picture detail at the "cutoff" resolution of a television system, however, as limited by the amplifier pass band, has at least the possibility of being clearly resolved. It is misleading to attach the same weight to this type of resolution as is attached to the *limiting resolution* of film. It would be nearer a true evaluation if the resolution of film were specified at that number of lines at which film matched the signal-to-noise ratio of a television system at its "cutoff." Such a comparison would place the resolution of 35-mm motion picture film, normally quoted at a limiting resolution of 1000 to 2000 lines, nearer to the resolution of a 500-line than a 1000-line television picture.

In general, the specification of the signal-to-noise ratio that a picture pickup device can transmit at an intermediate resolution is a more accurate and significant specification, not only of resolution, but also at the same time of the half-tone discrimination of the device, than is the specification of limiting resolution.

SATISFYING THE HUMAN EYE

Only one problem, that of presentation brightness, will be discussed here. Figure 8 shows the signal-to-noise ratio curves of a picture taken at scene brightness B_1, and viewed by the eye at presentation brightness B_1'. The viewed picture is assumed to be "noise free" and accordingly the B_1 curve lies above the B_1' curve. If, now, the presentation brightness is increased to B_2', the original scene brightness must also be increased, other things being constant, by the same factor to B_2' in order to match the increased discrimination of the eye. These considerations are significant because both motion pictures and television pictures aim at higher presentation or screen brightness.

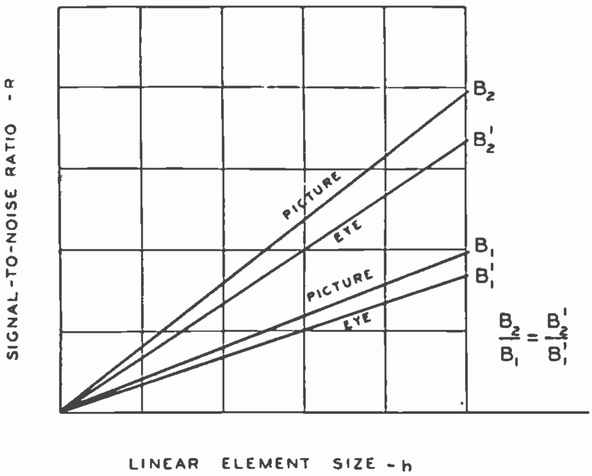

Fig. 8—Dependence of scene brightness on reproduction brightness.

The converse of the above operations makes an interesting test. Given a grainy motion picture or a "noisy" television picture, the most effective way of eliminating the fluctuations with the least cost to picture detail is to interpose a neutral filter between the eye and the picture. The discrimination of the eye is thereby readily reduced below the fluctuation limits of the picture. At the same time, the picture is shifted to a portion of the eye characteristic which shows higher apparent contrast and thus partially compensates for the loss of brightness. Figure 9 shows schematically the effect on picture detail of three ways of trying to eliminate "noise": reduction of picture brightness, increase in viewing distance and reduction of bandwidth of the picture. The last-named operation is peculiar to a television system and, while it reduces the total noise in the system, has little effect on the visibility of noise until an extremely coarse picture is obtained.

The curves B_1 are the signal-to-noise ratio characteristic of the picture. The curves B_1', B_1'', B_2' are the signal-to-noise ratio characteristics of the eye at the brightnesses B_1' and B_2'. In order that the fluctuations in the picture not be observed by the eye, the signal-to-noise ratio of the picture should be above B_1', B_1'', or B_2'. The limits

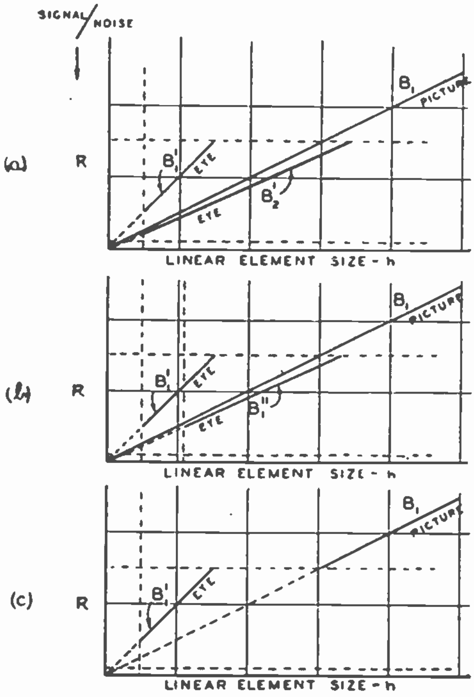

Fig. 9—(a) Noise reduction by lowered reproduction brightness; (b) Noise reduction by increased observer distance; (c) Noise reduction by bandwidth reduction.

of performance of the eye are shown by the three dotted lines. They mark out the minimum area that the eye can resolve by virtue of its cone structure, the minimum signal-to-noise ratio that it can perceive, and the maximum signal-to-noise ratio it can generate corresponding to the Weber-Fechner limit of 2 per cent brightness discrimination. The "cutoff" characteristics of the eye are shown as idealized sharp breaks to simplify the argument.

Starting with a noisy picture, that is, B_1' lying above B_1 as in Figure 9a, there are several formal operations that can be performed to get rid of the noise, that is, to insure that *all parts* of B_1 lie above the eye curve. Each of these operations corresponds to a physical operation and each affects the observed picture detail differently. In Figure 9a the eye curve B_1' is transformed into B_2' by a change of ordinate scale factor. This corresponds to interposing a neutral filter at the eye. The finest detail observable is still at the "cutoff" limit of the eye. In Figure 9b, the eye curve B_1' is transformed into B_1'' by a change of abscissa scale factor. This corresponds to backing away from the picture. Although the finest observable detail remains at the "cutoff" of the eye, this "cutoff" now corresponds to coarser detail in the picture. In Figure 9c, the pass band of the amplifier through which the original picture is transmitted is reduced to the point where the picture fluctuations are below the Weber-Fechner limit. This is an expensive way to remove noise—expensive in picture detail.

A final aspect of the significance of presentation brightness arises in the comparison of the low light performance of a man-made device with that of the human eye. Assume, for example, that the man-made device is as sensitive as the eye. If one picks up a scene whose brightness is 0.1 ft-L and views the reproduction at a presentation brightness of 10 ft-L, noise should be visible in the reproduction while it was not visible in the original scene. The higher presentation brightness gives the eye an unfair advantage. A more valid procedure would match the presentation brightness of the reproduction with the brightness of the original scene.

VISUAL NOISE

The phrase "signal-to-noise ratio of the eye" has been used frequently in the preceding discussion. One might expect to be able to "see" these fluctuations just as one sees the graininess of film or the noise in a television picture. The writer is convinced that such fluctuations are observable* particularly at low lights around 10^{-4} ft-L. A white surface then takes on a grainy appearance not unlike that of motion picture film. The observations in more detail are: in complete darkness little or no fluctuations are detectable, a fact which attests the substantial absence of local noise sources in the eye. Near threshold brightnesses, large area, low amplitude fluctuations appear. At higher brightnesses these fluctuations increase in amplitude and decrease in size. In the neighborhood of 10^{-2} ft-L the fluctuations tend to dis-

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* See also DeVries.

appear and a white surface takes on a "smooth" appearance and remains so at normal brightness levels. A secondary observation is that low-level blue light appears distinctly more grainy than low-level red light.

The last observation, together with known data on dark adaptation, fits in well with the assumption of a gain control mechanism in the eye. This gain control, just as the gain control in a television receiver or the lamp brightness used for film projection, does not alter the signal-to-noise ratio but does alter the visibility of noise by presenting the picture at a higher or lower brightness level. Thus, at high scene brightnesses, the gain control in the eye may be turned down to the point where the fluctuations are just not visible. (The sensitivity of the eye is apparently high enough to afford this luxury.) If one suddenly reduces the scene brightness, the gain control is still momentarily set at a low value and the picture is dim or not visible. As the gain control resets itself at a higher level, the picture appears to get brighter. This corresponds with the experience of dark adaptation. At these low light levels (10^{-1} ft-L) one has only to assume that the gain control is set high enough to make the fluctuations visible.

To account for the observations that low-level blue light appears to have more fluctuations than low-level red light, the gain control mechanism can be assumed to be set higher for blue than for red. This is not as "*ad hoc*" as it may appear. The reason is that, although at low-light levels* blue appears brighter and grainier than red, they both present the same resolution to the eye.[16] And since the resolution is determined by signal-to-noise ratio, this is in agreement with the assumption of a gain control that varies presentation brightness but not signal-to-noise ratio.

ACKNOWLEDGMENTS

The writer would like to acknowledge, without committing the acknowledged to the conclusions presented above, his indebtedness to Dr. D. O. North and Dr. G. A. Morton of these laboratories, and O. H. Schade of the RCA Victor Division for many profitable discussions of the subject of this paper.

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* The test is performed by starting with red and blue at the same brightness at high-light levels and attenuating both by the same neutral filter.

[16] Luckiesh, M., and Taylor, A. H., "A Summary of Researches in Seeing at Low Brightness Levels", *Illum. Eng.*, 38, 4 (Apr. 1943), p. 189.

ANALYSIS, SYNTHESIS, AND EVALUATION OF THE TRANSIENT RESPONSE OF TELEVISION APPARATUS*†

BY

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Summary—The sharpness of detail in a television picture is directly dependent upon the capability of the transmitter for the transmission of abrupt changes in picture half tone. A suitable test signal is a square wave of sufficiently long period.

Rules are deduced for the evaluation of the subjective sharpness to be expected in transmitted pictures and may be applied when the square-wave response of the transmitting apparatus is known.

Rapid chart methods have been devised for (1) the analysis of a square-wave output into sine-wave amplitude and phase response and (2) the synthesis of a square-wave response from a given set of amplitude and phase characteristics. Analysis furnishes an immediate solution to the familiar but troublesome problem of finding the sine-wave characteristics of television apparatus.

The four aspects of the application of square waves to television, i.e., measurement, analysis, synthesis, and evaluation, are presented as a basis for a unified and complete technique.

The authors hope that this paper will be a contribution to the general problem of working out electrical specifications for television transmitters and other television apparatus, giving information regarding the steepness of rise and the amplitude of overswing of the square-wave response.

I. INTRODUCTION

AS a result of the scanning process, the sharpness of detail in a television picture is directly dependent on the capability of the transmitter and receiver for the faithful transmission of signals arising from abrupt changes in picture half tone along the scanning line. Recognition of the validity of the Heaviside unit voltage, the electrical equivalent of an abrupt change in half tone, as a test signal, was accorded early in the art. Notwithstanding, the preponderance of emphasis has been placed upon the sine-wave characteristics of television apparatus, that is, upon the amplitude- and phase-versus-frequency characteristics.

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It has long been known that the response of a linear signal-transmitting system to a Heaviside unit voltage contains all the information necessary to determine both the phase-frequency and the amplitude-frequency characteristics. Conversely, the two frequency characteristics determine uniquely the response to a unit voltage (and incidentally the response to any other transient input wave). It is also known that the response to a single abrupt rise in a *repeating* square wave of sufficiently long period is essentially the same as the response to a Heaviside unit voltage insofar as that part of the transient response due to high frequencies is concerned.

In view of this implicit relationship between the sine-wave and the transient-response[1] characteristics of electrical circuits, it is surprising that the testing and specifying of high-frequency fidelity of both audio and video apparatus by the response to a square-wave has not become more common.[2]

Several circumstances have impeded rapid growth in the use of square waves to determine the high-frequency response. In the first place, the well-established sine-wave methods, which were developed for audio work, have the advantages of precedence and well-developed techniques for measurement, recording and plotting data, diagnosing imperfection, and comparing performances. Second, the lack of suitable oscillographic apparatus for accurately indicating the instantaneous response as a function of time has been a large contributing factor. A recently developed square-wave oscillograph[3] and square-wave generator provide a solution to this problem.

The laboriousness of classical methods for translating the results of sine-wave measurements into transient response and vice-versa, has tended improperly to make the two test methods seem unrelated and competitive instead of complementary. Furthermore, there has been no satisfactory means for evaluating *numerically* the fidelity of a par-

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[1] In this paper, the term *transient response* will be used as the equivalent of the expression *response to a Heaviside unit voltage*.

[2] The use of a 60-cycle square-wave generator and an oscillograph for investigation of the behavior of a television system at *low frequencies* of the order of the field scanning rate is well established. In these measurements, performance is judged by inspection of the "tilted" output wave and harmonic analysis of the wave is usually not desired. In general, a television system will have uniform response over a frequency range of many octaves in the region between the so-called "low-frequency" end and the "high-frequency" end, so that fidelity measurements at the two ends of the spectrum may be considered separately. This paper will be concerned only with the high-frequency end.

[3] R. D. Bell, A. V. Bedford, and H. N. Kozanowski, "A Portable High-Frequency Square-Wave Oscillograph for Television", *Proc. I.R.E.*, October, 1942.

ticular piece of apparatus or transmitting system from the transient response. In television applications it has been recognized that the "mean" steepness of rise of the transient-response wave is a measure of the sharpness of the picture which the system could transmit. Still there is no general agreement on a method of measurement and calculation of a mean value of steepness which is a consistent and an accurate indication of the picture sharpness for a wave having nonuniform steepness during the time of rise. The presence of overswing in the transient response makes evaluation even more difficult because of the effect on the visual sharpness of the picture and because of the introduction of spurious effects in the picture.

Accordingly, for the purpose of simplifying the passage between sine-wave response and transient response and interpreting the latter, we present below: (1) a graphical chart method for analyzing the response of a system to a square-wave input signal to obtain the sine-wave phase and amplitude characteristics; (2) a graphical chart method for synthesizing the response to a square wave from the sine-wave phase and amplitude characteristics; (3) a method for evaluating the mean steepness of a transient-response wave in terms of the *width of blur* produced in a television image by a wave which is similar in its visual effect and which has a *linear* change from one level to another; and (4) suggestions for the supplementing of sine-wave measurements by transient measurements.

II. ANALYSIS OF SQUARE-WAVE RESPONSE INTO PHASE-FREQUENCY AND AMPLITUDE-FREQUENCY CHARACTERISTICS

The analysis employs several permanent charts (Figures 2 to 5), the construction and use of which may be simply explained by reference to Figure 1.

Figure 1(a) shows the repeating square wave *e* applied to the apparatus under test. Wave *d* is the response measured at the output terminals. The fundamental period of *e* is not critical but must be taken long enough to insure that wave *d* has subsided to a substantially constant level during the latter part of each half cycle. Also the true time relation between waves *d* and *e* under test conditions has no significance in the present analysis and need not be known.

A basic hypothesis upon which the analysis of wave *d* rests **is** that a stepped wave *f* may be drawn which approximates in harmonic content that of wave *d*. It is clear from inspection that the waveform of *f* may be caused to approach that of *e* as closely as desired by taking the steps sufficiently small. Since wave *f* in the regions *a* to *b* and *a′*

to b' is of uniform amplitude, the lengths of the time intervals from a to b and a' to b' have no bearing on the shape of the transient portion of the waves. Hence, the fundamental period of wave f is unimportant provided that the value is great enough so that a part of f is uniform after each transition. The upward transition of wave f is the same as the downward transition except for inversion. Hence, the rectangles g'', h'', i'', etc., may be considered as continuations of rectangles g', h', i', etc. It follows that wave f has the components g, h, i, etc.

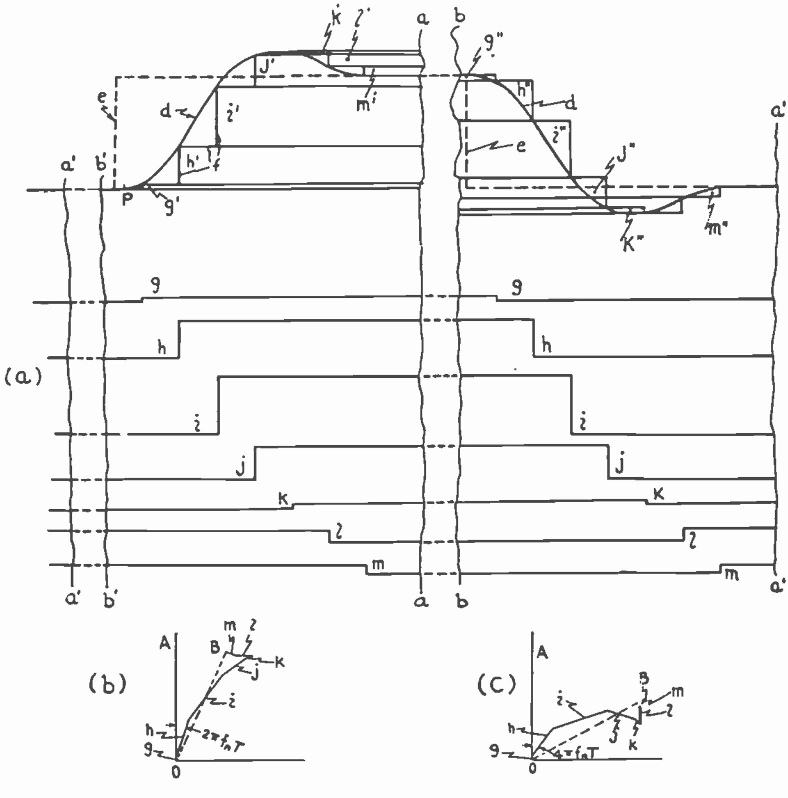

Fig. 1—(a) Approximation of a square-wave response d by a stepped wave f.
(b) Vector addition of components of frequency f_n in waves g, h, i, etc.
(c) Vector addition of components of frequency $2f_n$ in waves g, h, i, etc.

Each of these square-wave components is identical in shape to the input wave e and hence contains all harmonics in the same proportion as the input wave. Each of the square waves, however, has a different delay and hence, the harmonics of the various square waves occur in different phase relations. Therefore, the magnitude of each harmonic in the stepped wave, relative to the same harmonic in the input wave *e*, is indicated by the sum of a group of vectors having amplitudes proportional to the steps *g'*, *h'*, *i'*, etc., and angular positions corresponding to the different delays of the square waves *g*, *h*, *i*, etc. No cognizance need be taken of the fact that the harmonics of a square wave vary in amplitude inversely as their frequencies. Inasmuch as the fundamental frequency of the input wave may be allowed to approach zero (such that wave *e* becomes a Heaviside unit voltage), any reference to discrete harmonics may be dropped and the response of the apparatus approximated at any frequency to a degree of accuracy which depends on the fineness of the steps in wave *f*.

Figure 1(b) shows the vector addition involved in finding the response for the frequency f_n. Each vector component has the same angular position, namely, $2\pi f_n T$, with respect to the preceding component since the waves *g*, *h*, *i*, etc. correspond to points on the real response curve taken at equal time intervals *T*. Waves *l* and *m* are negative; hence, the vectors *l* and *m* are negative. The length of the vectorial sum *OB* gives the relative amplitude response of the apparatus tested and the angle *AOB* is the relative phase shift for the sine wave f_n. Figure 1(c) is drawn for a frequency equal to $2f_n$. The angle between successive vectors is $2\pi 2f_n T$. As in Figure 1(b), *OB* is the amplitude response at the frequency $2f_n$ and the angle *AOB* is the relative phase shift.

A set of analysis charts, Figures 2, 3, 4, and 5, were designed in order to reduce to a minimum the labor involved in performing the vectorial additions in Figures 1(b) and (c). Essentially, each chart serves as a protractor and a linear radial scale for use in locating the end points of the vectors. Two sets of time intervals, 1/20 and 1/30 microsecond, (referred to as 20- and 30-megacycle dots) between successive components *g*, *h*, *i*, etc., appear to be adequate for television applications. The basis for a choice of one of the two sets depends upon the degree of accuracy desired. This aspect is discussed later. Components (i.e. readings from the transient-response wave *d*) are numbered to correspond to radial lines on the charts. The angle between consecutively numbered radial lines is the angle used in the construction of Figures 1(b) and (c) (e.g., on Figure 3, the angle in scale 1 for the solution of the response at 1 megacycle is 18 degrees.) The component vectors *g*, *h*, *i*, etc., lie along the radial lines and the vectors are added as in Figure 1(b) by manipulation of a sheet of semitransparent paper. Detailed directions for the operation of the charts are given below. The charts and directions are drawn or printed preferably on cardboard or stiff paper.

A. Instructions for Using Figures 2, 3, 4 and 5 for Analysis of Square-Wave Response

Using a square-wave generator and an oscillograph or other means, obtain per cent voltage readings at 1/20-microsecond intervals (corresponding to oscilloscopic readings with 20-megacycle "dots" for timing) along the transient wave (Figure 6) such as shown in column (a) of Table I. (If more accurate analysis is desired, use readings at 1/30-microsecond intervals).

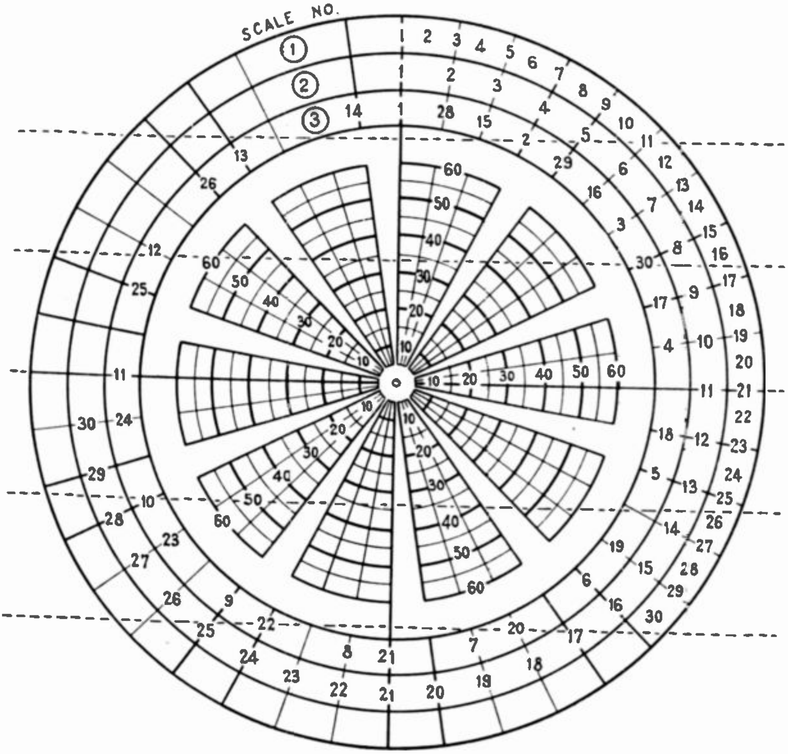

Fig. 2—Chart for square-wave analysis.

Scale	20-Mc Dots	30-Mc Dots
①	0.25 Mc	0.375 Mc
②	0.5 Mc	0.75 Mc
③	1.5 Mc	2.25 Mc

Readings should begin at the zero-voltage level before the beginning of the transient and end at a 100 per cent point where the voltage becomes uniform after the transient rise has been completed and substantial rest attained. Wave *d* of Figure 1 is a typical plot of such a wave but plotting is not necessary for the purpose of the analysis.

Compile column (b) by taking the differences in adjacent readings in column (a). These numbers represent the amplitudes of the increments in the "stairstep" wave *f*. Number the increments consecutively as in column (c).

Draw a horizontal *x* axis and a vertical *y* axis on a sheet of sufficiently transparent tissue paper. Place the tissue paper on Figure 4,

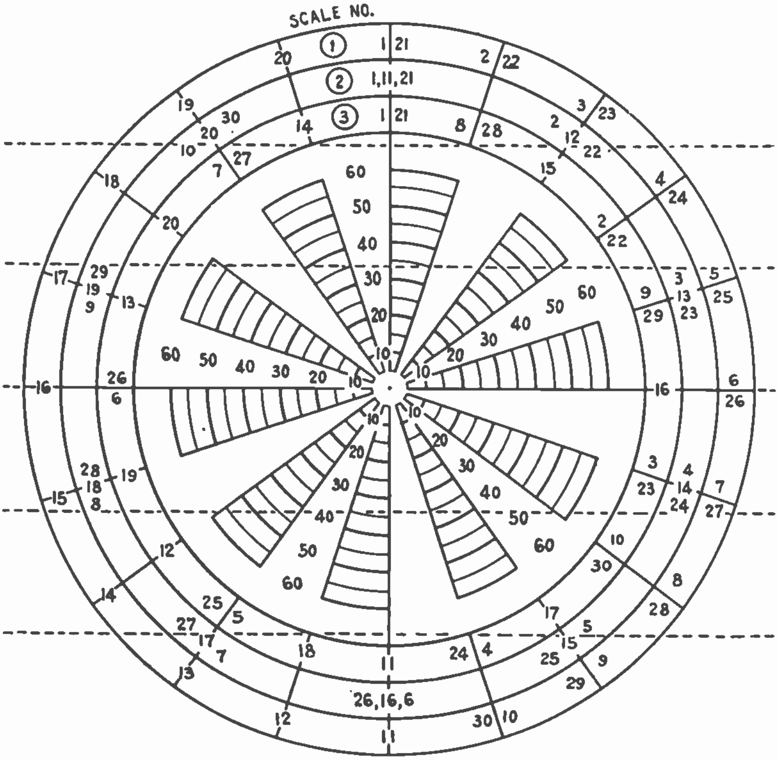

Fig. 3—Chart for square-wave analysis.

Scale	20-Mc Dots	30-Mc Dots
①	1 Mc	1.5 Mc
②	2 Mc	3 Mc
③	3 Mc	4.5 Mc

for example, with the origin of the *x-y* system at the center of the chart. (It will be noted that Figures 2, 3, and 4 are for different frequencies.) Make a pencil dot on the tissue on the radial line marked "1" in circular scale (1) at a distance from the center of the chart equal to increment No. 1 in column (b). Move the tissue so that this dot coincides with the center of the chart, keeping the *x* axis parallel to the horizontal dotted lines on the chart. Make a second dot on the radial line marked "2" in circular scale (1) at a radius equal to increment No. 2. Move the tissue until the second dot is at the center of the chart and repeat the procedure for the remaining increments. Positive increments are plotted in the radial direction *toward* the increment number in circular scale (1) while negative increments are plotted in the opposite direction. Dots may be numbered to avoid errors.

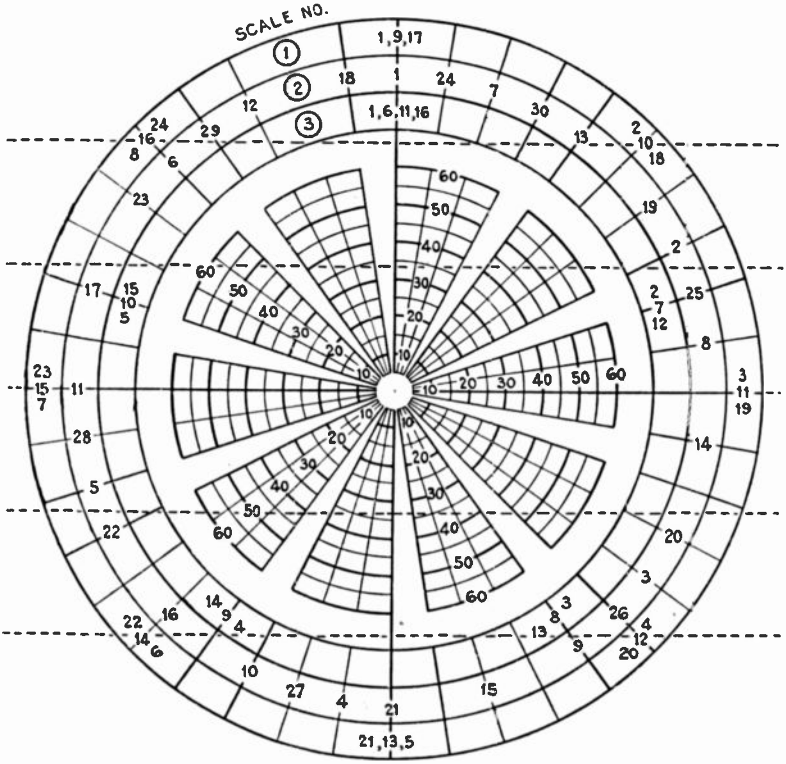

Fig. 4—Chart for square-wave analysis.

Scale	20-Mc Dots	30-Mc Dots
①	2.5 Mc	3.75 Mc
②	3.5 Mc	5.25 Mc
③	4.0 Mc	6.0 Mc

Draw a vector from the x-y origin to the final dot located. Place the tissue on Figure 5 and read the length of the vector on the "amplitude scale." This length is the 2.5-megacycle amplitude response in per cent. Read the angle between the y axis and the final vector by using the protractor in Figure 5. This angle[1] is the phase lag of the 2.5-megacycle component of wave (*d*) relative to point *P* in Figure 1 where *P* is ½ a reading interval before the second reading in column (a). The phase angles obtained for the various frequency components are correct relative to one another but the absolute time delay through the system is not obtained. The phase angles may be converted to time delay by means of the delay graph of Figure 5. In the example given above, the amplitude response at 2.5 megacycles is 85 per cent. The relative phase angle is 103 degrees which corresponds to a time delay

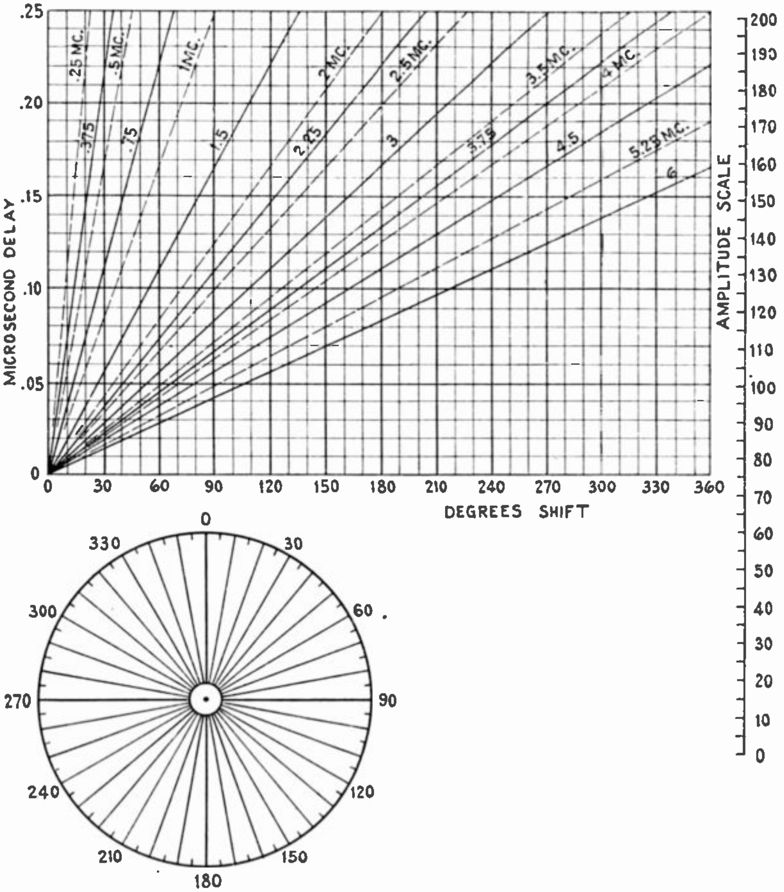

Fig. 5—Chart for square-wave analysis.

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[1] The angle read between the *y* axis and the vector is the phase angle for the 2.5-megacycle response of the stepped wave *f* with respect to the time of the second reading of column (a). Wave *f* lags the true-response wave *d* by one half a reading interval.

of 0.114 microsecond. Obtain the response at other frequencies in a similar manner by using the increments in conjunction with the other scales of Figure 2 and the scales of other charts.

The charts may be used for other frequency ranges such as the audio range. In such applications, the time interval between successive readings on the transient-response wave will not usually correspond to 20- or 30-megacycle dots. However, the analysis charts shown in Figures 2, 3, 4, and 5 may be employed as in the example given in Table I, when the appropriate multiplying factor for frequency is computed. This factor is equal to T_0/T_X where T_0 is $1/30 \times 10^{-6}$ or $1/20 \times 10^{-6}$ second depending upon the original dot frequency for which the charts were designed. T_X is the new time interval in seconds.

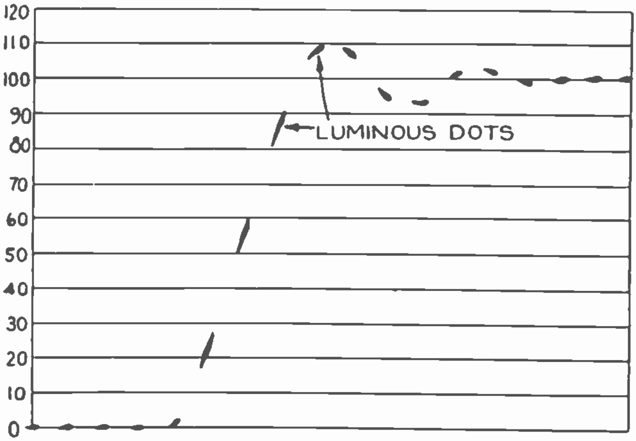

Fig. 6—Illustrative example of a square-wave response reproduced by an oscillograph showing readings at 0.05-microsecond intervals (20-megacycle dots).

B. *Examples of Accuracy of Chart Analysis in Specific Applications*

In the instance of a compensated resistance-coupled amplifier, mathematical formulas are available for the exact calculations of the amplitude and delay characteristics and the response to a unit function.[5] Hence, some conception of the accuracy of chart analysis may be gained by observing the agreement of data thus determined with the exact characteristics. Figure 7 contains such data based on the analysis of the unit-function-response wave shown in Figure 8. Figure 9 contains similar data based on the response wave of Figure 10. Since the delay characteristics determined by graphical analysis have only

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[5] Appendix II.

TABLE I

(a)	(b)	(c)	(a)	(b)	(c)
0	—	—	95	−13	7
2	2	1	93	− 2	8
22	20	2	101	8	9
56	34	3	102	1	10
87	31	4	99	− 3	11
108	21	5	100	1	12
108	0	6	100	0	13

relative significance, the data points corresponding to a 20-megacycle-dot readings were shifted a constant amount (0.026 microsecond) in Figures 7 and 9 so that the best correspondence with the mathematically determined delay curves was reached in order that the results of graphical and mathematical methods may be compared directly. A similar shift of 0.017 microsecond was made in the case of the 30-megacycle-dot data. The deviations of the data points in Figures 7 and 9 from the exact curves represent the degree of approximation of analysis as applied to two specific cases. For the general case, no definite limits can be set up for the error in sine-wave characteristics as determined by chart analysis. As in Fourier analysis,[6] the error

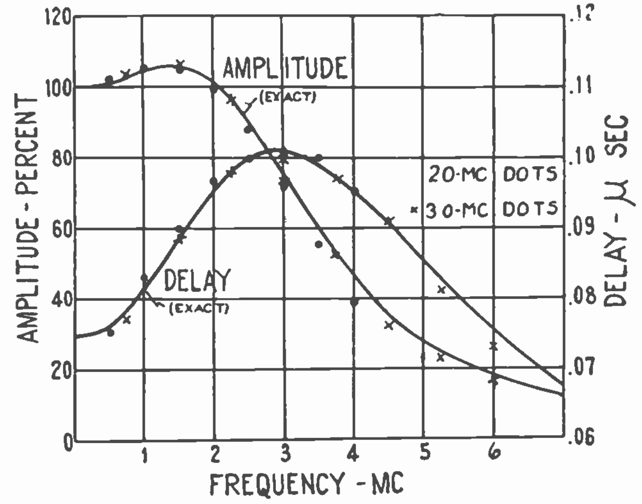

Fig. 7—Curves show exact amplitude and delay characteristics of a 2-stage compensated resistance-capacitance amplifier in which $K = \sqrt{2}$, $f_0 = 3$ megacycles. Amplitude and delay characteristics determined by chart analysis of Figure 8 are shown by data points.

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[6] The amplitude characteristic of a circuit, as determined graphically from the square-wave response, may be converted into the amplitude characteristic corresponding to Fourier analysis of the response of the same circuit to a square *pulse*. This conversion is discussed in Appendix III.

depends upon the specific curve which is analyzed and upon the length of the time interval between readings, or in other terms, upon how well the curve is defined by the series of dots. A better approximation to the exact amplitude or delay characteristic at the higher frequencies is afforded by a solution based on 30-megacycle-dot readings rather than on 20-megacycle-dot readings.

It is conceivable that as a consequence of strong components of very high frequency the transient-response wave could make violent excursions between adjacent readings such that a plot of the readings would not clearly define the wave even for the lower-frequency components. In such a case, inspection would show that analysis would be inaccurate. In general, it has been found that if the 20-megacycle timing dots trace out the square-wave response unmistakably, the 20-megacycle dots are sufficiently accurate for analysis out to 3.5 megacycles. Under the same conditions, 30-megacycle dots are adequate out to 5.25 megacycles.

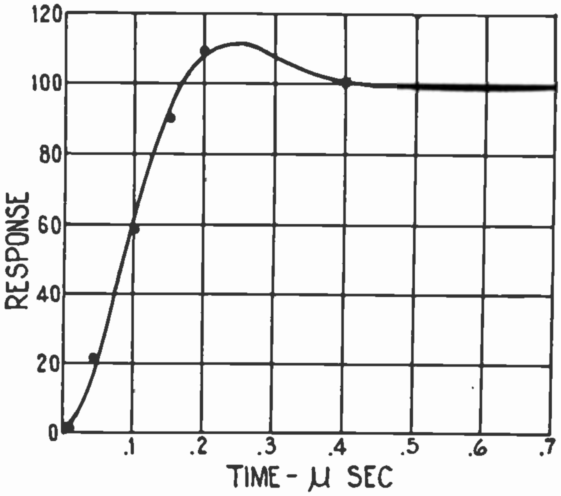

Fig. 8—Exact response to a unit function of a 2-stage compensated resistance-capacitance amplifier where $K = \sqrt{2}$ and $f_o = 3$ megacycles. Points shown were obtained by synthesis from theoretical amplitude and delay characteristics in Figure 7.

III. SYNTHESIS OF SQUARE-WAVE RESPONSE FROM THE SINE-WAVE CHARACTERISTICS

A closed mathematical formula[7] is available for the calculation of

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[7] A. V. Bedford and G. L. Fredendall, "Transient Response of Multistage Video-Frequency Amplifiers", *Proc. I.R.E.*, Vol. 27, pp. 277-285; April, 1939.

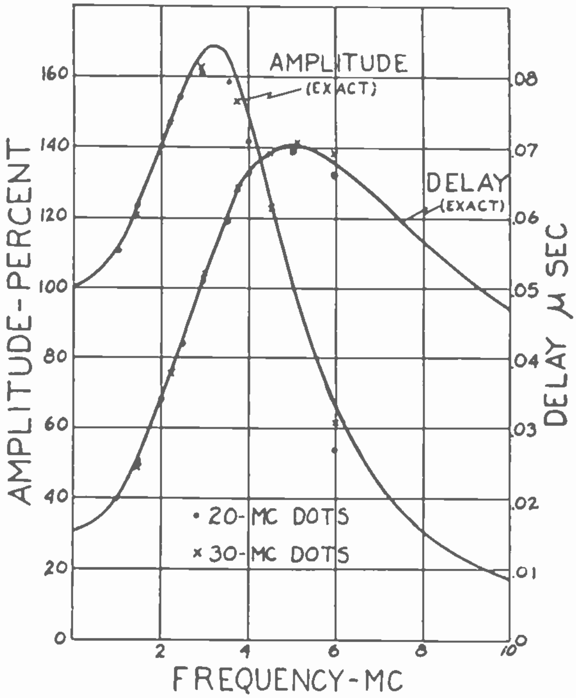

Fig. 9—Theoretical amplitude and delay characteristics of a 2-stage compensated resistance-capacitance amplifier in which $K = 1.1$ and $f_0 = 4$ megacycles. Amplitude and delay characteristics determined by chart analysis of Figure 10 are shown by data points.

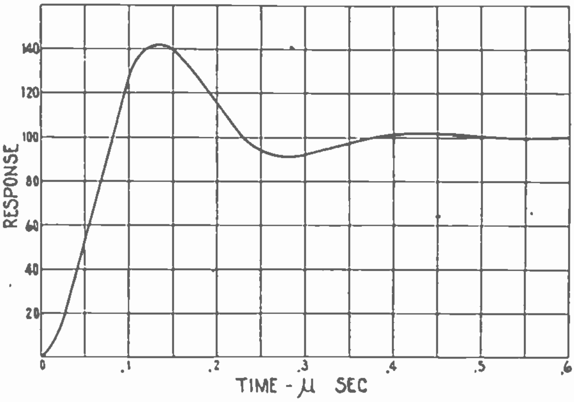

Fig. 10—Exact response to a unit function of a 2-stage compensated resistance-capacitance amplifier in which $K = 1.1$ and $f_0 = 4$ megacycles.

the response of a linear electrical system to a unit function from the sine-wave amplitude and phase characteristics. When a numerical answer is sought, however, the formula is rarely practicable. As pointed out earlier, the solution for the response to a square wave may usually be substituted when the period required does not necessitate the computation of many terms of the Fourier series. The period $1/f_p$ chosen must be long enough to permit the transient response of the circuit under consideration to assume a substantially constant value between consecutive abrupt changes in the square wave.

If the square wave applied to the circuit is

$$E(t) = \frac{1}{2} + \frac{2}{\pi}\,(\sin 2\pi f_p t + \tfrac{1}{3} \sin 6\pi f_p t + \tfrac{1}{5} \sin 10\pi f_p t + \cdots + \frac{1}{n} \sin 2\pi n f_p t + \cdots \qquad (1)$$

then the response is

$$e(t) = \frac{1}{2} + \frac{2}{\pi}\,(A_1 \sin 2\pi f_p (t - D_1) + \frac{A_3}{3} \sin 6\pi f_p (t - D_3) + \cdots + \frac{A_n}{n} \sin 2\pi n f_p (t - D_n) + \cdots \qquad (2)$$

where A_n and D_n are the amplitude response and delay of the circuit for the nth harmonic. $D_n =$ phase shift in radians$/2\pi n f_p$.

Synthesis charts, Figures 11, 12, and 13, were developed for the summing of the significant terms of the series (2) above. The principle of the charts is based on the vector representation of a sine function. For example, the term $A_n/n \sin 2\pi n f_p (t - D_n)$ is represented in Figure 14(a) by the length of the perpendicular AC to the vector OD. The value of the term may be found rapidly, for specific values of t which have been determined in advance, by dividing the circumference of a circle into a number of equal parts.

For example, if the circle in Figure 14(b) is divided into N equal segments, then the value of the sine term may be found every $1/Nnf_p$ second starting at $t = 0$. It is convenient to designate a specific value of time t by one of the whole numbers between 0 and N.

A straightedge may be placed along the radial line which makes the required angle $2\pi n f_p D_n$ with the x axis. (See Figure 14(b)). The quantity $2\pi n f_p D_n$ will usually be expressed in terms of the unit angular segment used in marking off the circumference of the circle. If a draftsman's triangle is moved along *OD*, successive values of the sine term corresponding to regular intervals of time may be read on a calibrated scale *AC* (marked on one leg of the triangle) by noting the points of intersection of the radial lines and an imaginary circle having a diameter *OA* equal to A_n/n. The multiplying factor $2/\pi$ in (2) may be taken into account by suitable calibration of the scale *AC*. A number of concentric circles permanently drawn on the chart, so as to divide the diameter of the largest circle into a convenient number of equal parts, will aid in finding the length *OA* as the triangle is shifted. Readings along the calibrated edge are positive above the line *OD* and negative below *OD*.

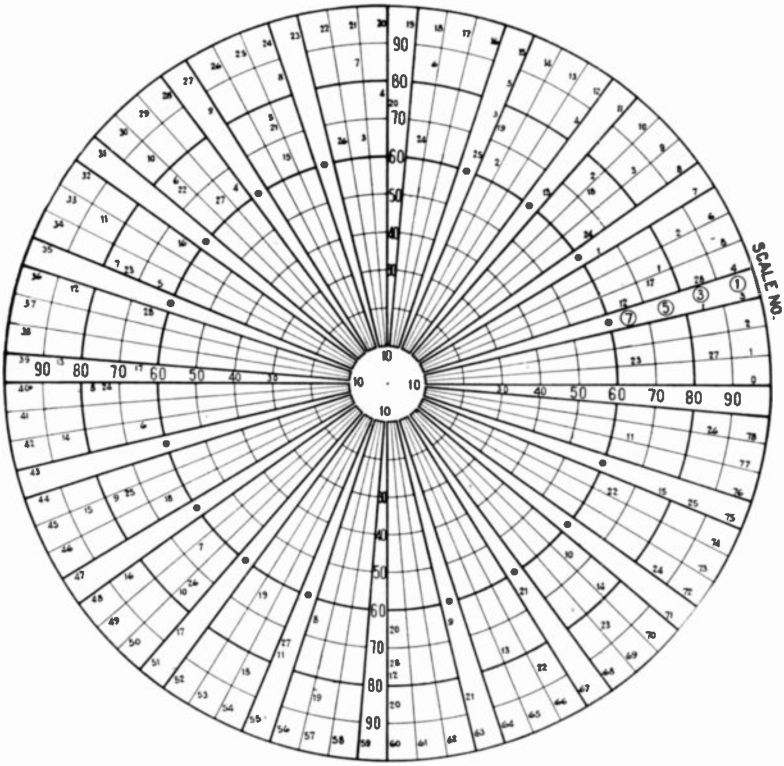

Fig. 11—Chart for square-wave synthesis.

Scale	0.05-microsecond intervals
①	0.25 Mc
③	0.75 Mc
⑤	1.25 Mc
⑦	1.75 Mc

The same chart may be used for a number of different harmonic terms by simply numbering the radial lines appropriately for each term. Thus, in Figure 14(b), if the circular scale (1) represents the fundamental term sin $2\pi f_p t$, then circular scale (3) will represent the 3rd-harmonic term sin $6\pi f_p t$ in which the unit angle is 3 times the unit angle in scale (1) etc. The steps involved in finding and summing the sine terms in (2) have been systematized in a set of synthesis charts, Figures 11, 12, and 13. The following directions have been drawn up to facilitate rapid use of the synthesis charts.

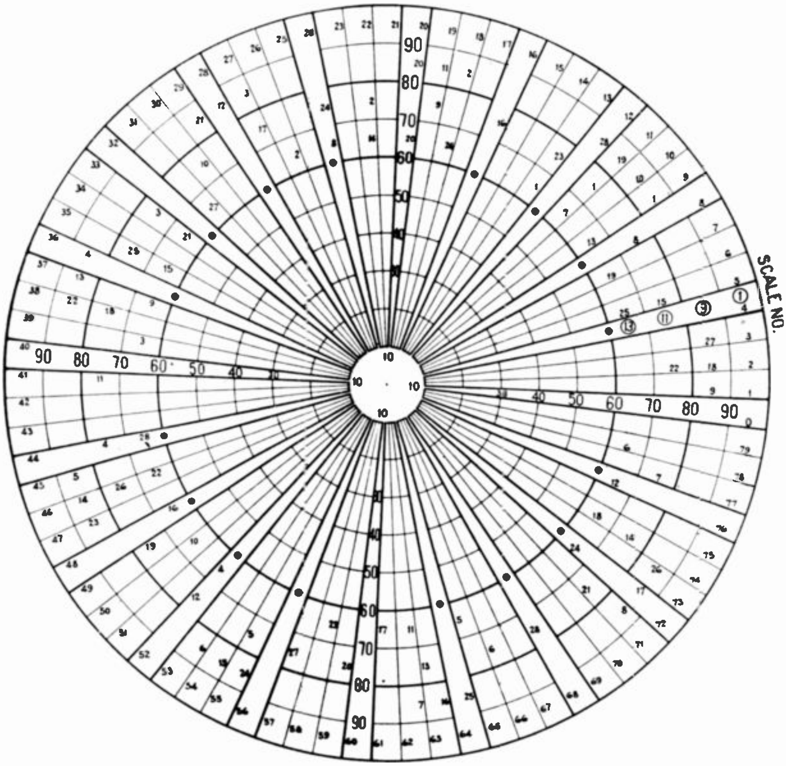

Fig. 12—Chart for square-wave synthesis.

Scale	0.05-microsecond intervals
①	0.25 Mc
⑨	2.25 Mc
⑪	2.75 Mc
⑬	3.75 Mc

A. Directions for Use of Synthesis Charts

Figures 11, 12, and 13 are used in the compilation of the response of an electric circuit to a square wave as represented by the first 10 odd harmonics (1st, 3rd, . . ., 19th) when the sine-wave, phase-delay, and amplitude characteristics of the circuit are known.

The particular charts illustrated were arbitrarily designed for a fundamental frequency of 0.25 megacycle which has been found satisfactory for television purposes. In this instance the square-wave response is found at intervals of 0.05-microsecond (i.e., 1/80 of the period). The result is a close approximation to the response of a circuit to a unit function when (1) the square-wave response attains substantially a steady value within 2 microseconds (i.e., one half the

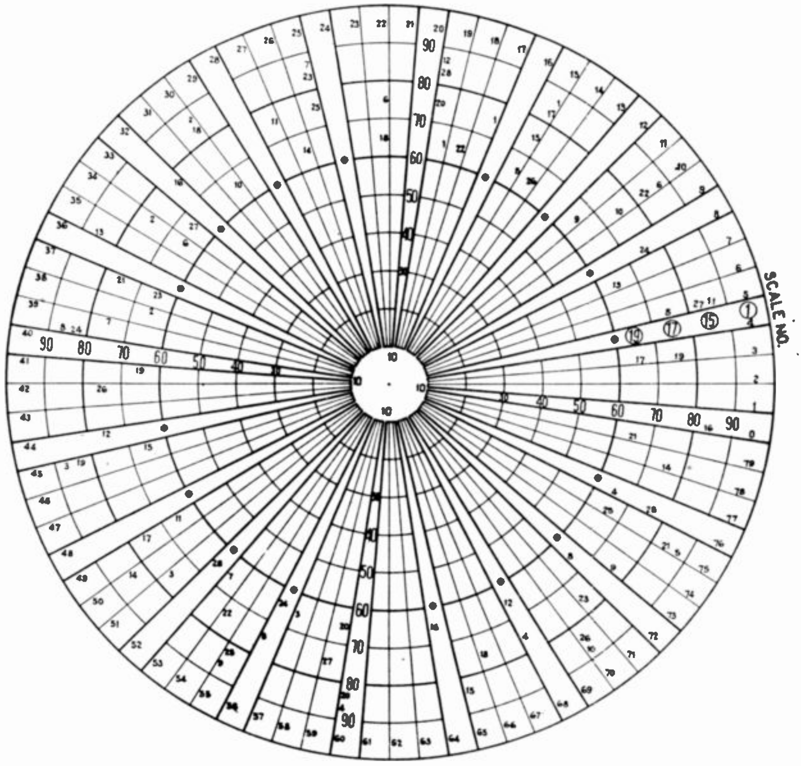

Fig. 13—Chart for square-wave synthesis.

Scale	0.05-microsecond intervals
①	0.25 Mc
⑮	3.75 Mc
⑰	4.25 Mc
⑲	4.75 Mc

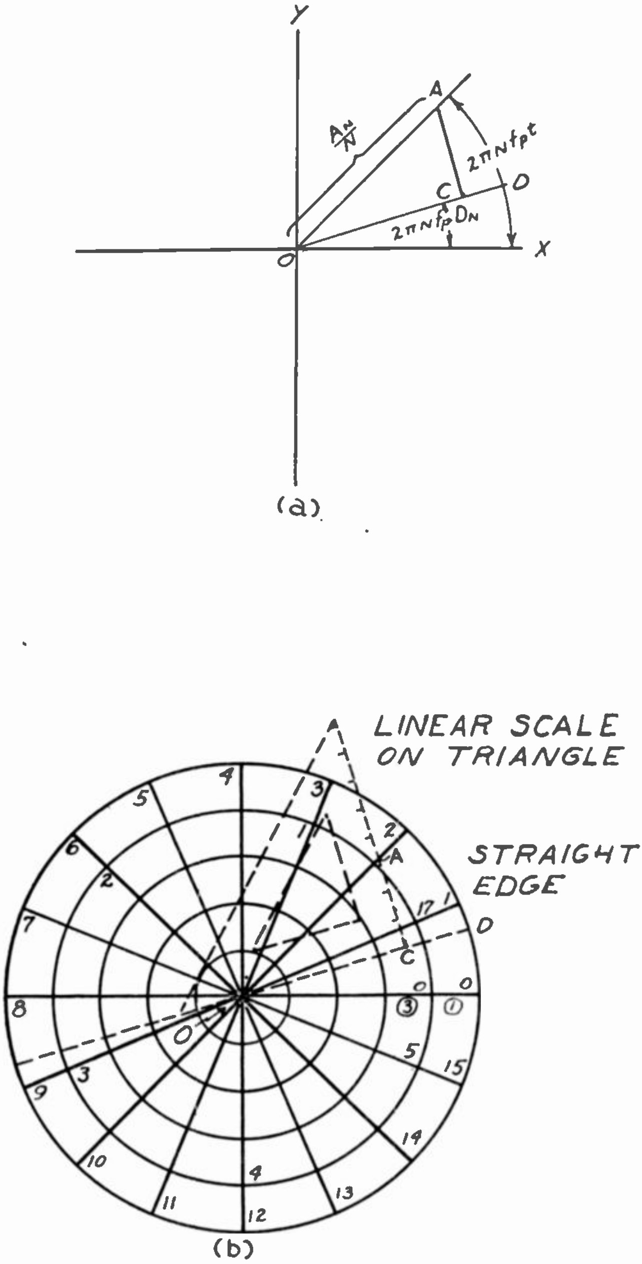

Fig. 14

(a) Basic principle of construction of synthesis charts.
(b) Use of synthesis charts.

period of the fundamental frequency) and (2) the contribution of harmonics beyond the 19th (4.75 megacycles) may be neglected.

The rapid use of the charts is expedited by observation of the following procedure in the preparing of Table II. A synthesis of the amplifier square-wave response shown in Figure 8 from the theoretical amplitude and phase characteristics in Figure 7 is used as an example.

(1) Record in column (c) the amplitude response in per cent corresponding to each harmonic divided by the order of the harmonic.

(2) Shift the real delay curve up or down but parallel to itself to any new position for which the minimum delay in the essential frequency band is equal to, or near, zero. Shifting (i.e., subtracting a constant delay) is a convenience in manipulating the charts but is not a necessity.

TABLE II

SYNTHESIS OF 32-STAGE AMPLIFIER

(a) Order of Harmonic, n	(b) Frequency of Harmonic (Mc)	(c) Amplitude Response of Harmonic $\div n$ (in per cent)	(d) Delay of Harmonic $\times 20n$	Contributions of Harmonics Read from Charts				
				(0) $t=0$ microseconds	(1) $t=0.05$	(2) $t=0.1$	(27) $t=1.35$	(28) $t=1.4$
1	0.25	100	1.5	−7.5	−2.5	+2.5		
3	0.75	34	4.7	−7.8	−2.9	+2.3		
5	1.25	21	8.5	−8.4	−3.6	+1.6		
7	1.75	15	12.9	−8.1	−4.3	+0.8		
9	2.25	11	17.6	−6.6	−4.3	+0.2		
11	2.75	7.5	22.2	−4.7	−3.7	−0.1		
13	3.25	5	26.0	−3.0	−2.8	0		
15	3.75	3.5	29.0	−1.7	−2.0	+0.2		
17	4.25	2.5	31.6	−0.9	−1.4	+0.7		
19	4.75	2	33.4	−0.5	−0.9	+0.4		
Add +50				+50	+50	+50		
Instantaneous response				+ 1	+22	+59		

(3) Record in column (d) the real delay (in microseconds) (or shifted delay, if performed) of each harmonic multiplied by 20 times the order of the harmonic.

(4) Proceed as follows to enter the contributions of the various harmonics in columns (0) to (28):

(A) Pass a straightedge through the center of Figure 11 along the radial line corresponding to a number on the circular scale (1) equal to the delay of the fundamental frequency recorded in column (d). Circular scale (1) refers to the series of radial lines numbered 0 to 80 in ring (1). The straightedge remains fixed until steps (B) and (C) below have been completed.

(B) On the radial line number 0 on scale (1) find a point which is at a distance from the center of the chart equal to the number in column (c). The length of the perpendicular from this point to the straightedge is the contribution of the fundamental frequency at $t = 0$. If the point referred to is *below* the axis, the contribution is entered in column (0) of the table with a negative sign. If above, the contribution is entered with a positive sign. The perpendicular is measured with a calibrated triangle illustrated in Figure 15. One leg slides along the straightedge while the calibrated leg forms the perpendicular.

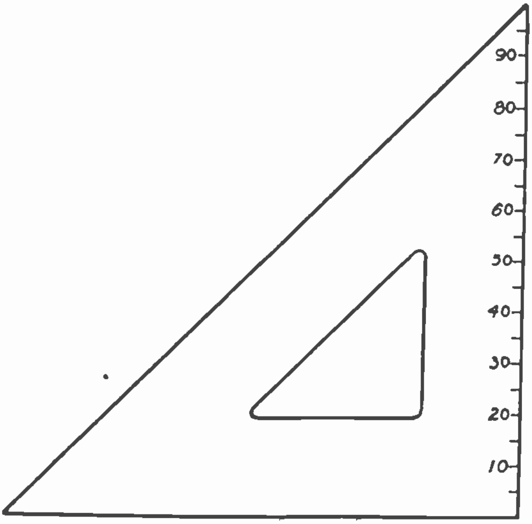

Fig. 15—Calibrated triangle[8] for reading contributions of harmonics from synthesis charts.

(C) The contribution of the fundamental at $t = 0.05$ microsecond, to be entered in column (1) opposite 0.25 megacycle, is found by locating the point 1 on scale (1) and proceeding as directed for point 0. Point 2 corresponds to 2×0.05 microsecond. In a similar manner, the first row of entries in Table II are made.

(D) In order to find contributions of the 3rd harmonic, pass the straightedge through the number on scale (1) corresponding to the delay factor from column (d). Use scale

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[8] The 100 per cent point on the radial scales of Figures 11, 12, and 13 should coincide with the 63.6 $(= (\pi 2)\ 100)$ per cent point on the calibrated triangle when the triangle is drawn to the same scale as the figures.

numbers on scale (3) in order to locate the perpendicular and proceed with the measurement of its length as directed above for the fundamental frequency.

(E) Complete the table for the other harmonics in a manner similar to the above.

(F) The instantaneous value of the square-wave response at $t = 0$ microseconds is found by summing column (0) and adding 50 per cent. Similarly, the response is found at other times by summing the appropriate column and adding 50 per cent.

B. *Examples of Synthesis of Square-Wave Response*

Several points on the square-wave response of a 2-stage amplifier are synthesized in Figure 8 from the amplitude and phase characteristics in Figure 7. Since the theoretical response is shown, a direct indication of the accuracy of the synthesis is available. The discrepancy is seen to be only a few per cent in this example.

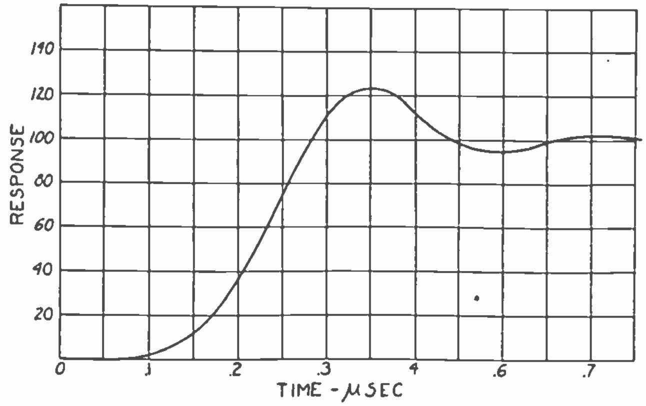

Fig. 16—Synthesis of the square-wave response of a 32-stage compensated resistance-capacitance amplifier in which $K = 1.51$ and $f_0 = 7.7$ megacycles.

The chart method greatly simplifies and shortens the synthesis of square-wave response in complex cases in which rigorous mathematical formulas for response are impracticable. A typical example is the response of a 32-stage amplifier which was synthesized in Figure 16 from the theoretical amplitude and phase characteristics of Figure 17. In this particular case, an absolute measure of the error of the synthesis is not known. The deviation of any synthesized square-wave response from the exact square-wave response is due only to errors in manipulation of the charts and the neglect of contributions of harmonics beyond the limits of the charts (4.25 megacycles).

IV. Evaluation of the Square-Wave Response

Assume that a subject containing an extensive dark area and an extensive white area with a sharp vertical junction between the two is used as a test subject for a determination of the fidelity of transmission of a television system.[9] An ideal scanning device in crossing the junction would generate a unit voltage which becomes the input signal for a television transmission system under test. The output signal or response of the system may be symbolized in Figure 18. The waveshape may be determined experimentally by a square-wave oscillograph or by synthesis from known sine-wave characteristics of the system. If the test subject were reproduced by an ideal scanning device (i.e., one having negligible aperture losses), actuated by the output signal,

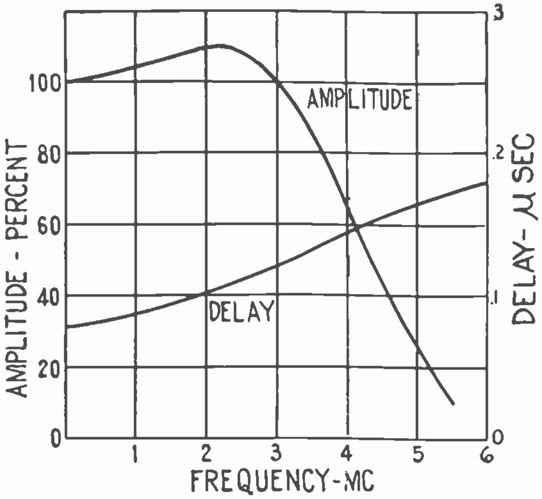

Fig. 17—Exact amplitude and delay characteristics of a 32-stage compensated resistance-capacitance amplifier in which $K = 1.51$ and $f_0 = 7.7$ megacycles.

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[9] The use of a single abrupt transition from one brightness to another brightness as a test subject for measuring "resolution" in a television picture was developed by R. D. Kell, A. V. Bedford, and G. L. Fredendall in "Determination of Optimum Number of Lines in Television System", *RCA REVIEW*, Vol. V, pp. 7-30; July, 1940. A test pattern consisting of several converging bars is much more commonly used in evaluating an entire television system on account of convenience. The results obtained, however, are less significant because the resolution of the individual bars is not affected by the phase fidelity, whereas it is known that phase fidelity affects the sharpness and utility of most television pictures.

the variation of light intensity on the screen along the scanning lines would be as shown in the figure when using for the ordinates and abscissas the light intensity and distance, respectively. The distance required for the complete change from black to white is greater than zero due to the finite rate of rise of the response curve. Upon close observation of the screen, the junction would appear blurred. Furthermore, several alternate light and gray striations following the junction would be observed as a consequence of the damped oscillation in the response. The overswing shown in Figure 18 would not be entirely objectionable for television purposes because at the optimum viewing distance *most* of the overswing is not distinguishable from the transition and the net effect would be substantially a single transition. In fact, the visual sharpness of the transition is *enhanced* by the overswing in the response wave. If a quantitative measure of the sharpness were found, the response wave in Figure 18, for example, could be compared directly with other waves having different shapes, and the relative merit of television apparatus with reference to picture sharpness could be determined.

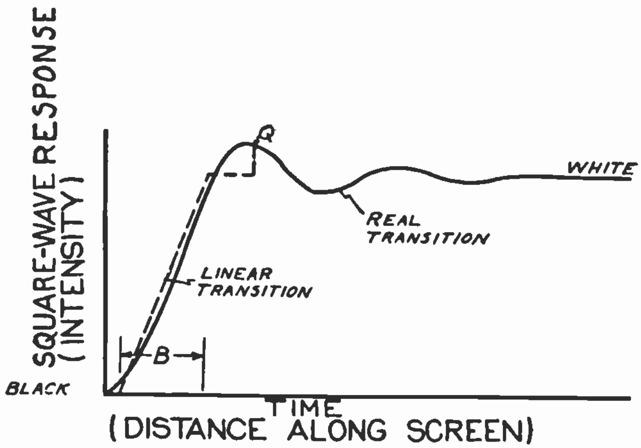

Fig. 18—Nonlinear response of a hypothetical complete television system and the equivalent linear response in terms of light along the screen surface or electrical response of the system.

With the object of finding such a measure of sharpness, the writers constructed several different synthetic black-white transitions with ink on cardboard using fine shading lines of variable widths to reproduce accurate half-tone values. When the transitions were viewed at a distance for which the "blur" was just discernible, that is, the optimum viewing distance it was observed that a simple *linear* transition similar to that shown dotted in Figure 18 could be found for each nonlinear transition such that the observer was unable to distinguish between the nonlinear transition and its linear equivalent.

The width of the visually equivalent linear transition was termed the equivalent "blur" of the nonlinear transition. The blur may be specified in units of distance along the picture screen or in the corresponding units of time. Complex transition curves containing damped oscillations of sufficiently long duration are properly represented by a linear transition followed by that part of the oscillation which the eye does not include with the transition.

The comparison method of determining blur is objectionable because the labor involved is great and the evaluation of each transition depends somewhat upon the observer's judgment. A method not subject to these objections has been devised whereby the blur may be found directly from a plot of the light intensity along a transition. The steps involved in applying the method are given below in a "Generalized Statement" which defines the conditions under which a linear transition is visually equivalent to a nonlinear transition such as Figure 18. For clarity, this "Generalized Statement" is presented in the form of a law or theorem, but we do not presume to use these terms until the statement has been proved more adequately by theory or experiment than is done in this paper.

Generalized Statement

A linear transition having a uniform rate of change of intensity along the surface from a first mean brightness to a second mean brightness is visually equivalent at the optimum viewing distance to any nonlinear transition from the first mean brightness to the second mean brightness when conditions 1 and 2 below are fulfilled.

Condition 1—The summation of the weighted differences of the light intensities of the linear transition and of the nonlinear transition is zero, where the weighted differences are the real differences of light intensity along the transitions multiplied by a weighting factor. The The weighting factor varies linearly with distance from a value of — 1 at the first inflection point of the linear transition to + 1 at the second inflection point; also linearly with distance from a value of zero at a point preceding the first inflection point by half the distance between inflection points, to the — 1 value at the first inflection point; also linearly with distance from the + 1 value at the second inflection point to zero at a point following the second inflection point by half the distance between the inflection points; and is zero for all other points.

Condition 2—The summation of the differences of the light intensities along the transitions is zero, over the range where the weighting factor is not zero.

According to the generalized statement, the linear transition E in Figure 19 is visually equivalent to the transition A. In this and other cases, it is convenient to consider the vertical dimensions J of the difference areas C, D, F, etc., as positive when A is above E and negative when A is below E. Then, if these dimensions are multiplied by the corresponding ordinates of the weighting curve N and plotted as at V, the positive and negative areas C', D', E', etc., are formed. Condition 1 is fulfilled if the algebraic sum of these areas C', D', F', etc., equals zero. Condition 2 simply requires that the algebraic sum of the positive and negative areas C, D, F, etc., equals zero.

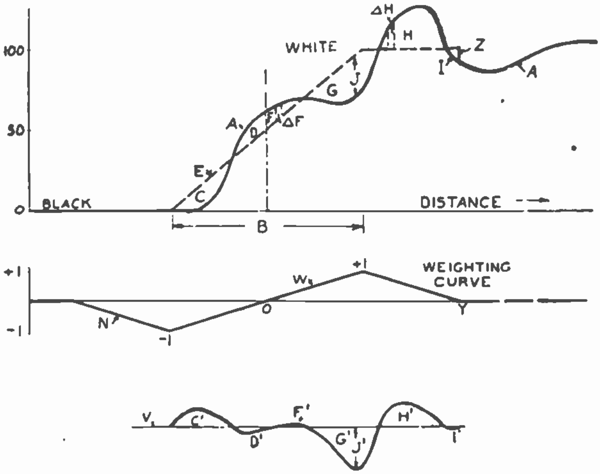

Fig. 19—Linear transition E is equivalent to A. By condition 2, the algebraic sum of areas C, D, F, G, H, and I is zero. By condition 1, the sum of weighted areas C, D, F, etc., shown at C', D', F', etc., is zero.

The location of a line E that fulfills the two conditions for a particular nonuniform transition is obtained by a trial-and-error method. The equivalent blur is indicated as the distance B. Each trial curve E requires a different weighting curve N since the weighting curve is itself defined by the equivalent linear-transition curve. Such a procedure would appear to be laborious since the line E has two variable characteristics, slope and position. However, the number of trials required is much reduced by the knowledge that for small changes of slope of line E only condition 1 is affected and for small changes of position only condition 2 is affected.

Although the generalized statement for defining a linear transition which is visually equivalent to a nonlinear transition is essentially empirical, the conditions appear to be consistent with observed properties of the eye. It is well known that the eye responds to a surface consisting of many tiny, uniformly distributed white dots on a black background (or black dots on a white background) as though the surface were uniformly white but the illumination reduced so that the total reflected light is the same. In other words, the eye responds to the average[10] brightness or to the total brightness. Condition 2 is consistent with this observation in that it requires equality of the total amount of light of the equivalent transition and the nonlinear transition.

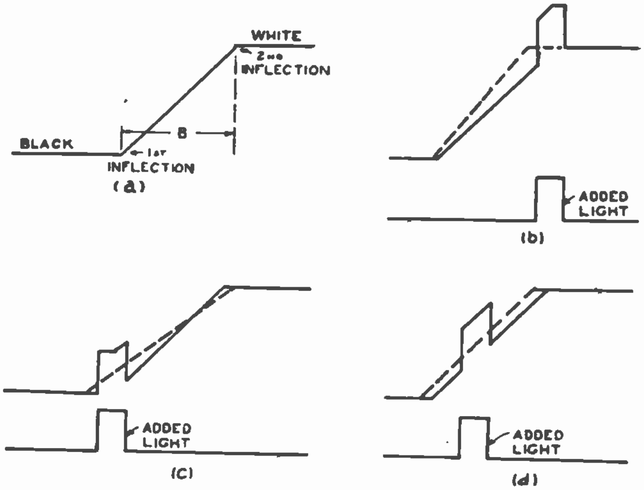

Fig. 20—Actual transitions are shown in solid lines and equivalent linear transitions are shown in dotted lines.

The plausibility of condition 1 may be established by using the simple transitions of Figure 20 to illustrate some basic factors which affect apparent steepness. Figure 20(a) represents a linear change from black to white or from any half tone to another half tone. The blur by our definition is the distance *B*. Now assume that an extra amount of light is added near the second inflection point as shown in Figure 20(b). As a result, the transition has been made effectively steeper so that the equivalent linear transition must also be made steeper as shown by the dotted line.

If the light were added near the first inflection point as shown in

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[10] The eye also responds to the average brightness of a rapidly flickering source.

Figure 20(c), it is apparent that the transition would be *less* steep as indicated by the dotted linear transition. From these two observations, it is plausible that the addition of light at the *middle* of the transition as in Figure 20(d) would not alter the effective steepness. The equivalent linear transition, however, would be displaced to the left as required by condition 2.

If light were added at a point considerably ahead of the first inflection point of the linear transition or considerably after the second inflection point, it is reasonable that the effect on the steepness would be less than if the light were added near the inflection points.

Let us refer back to the example of Figure 19 in which curve *E* is assumed to be the visual equivalent of the curve *A*. Obviously, *E* is *identical* to curve *A* except for the difference areas *C*, *D*, *G*, etc. Some of the areas such as *C* and *H* render curve *A* effectively steeper than the equivalent curve but other areas such as *D* and *G* detract from the steepness of curve *A*.

If curve *E* is to be the equivalent of curve *A*, the algebraic sum of the *effects* of all the aiding difference areas *C*, *H*, etc., must be canceled by the opposing areas. Condition 1 states this requirement and provides a weighting factor for determining the effectiveness of each difference area resulting from its location, as discussed in connection with Figure 20.

A linear weighting curve has been taken arbitrarily due to the absence of evidence which would point to a specific form for the curve. A linear variation is simple and may also be considered as a mean between the various possible concave and convex forms.

It should be noted especially that in the application of the generalized statement to transitions which have appreciable irregularities in the region in which the weighting curve is zero (beyond *Y*), the irregularities are not included in the equivalent blur. As shown in Figure 19, the equivalent curve may be continued through *Z* to the actual curve *A*. It should be noted also that the equivalent transition defined by the statement is intended to be equivalent only when viewed at a distance for which the nonlinear transition is indistinguishable from a linear equivalent transition. At such a distance, a very definite blur may be visible without the *shape* of the transition being discernible.

The observations made in the several paragraphs above lend support to the concept of an equivalent linear blur but obviously do not place the concept on a firm physical basis. Furthermore, the physical accuracy of the particular conditions set forth in the generalized statement have not been established because of inadequate knowledge of the eye and brain. However, actual viewing tests with specific transitions lead us to believe that the application of the generalized statement is sufficiently accurate to serve a useful purpose for the evaluation of the transitions found in television.

The authors feel that such a measure of blur also can have utility in specifying the quality of many other devices related to vision, such as lens system, photographic film and processes, facsimile transmission, printing, duplicating processes, and paper.

V. APPLICATION OF SQUARE-WAVE METHODS IN TELEVISION

Square-wave and sine-wave measurements should be mutual aids in the solution of many television problems. In some instances, a square-wave measurement may furnish the data for analysis into amplitude and delay characteristics. In others, the evaluation of blur is indicated. In some applications, sine-wave measurement and synthesis of the square-wave response may be indicated, followed perhaps by evaluation. Most applications will not require the use of all three processes of square-wave treatment.

We know no simple satisfactory method of judging the degree of fidelity of a television system from inspection of the sine-wave characteristics. If the square-wave response of the system cannot be secured experimentally, resort must be had to synthesis from amplitude and phase data. However, acceptable tolerances in terms of sine-wave performance may be more easily determined when the amplitude and phase characteristics corresponding to a large variety of transient-response curves have been determined and cataloged.

In design work in which the characteristics of television apparatus must be determined by calculation based on circuit constants, synthesis from the sine-wave characteristics probably constitutes the only feasible means for obtaining the square-wave response. When the apparatus is susceptible to experimental test, the most expeditious method is direct measurement of the square-wave response by oscillographic equipment such as described in a companion paper.[3] An analysis for amplitude and phase characteristics may then be performed in order to facilitate the design of equalizing networks if required or for any other purpose. In this connection it is significant that the experimental difficulty of phase measurements of extensive apparatus such as a complete television system including transmitter and receiver by sine-wave methods is usually so great that the attempt is not often made.[11, 12] Notwithstanding, a reasonably linear phase shift is conceded to rank with amplitude uniformity in importance. The extreme ease with which square-wave response can be recorded with suitable equipment has been pointed out in the companion paper. The analysis of the response for delay versus frequency through the use of charts is simple and immediate.

In particular cases, square-wave measurements may provide useful data which are almost impossible to obtain by sine-wave measurements and synthesis. An example is the modulation amplifier of a television transmitter in which the output impedance may be relatively high at low frequencies, in order to permit a high output voltage, and relatively low at high frequencies such that the frequency fidelity of the stage alone is poor. Adequate equalization may be inserted in earlier stages of the system but high-frequency components may be saturated at high levels. A sine-wave characteristic of the entire amplifier taken at a low level for which saturation is negligible would indicate high fidelity. It would be almost impossible to synthesize from the low-level sine-wave data the transient response for a high level corresponding to conditions occurring in common use. Square-wave oscillographic tests, however, would indicate the transient response corresponding to any desired level. An evaluation of the square-wave response for the purpose of finding the blur corresponding to various levels would be significant but a determination of the sine-wave response at levels where saturation exists would have no meaning in the usual sense.

In general, square-wave methods have greatest usefulness (as compared with sine-wave methods) in dealing with performance of units which are likely to contribute a substantial amount of the distortion in the transmission characteristic of the system. Included in this classification are entire transmitters, entire receivers, long transmission lines, pickup chains, and any single amplifier stages which may be regarded as "bottleneck" stages. It would be rather pointless to find the square-wave response and the equivalent blur of a single stage of a video voltage amplifier of good fidelity in a system in which many similar stages exist. Usually the distortion of a single stage is so small that only the accumulated effect of several stages is clearly evident in the over-all square-wave response. The writers are not aware of the existence of a practicable method of combining the individual

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[11] B. D. Loughlin, "A Phase Curve Tracer for Television", *Proc. I.R.E.*, Vol. 29, pp. 107-115; March, 1941. Loughlin describes apparatus which furnishes a direct plot of phase versus frequency on the screen of a cathode-ray tube. The complexity of the apparatus may limit its general utility.

[12] M. E. Strieby and J. F. Wentz, "Television Transmission Over Wire Lines", *Bell Sys. Tech. Jour.*, Vol. 20, pp. 62-81; January, 1941.

square-wave response of two or more units for the purpose of finding the over-all square-wave response. Even if a feasible method should exist, the experimental errors involved in many individual measurements would jeopardize the dependability of the calculated over-all response.

If units of good fidelity must be considered individually, it becomes expedient to determine first the sine-wave characteristics and then to combine the sine-wave characteristics so that a synthesis of the over-all square-wave response may be performed.

The accuracy of square-wave methods is more than adequate to reveal those imperfections of transmission which cause discernible effects in a television picture. Therefore, the authors hope that this paper will be a contribution to the general problem of working out electrical specifications for television transmitters and other television apparatus, giving information regarding the steepness of rise and the amplitude of over-swing of the square-wave response.

APPENDIX I

ANALYSIS OF SQUARE-WAVE RESPONSE

Let the square-wave response be approximated by a stepped wave f as shown in Figure 1. The kth step may be approximated by the series

$$A_k \left[\frac{1}{2} + \frac{2}{\pi} \sum_{1}^{N} \frac{1}{n} \sin n\omega_o (t - T_K) \right] \tag{3}$$

where N is very large.

The sum of M steps leads to the following expression for the stepped wave

$$\sum_{m=1}^{M} \frac{A_m}{2} + \frac{2}{\pi} \sum_{m=1}^{M} \left[A_m \sum_{n=1}^{N} \frac{1}{n} \sin n\omega_o (t - T_m) \right]. \tag{4}$$

The constant term in (3) is not of interest. If the order of summation of the second term is reversed, there results

$$\frac{2}{\pi} \sum_{n=1}^{N} \frac{1}{n} \sum_{m=1}^{M} \left[A_m \sin n\omega_o (t - T_m) \right]. \tag{5}$$

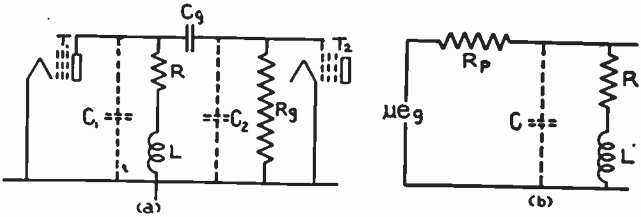
The inner sum written for some specific value of n as N_1 is the total N_1th harmonic content of the stepped wave. That is,

$$\begin{aligned} E_{N1} &= \frac{2}{n}\frac{1}{N_1}\sum_{m=1}^{M} A_m \sin N_1\omega_o(t-T_m) \\ &= \frac{2}{\pi}\frac{1}{N_1} B_{N_1} \sin N_1\omega_o\,(t-t_{N_1}). \end{aligned} \qquad (6)$$

Since the input square-wave signal may be approximated by the form

$$\frac{1}{2}+\frac{2}{\pi}\sum_{n=1}^{N}\frac{1}{n}\sin n\omega_o t$$

it follows that the N_1 component of the input is



(a) Compensated resistance-capacitance amplifier. (b) Equivalent circuit for high frequencies.

Fig. 21

$$\frac{2}{\pi}\frac{1}{N_1}\sin N_1\omega_o t.$$

The amplitude response of the circuit is, therefore, B_{N1} and the time delay is T_{N1}. These two quantities may be found graphically as indicated in Figure 1(a) and (b).

Since ω_o may be allowed to approach zero, it follows that reference to a fundamental frequency is not required and the amplitude and phase corresponding to any frequency may be found.

APPENDIX II

THE COMPENSATED RESISTANCE-CAPACITANCE AMPLIFIER

A schematic diagram of the amplifier appears in Figure 21. When $R_p >> Z\ (L, C, R)$ the response of the equivalent circuit to a unit function is expressed by the following equation:

$$e = \frac{\mu R}{R_p} [1 - e^{-\pi f_0 K t} \{A \sin (\Omega_1 t + \psi) + Bt \sin (\Omega_1 t + \beta\}]$$

in which

$$t = \text{time}$$
$$f_0 = 1/(2\pi\sqrt{LC})$$
$$A = \sqrt{1 + M^2}$$
$$K = 2\pi f_0 RC$$
$$M = \frac{-3 + 3K^2 - K^4/2}{4K(1 - K^2/4)^{3/2}}$$
$$N = \frac{\pi f_0 (K^2 - 1)}{K^2\sqrt{1 - K^2/4}}$$
$$\Omega_1 = 2\pi f_0 \sqrt{1 - K^2/4}$$
$$B = \sqrt{P^2 + N^2}$$
$$\psi = \tan^{-1} \frac{1}{M}$$
$$\beta = \tan^{-1} \frac{N}{P}$$
$$P = \frac{\pi f_0 (3 - K^2)}{2K(1 - K^2/4)}.$$

The amplitude response is

$$\text{amplitude} = \frac{\mu R}{R_p} \sqrt{\frac{1 + B^2/K^2}{B^2K^2 + (B^2 - 1)^2}}$$

$$\text{delay} = \frac{1}{2\pi f} \tan^{-1} (1 - K^2 - B^2) B/K$$

in which

$$f = \text{frequency}$$
$$B = f/f_0$$
$$K = 2\pi f_0 RC.$$

APPENDIX III

RELATION BETWEEN SQUARE-WAVE ANALYSIS AND FOURIER ANALYSIS

Assume that the response of a linear electrical network to a unit function is represented by $e_1(t)$ in Figure 22(a). The response $e_2(t)$ of the same circuit to a square pulse Δt seconds long is found by shifting $e_1(t)$ to the right along the time axis by an amount Δt and subtracting the result from $e_1(t)$ that is, $e_2(t) = e_1(t) - e_1(t + \Delta t)$. If the condition is imposed that $e_1(t) = 1$ for $t \geqq t_q$, then $e_2(t) = 0$ for $t \geqq (t_q + \Delta t)$.

It is clear that the waveform of the response $e_3(t)$ of the same circuit to the periodic pulse wave of Figure 22(b) is given for each cycle by $e_2(t)$ if $T_0 \geqq (t_q + \Delta t)$. Hence, the coefficients of the Fourier series written for $e_3(t)$ may be expressed in terms of the ordinate readings of $e_1(t)$ taken at successive equal time intervals. Thus

$$e_3(t) = a_0 + a_1 \cos 2\pi f_0 t + \cdots + a_k \cos K2\pi f_0 t + \cdots$$

$$+ a_{n/2} \cos \frac{n}{2} 2\pi f_0 t + b_1 \sin 2\pi f_0 t + \cdots$$

$$+ b_k \sin K2\pi f_0 t + \cdots$$

$$+ {}_{(n/2)-1} \sin \left(\frac{n}{2} - 1 \right) 2\pi f_0 t$$

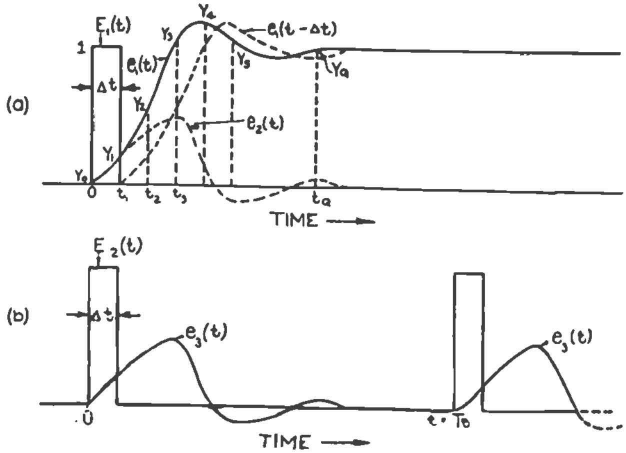


Fig. 22—Fourier analysis applied to the response of a circuit to a square pulse compared with square-wave analysis applied to the response to a unit function.

where

$$\begin{aligned} a_k &= 2\Delta t f_0 [\, (y_1 - y_0) \cos K2\pi\Delta t f_0 \\ &\quad + (y_2 - y_1) \cos K4\pi\Delta t f_0 + \cdots \\ &\quad + (y_q - y_{q-1}) \cos Kq2\pi\Delta t f_0] \\ &= 2\Delta t f_0 A \end{aligned}$$

and

$$\begin{aligned} b_k &= 2\Delta t f_0 [(y_1 - y_0) \sin K2\pi\Delta t f_0 \\ &\quad + (y_2 - y_1) \sin K4\pi\Delta t f_0 + \cdots \\ &\quad + (y_q - y_{q-1}) \sin k_q 2\pi\Delta t f_0] \\ &= 2\Delta t f_0 B\cdot \end{aligned}$$

The Fourier series for $E_2(t)$ is

$$E_2(t) = \frac{T_0}{\Delta t} + \frac{2}{\pi} \sum_K \frac{1}{k} \sin K\pi \frac{\Delta t}{T} \cos 2\pi K f_0 t\cdot$$

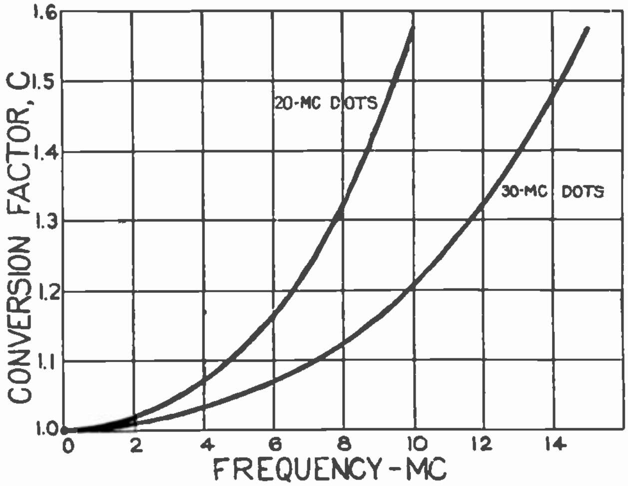


Fig. 23—Conversion factor for obtaining Fourier analysis from square-wave analysis. C times amplitude according to square-wave analysis equals amplitude according to Fourier analysis.

Hence, the amplitude characteristic of the circuit is given by

$$R(K, f_0) = \frac{\pi \Delta t K f_0}{\sin \pi \Delta t K f} \sqrt{A^2 + B^2}$$

$$\text{or } \lim_{f_0 \to 0} R(K, f_0) = R(f) = \frac{\pi \Delta t f}{\sin \pi \Delta t f} \left[\lim_{f_0 \to 0} \sqrt{A^2 + B^2} \right]$$

and the delay characteristic is given by

$$\lim_{f_0 \to 0} \frac{\theta(K, f_0)}{2\pi f} = \frac{1}{2\pi f} \tan^{-1} \frac{B}{A}.$$

Now $\sqrt{A^2 + B^2}$ is the same expression which is obtained for the amplitude characteristic by the graphical method of square-wave analysis. Hence, $\pi \Delta t f / \sin \pi \Delta t f$ may be regarded as the conversion factor which converts square-wave analysis into Fourier analysis when latter is based on the square pulse.

The magnitude of the conversion factor is shown in Figure 23. It may be concluded that for most applications of square-wave analysis in television, the conversion factor may be taken equal to 1. The delay characteristics as determined by square-wave analysis and Fourier analysis are identical.

TRANSMISSION OF TELEVISION SOUND ON THE PICTURE CARRIER*†

BY

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Summary—*Several pulse methods for the transmission of television sound on the picture carrier during the line-blanking intervals are analyzed from the points of view of signal-to-noise ratio, audio fidelity, and transmitter and receiver design.*

The advantages of duplex transmission are: (1) elimination of a separate sound transmitter; (2) elimination of the ambiguity and difficulty which may occur when a standard frequency-modulated sound signal is tuned in; (3) freedom of the audio output from the type of distortion which occurs in frequency-modulated receivers as a consequence of excessive drift of the frequency of the local oscillator; and (4) improvement of the phase characteristic of the picture intermediate-frequency amplifier resulting from elimination of trap circuits.

The highest audio-modulation frequency in duplex systems must not exceed one half of the line-scanning frequency. This is a disadvantage under the present television standards which specify a line frequency of 15,750 cycles per second.

With the exception of pulsed frequency modulation, the signal-to-noise ratios of sound in duplex systems are not so great as the ratio offered by the transmission of a standard frequency-modulated carrier. The comparison is subject to the condition that the amplitude of the frequency-modulated carrier is 0.7 of the peak amplitude of the duplex carrier. The signal-to-noise ratio of a pulsed frequency-modulated signal may equal the ratio of a standard frequency-modulated signal up to a critical distance from the transmitter, but is less at greater distance.

INTRODUCTION

FROM time to time, proposals have been made that the sound accompaniment for television may be transmitted by a modulation of the picture carrier during the line-blanking intervals when no picture detail is transmitted. Improved reception of picture and sound, decreased investment in receivers and transmitters, and greater channel width for the picture signal are mentioned as possibilities of a "duplex" transmission of picture and sound. The purpose of this paper is to assist engineers in their evaluation of duplex transmission as a

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‡ Columbia Broadcasting System, New York, N. Y. Work covered in this paper was done while the author was a member of RCA Laboratories, Purdue University, Lafayette, Ind.

practicable system by offering an analysis of several methods of duplexing.

A review of the method of sound transmission and reception in use is helpful as a setting for the discussion. In the present arrangement, sound is transmitted by a frequency-modulated sound transmitter operating on a carrier frequency which is 4.5 megacycles above the picture carrier. At the position of the sound carrier, the recommended standards[1] state that the field strength of the picture sidebands shall not exceed 0.0005 of the picture carrier. There is essentially nothing at the transmitting point to distinguish a television sound transmitter

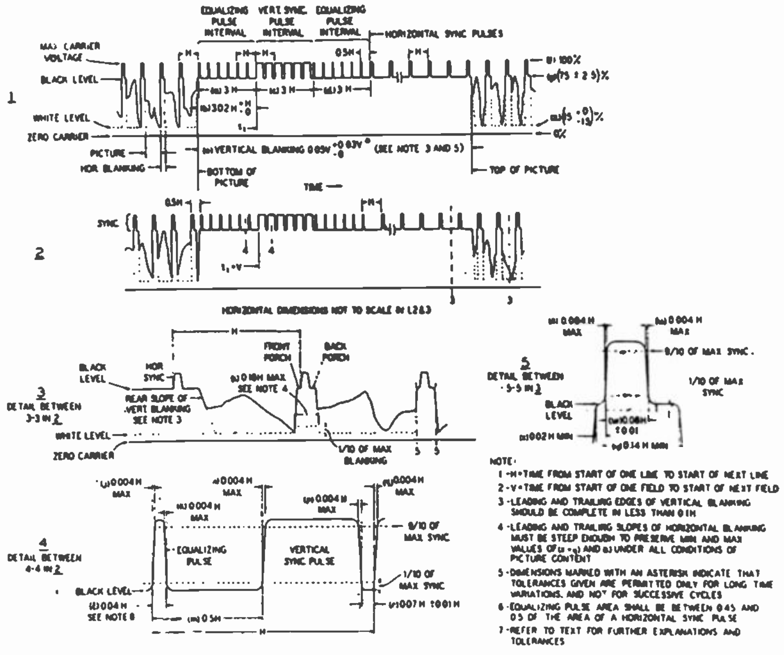

Fig. 1—Television synchronizing wave form.

from a conventional frequency-modulation transmitter designed for operation in the frequency-modulation band. The sound receiver is likewise conventional and may share only the same heterodyne oscillator with the picture receiver.[2]

The picture transmitter is amplitude-modulated and radiates a wave form illustrated by Figure 1. Picture content is transmitted

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[1] Final report of the Radio Technical Planning Board.

[2] In some designs, the sound and picture signals are amplified at intermediate frequency in one or more common stages before branching off into separate intermediate-frequency amplifiers occurs.

during about 85 per cent of the total "time on the air." No picture detail is transmitted during the blanking intervals of the scanning tubes in the transmitter and receiver. Such "idle" intervals amount to about 15 per cent of the total time.

Proposals[3, 4] for duplexing have been directed, therefore, toward the utilization of some part of the blanking interval for sound transmission. Thus the television transmitter would be converted into a picture-sound duplex transmitter which radiates picture intelligence during 85 per cent of the time and sound intelligence during some part of the remaining 15 per cent, using only one antenna and one radio-frequency power amplifier. There would be a synchronized electronic switch in the receiver for the opening of the sound channel to the video signal sometime during the blanking interval.

Factors which are involved in a comparison of duplex methods and the present method of continuous transmission of sound are the following: (1) audio fidelity; (2) signal-to-noise ratio; (3) amount of interaction between video and audio signals; (4) permanency of receiver alignment; (5) picture quality; (6) ease of receiver tuning; (7) cost of receiver; (8) cost of transmitter.

Certain advantages and disadvantages of a duplex system may be predicted in advance of a theoretical and experimental analysis. First, there is no separate sound transmitter and antenna. This economic advantage can not be accorded much weight unless there is a resultant economy in the television receiver because the ratio of receivers to transmitters is so great that the economics of the receiver is the controlling factor.

A significant advantage of a duplex receiver is the freedom of the audio output from the type of distortion which occurs in frequency-modulation receivers as a consequence of excessive drift of the frequency of the local oscillator. Audio modulation is conveyed in a duplex system by the envelope of a radio-frequency carrier and is relatively unaffected by instability of the local oscillator.

Sound-rejector circuits in the picture intermediate-frequency amplifier can be removed with a resulting reduction in phase distortion and some improvement in picture quality.

The ambiguity involved in tuning-in a conventional frequency-modulation signal is removed. In the present system a choice must be

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[3] H. E. Kallman, "Audio and video on a single carrier", *Electronics*, vol. 14, pp. 39-42; May, 1941.

[4] Numerous patents including: U. S. patents, No. 1,655,543, R. A. Heising; No. 1,887,237, J. L. Finch; No. 2,061,734, R. D. Kell; No. 2,083,245, H. Shore and J. N. Whittaker; No. 2,086,918, D. G. C. Luck; No. 2,089,639, A. V. Bedford; No. 2,227,108, H. A. Rosenstein; No. 2,257,562, H. Branson; No. 20,153, E. F. W. Alexanderson (reissue).

made of tuning for minimum interference in the picture or minimum noise in the sound in receivers which are somewhat misaligned.

There is also available a small extension of the video band into the space now assigned as a guard band for the sound; this amounts to about one quarter of a megacycle.

It may be anticipated that the signal-to-noise ratio with duplex sound would be unfavorable as a consequence of the reduced time for transmission of sound. This is a serious obstacle in some duplex systems. A further limitation is the imposition of a maximum audio frequency that may be transmitted without the introduction of spurious frequencies into the audio spectrum. This restriction has been noted before in pulse transmission of sound. Finally, there is the complication of synchronizing the electronic sound selector in the receiver with the line-scanning frequency.

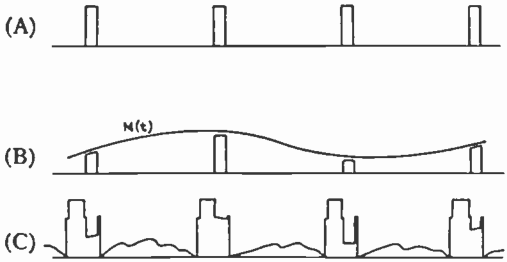

Fig. 2—Amplitude-modulated duplex wave forms.
(A) pulse carrier
(B) audio wave form *M(t)* and amplitude-modulated pulse carrier
(C) amplitude-modulated pulses combined with television wave form

DUPLEX METHODS

1. *Amplitude-Modulated Pulses*

One of the most obvious duplex systems is the modulation of the amplitude of a rectangular pulse wave in accordance with the audio signal and the insertion of the modulated pulses in the line-blanking interval of the video signal. Figure 2 illustrates the successive steps at the transmitter. The pulse wave form shown in (A) of Figure 2 is amplitude modulated by the audio signal $M(t)$, as illustrated in (B). In (C) the modulated pulses have been inserted in the part of the blanking interval following the synchronizing pulse. Such a composite wave would be applied as modulation of the picture carrier.

If it is granted that the synchronized electronic switch in the receiver is able to select the pulses from blanking so that the pulse wave in Figure 2(B) is recovered, the audio fidelity of the duplex system can be found from the solution for the frequency components of (B). An analysis shows that the spectrum consists of the applied audio-modulation frequency f_0 and a large number of sidebands $(f_c \pm f_0)$, $(2f_c \pm f_0)$ $(3f_c \pm f_0) \cdots (nf_c \pm f_0)$ in which f_c is the fundamental frequency (line frequency) of the pulse wave.[5] The amplitude of each group obeys the damped sine-wave law $\sin n\pi r/n$ where n is the order of the sideband and r is the ratio of the width of the pulse to the fundamental period. In a television application, r could not exceed about 0.06 and hence the amplitude factor, $\sin n\pi r/n$ changes slowly. When f_0 exceeds $1/2f_c$, there is overlapping of the first-order lower sideband and the audio-frequency component, as well as general overlapping of adjacent sidebands of higher order. We do not have knowledge of any detector whereby an undistorted audio signal can be recovered from this multiplicity of overlapping sidebands. However, if the frequency of the audio modulation is restricted by a low-pass filter at the transmitter to less than one half the fundamental pulse frequency, this confusion is avoided. A similar filter must be installed in the receiver for the rejection of frequencies above $f_c/2$. Such a low-pass filter in the receiver functions as a distortionless detector of the audio modulation. Therefore, the theoretical upper limit of the audio bandwidth is equal to one-half of line frequency, or 7875 cycles with the present standards.

Vertically scanned pictures would allow a greater maximum audio frequency of $4/3 \times 7875$, or 10,500 cycles.[6] However, it has been observed in laboratory tests that moving subjects scanned vertically with an interlaced pattern do not, in general, reproduce with as much detail as horizontally scanned subjects. This is due, probably, to the predominance of horizontal motion in average subject matter. Hence, it appears that vertical scanning must be rejected as a means of increasing the maximum audio frequency.

The most promising way of increasing the upper audio limit in a monochrome system is an increase in the video bandwidth. The two quantities are related by the formula[7]

$$f_a = K_1 \sqrt{f_v} \qquad (1)$$

in which

f_a = maximum audio frequency

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[5] Appendix I.

[6] The line frequency in a vertical scanning system is 4/3 the line frequency of the standard horizontal scanning system with an aspect ratio of 4 to 3.

[7] Appendix II.

f_v = video bandwidth

K_1 = a constant.

Figure 3 shows the correlation between sound band, video band, and number of lines for monochrome and color transmissions. The latter is assumed to be a sequential tricolor system with an interlace ratio of 2:1 and a color-field frequency of 120 cycles.[8] For a given video bandwidth, the quality of duplex sound in terms of audio bandwidth may be made 1.4 times better for color than for monochrome television. Thus, high-fidelity sound (11,000 cycles) may be duplexed along with a color picture of about 360 lines over a video channel of 4 megacycles, while a monochrome picture of 525 lines, which occupies the same video band, accommodates only about 7800 cycles. A maximum audio frequency of 11,000 cycles would require over 700 lines in a monochrome system.

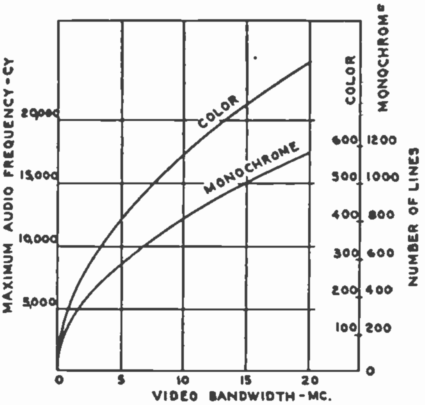

Fig. 3—Maximum audio frequency versus video bandwidth and number of lines.

The corresponding radio-frequency channel in all cases is approximately 30 per cent greater than the video bandwidth as a consequence of the additional space required by the vestigial sideband.

A proposal for sound transmission has been disclosed which removes the limitation on the maximum audio frequency of one-half line frequency.[9] In effect, the system provides for the modulation of a rectangular pulse carrier of two times line frequency and the subsequent delay of alternate pulses to a time position which permits transmission of pairs of pulses during horizontal blanking time. At the receiving

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[8] P. C. Goldmark, J. N. Dyer, E. R. Piore, and J. M. Hollywood, "Color television", *Proc. I.R.E.*, vol. 30, pp. 162-182; April, 1942.

[9] A. V. Bedford, U. S. Patent No. 2,089,639.

point, the previously undelayed pulses are delayed before detection, thus restoring the modulated pulse signal to its original form as a wave of double line frequency. The maximum audio frequency has thereby been increased to line frequency. This system, however, does not appear to be economically feasible from the point of view of receiver design at the present time.

TABLE I

SIGNAL-TO-NOISE RATIOS (root-mean-square)

Method of Transmission	(1) Signal-to-Noise	(2) Signal-to-Noise (Critical)	(3) Signal-to-Noise $P/N = K/d^2$	(4) Signal-to-Noise $P/N = K/d^2$	(5) Signal-to-Noise (Critical)
Standard Amplitude Modulation	$\frac{2P}{\sqrt{f_a}\,N}$	None	$\frac{2}{\sqrt{f_a}}\frac{K}{d^2}$	$0.023\frac{K}{d^2}$	None
Standard Frequency Modulation	$\frac{\sqrt{3}\,f_d}{f_a^{3/2}}\frac{P}{N}$	$\sqrt{6}\left(\frac{f_d}{f_a}\right)^{3/2}$	$\frac{\sqrt{3}\,f_d}{f_a^{3/2}}\frac{K}{d^2}$	$f_d = 150$ kilocycles $0.417\frac{K}{d^2}$	$f_d = 150$ kilocycles 217
Amplitude-Modulated Pulses	$\frac{3\sqrt{2}\sqrt{r}}{4\sqrt{f_a}}\frac{P}{N}$	None	$\frac{3\sqrt{2}\sqrt{r}}{4/\sqrt{f_a}}\frac{K}{d^2}$	$0.003\frac{K}{d^2}$	None
Symmetrical Width-Modulated Pulses	$\frac{3w}{2t_s\sqrt{f_v}}\frac{P}{N}$	$\frac{4w}{t_s}$	$\frac{3w}{2t_s\sqrt{f_v}}\frac{K}{d^2}$	$0.012\frac{K}{d^2}$	62
Dissymetrical Width-Modulated Pulses	$\frac{3w}{2\,t_s\,\sqrt{f_v}}\frac{P}{N}$	$\frac{4\sqrt{2}\,w}{t_s}$	$\frac{3w}{\sqrt{2}\,t_s\,\sqrt{f_v}}\frac{K}{d^2}$	$0.016\frac{K}{d^2}$	87
Pulses of Frequency Modulation	$\frac{3\sqrt{6}\sqrt{r}\,f_d}{8f_a^{3/2}}\frac{P}{N}$	$\sqrt{6}\sqrt{r}\left(\frac{f_d}{f_a}\right)^{3/2}$	$\frac{3\sqrt{6}\sqrt{r}\,f_d}{8f_a^{3/2}}\frac{K}{d^2}$	$f_d = 150$ kilocycles $0.052\frac{K}{d^2}$	$f_d = 150$ kilocycles 107
Pulses of Frequency Modulation	$\frac{3\sqrt{6}\sqrt{r}\,f_d}{8f_a^{3/2}}\frac{P}{N}$	$\sqrt{6}\sqrt{r}\left(\frac{f_d}{f_a}\right)^{3/2}$	$\frac{3\sqrt{6}\sqrt{r}\,f_d}{8f_a^{3/2}}\frac{K}{d^2}$	$f_d = 1200$ kilocycles $0.417\frac{K}{d^2}$	$f_d = 1200$ kilocycles 2420

The signal-to-noise formulas in columns (1). (3). and (4) of the table for standard frequency modulation, width-modulated pulses, and pulses of frequency modulation during postblanking, are valid only for ratios higher than the critical ratio since the formulas are derived with the assumption that noise limiting is effective. Thus it may appear, with only a casual reading of the table, that the ratio is always the same for standard frequency modulation and pulses of frequency modulation $f_d = 1200$ kilocycles. The fact is that the two types of transmission yield equal signal-to-noise ratios only when the critical ratio for pulses of frequency modulation is exceeded. At greater distances from the transmitter standard frequency modulation is superior.

Values of constants: $f_a = 7500$ cycles per second; $r = 0.06$; $w/t_s = 15.4$; $f_v = 4 \times 10^6$ cycles per second.

Success or failure of a method of transmission often rests on the degree of immunity to noise. The signal-to-noise ratios appearing in Table I provide a direct comparison of amplitude-modulation pulse transmission and other systems. Comments on the significance of the ratios, and the bearing on modulated pulses as an audio service for television are made later. It is clear that the amplitude-modulation pulses suffer a disadvantage in that superimposed noise cannot be reduced by amplitude limiting to the extent possible in certain other systems.

2. Width-Modulated Rectangular Pulses

A more promising form of pulse modulation, from the standpoint of signal-to-noise ratio, is a constant-amplitude pulse system wherein the width of a pulse is proportional to the amplitude of the audio signal. Two examples of width-modulated signals are illustrated in Figure 4(B) and (C). In type (1), a dissymmetrical modulation, the leading edges of the pulses occur periodically, but the widths are proportional to the instantaneous amplitude of the audio signal $M(t)$ at the instant of the leading edge: that is, only the trailing edge is "modulated." Type (2) is a symmetrical modulation, the width of a pulse being proportional to the instantaneous amplitude at the instant corresponding to the center line of the unmodulated pulse. The center lines are periodically spaced; thus, both leading and trailing edges of type (2) are modulated. Such pulse waves may be inverted in polarity and combined with the standard television wave form as shown in Figure 4(D) for type (2) modulation. Any amplitude variation of the pulse due to noise may be removed by limiting in the receiver following separation of the pulse from the video signal if the peak noise does not exceed one half of the pulse amplitude.

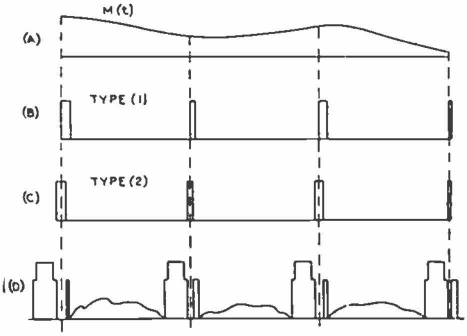


Fig. 4—Width-modulated duplex wave forms.
(A) audio modulation $M(t)$
(B) pulse carrier dissymmetrically width-modulated by $M(t)$—type (1)
(C) pulse carrier symmetrically width-modulated by $M(t)$—type (2)
(D) combination of type (2) and television wave form

Equation (28) in Appendix III is the expression for a pulse wave width-modulated in the symmetrical manner (type 2) by a sine wave. The similarity to standard frequency modulation is noticeable in the sequence of sidebands which are generated. Thus when a pulse carrier of fundamental frequency f_c is width-modulated at a rate of f_0 cycles per second, the resultant wave contains component frequencies f_c, $(f_c + f_0)$, $(f_c - f_0)$, $(f_c + 2f_0)$, $(f_c - 2f_0)$, etc., as well as corresponding sidebands for each harmonic of the fundamental f_c; namely $2f_c$, $(2f_c + f_0)$, $(2f_c - f_0)$, $(2f_c + 2f_0)$, $(2f_c - 2f_0)$, etc. In addition, the frequency terms containing only the modulating frequency f_0 and its harmonics $2f_0$, $3f_0$, etc. appear. The general term is $(Mf_c \pm Nf_0)$

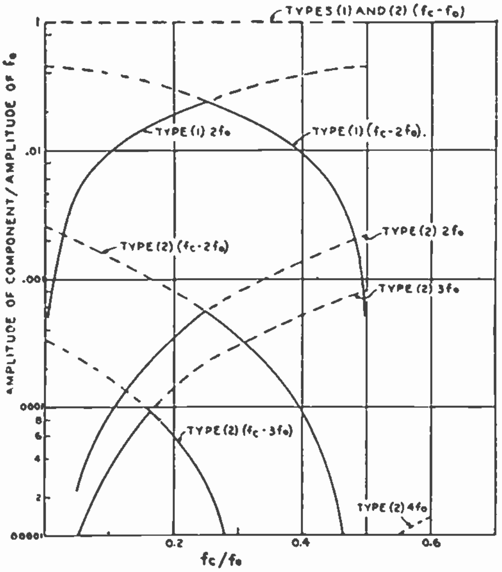
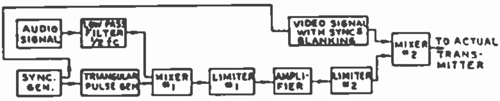


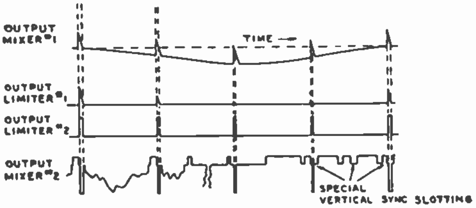
Fig. 5—Frequency components resulting from width modulation of a pulse carrier wave by a sine wave.
Type (1) = dissymmetrical modulation
Type (2) = symmetrical modulation
f_c = pulse frequency f_0 = modulating frequency

where M and N are positive integers or zero.

Since overlapping of f_0 and the sideband $(f_c - f_0)$ must be prevented, the modulating frequency should not exceed $f_c/2$. At the receiving end, the modulated pulse-signal may be applied to a low-pass filter which rejects all sidebands and harmonics exceeding one half the frequency of the fundamental f_c. However, all distortion terms are not thereby excluded; harmonics of f_0 and the sidebands of higher order $(f_c - 2f_0)$, $(f_c - 3f_0)$, may fall within the pass band. The magnitudes of the most important distortion terms in (28) have been plotted in Figure 5. The modulation constant $\propto$ was taken equal to 1, the value corresponding to maximum modulation. A value of 3 per cent was assigned to w, the unmodulated pulse width. Therefore, the widths of the pulses in the modulated wave vary from 0 to 6 per cent of the period of the carrier. This is substantially the maximum variation under the specifications given for the television synchronizing wave form (Figure 1). The broken-line portion of a curve indicates the range of the component which is suppressed by the low-pass filter in the receiver. Thus the component $(f_c - 2f_0)$ is suppressed for the range $f_0 < f_c/4$ but is transmitted when $f_0 > f_c/4$. In a reverse manner the component $2f_0$ is transmitted when $f_0 < f_c/4$ and suppressed when $f_0 > f_c/4$. The maximum value attained by either component is approximately 0.05 per cent of the amplitude of the audio component f_0.



(A) TRANSMITTER



(B) GENERATION OF WIDTH-MODULATED PULSE

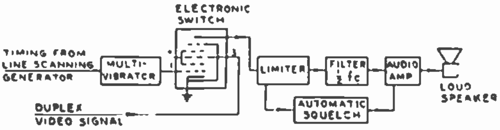
Fig. 6—Width-modulated duplex-system transmitter.

An analysis of a dissymmetrical width-modulated pulse wave (Appendix III, equation (31)) displays the same general characteristics as the symmetrical modulation. Harmonics of the audio frequency, as well as numerous sidebands, are present, but the magnitudes shown in Figure 5 are greater than in type (2) modulation. The largest contribution of any distortion term is 2.5 per cent.

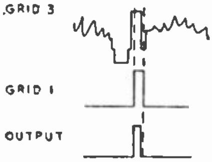
Signal-to-noise ratios calculated according to the derivations in Appendix IV appear in Table I and Figure 15.

Figure 6(A) illustrates one method for the production of dissymmetrical width-modulated pulses. The starting point is a wave of narrow triangular pulses (Figure 6(B)) which is derived from driving pulses at line frequency normally generated by the synchronizing generator. To the triangular pulses is added the audio signal from which the frequency components higher than one-half the line frequency have been removed by a low-pass filter. Limiter No. 1 removes the audio wave below the base line as shown in Figure 6(B). The residue is greatly amplified and then acted upon by limiter No. 2, with the result that width-modulated pulses of substantially rectangular shape are produced. These are inserted in the line-blanking interval following the synchronizing pulse (postblanking). The standard field synchronizing pulse must be slotted down to black level during the line pulses, as shown in Figure 6(B), in order that the width-modulated pulses when applied may extend to white level. The combined video signal is applied to the picture transmitter in the customary manner.

Figure 7(A) illustrates the functional arrangement of the receiver. The selection of the width-modulated pulses from the video signal and rejection of picture components is performed by an electronic switch, usually a vacuum tube having two control grids. A keying pulse originating in a multivibrator which is synchronized by the line-scanning circuit is impressed on grid 1 of the switch. The width and timing of the pulse is critical for the most favorable signal-to-noise ratio. The duplex video applied to grid 3 causes plate current to flow only when the tube is keyed on. In this way the width-modulated pulses are isolated.



(A) DUPLEX RECEIVER



(B) OPERATION OF ELECTRONIC SWITCH

Fig. 7—Width-modulated duplex system. Type (1) receiver.

Amplitude noise is removed by limiting if the peak noise does not exceed one-half of the pulse amplitude but the variation of the pulse width due to noise is not removable. Such variation constitutes a width modulation and is reproduced as audible noise. When the synchronization of the receiver is impaired by noise, the timing of the switch is likewise affected, and parts of blanking and picture may be admitted into the audio amplifier and appear as noise in the loud speaker. There is a marked increase in the immunity of the system to noise if the receiver is synchronized by automatic frequency control.[10]

Audio components in excess of one-half line frequency are removed by a low-pass filter.

Additional kinescope blanking must be provided in the receiver since the duplex signal extends to white level during the sound pulse. In Figure 7(A) blanking is derived from the multivibrator simultaneously with the keying pulses.

Means for excluding signal from the audio circuits when the receiver is not in synchronism is very desirable. Without such a device, video components are admitted to the audio system with an annoying audible result. A circuit may be devised which is sensitive to the changed character of the signal passed by the electronic switch during intervals of missynchronization and applies a bias beyond cutoff to the audio amplifier.

In September, 1943, television signals containing width-modulated pulses of the dissymmetrical type were transmitted by television station WNBT, and successfully received in Princeton, New Jersey, using the system outlined above.

3. Pulse Time Modulation

Another form of modulation known as "pulse time modulation" is related to width modulation.[11] In pulse time modulation the pulse amplitude and width remain constant, but the time interval between successive pulses is varied in accordance with the instantaneous amplitude of the audio signal and the rate of this variation corresponds to the instantaneous frequency of the signal. Such a pulse wave may be regarded as the sum of two width-modulated pulse waves of the dissymmetrical type of opposite polarities as illustrated in Figure 17. The frequency components of the pulse time wave are therefore solvable from (31).

4. Pulsed Frequency Modulation

In contrast with the foregoing duplex methods involving rectangular pulses for the transmission of sound during the line-blanking interval, there is a method which may be called "pulsed frequency modulation," that employs wave bursts of a frequency-modulated subcarrier for the same purpose. The bursts are generated at the trans-

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[10] K. R. Wendt and G. L. Fredendall, "Automatic frequency and phase control of synchronization in television receivers", *Proc. I.R.E.*, vol. 31, pp. 7-15; January, 1943.

[11] E. M. Deloraine and Emil Labin, "Pulse time modulation", *Elec. Commun.*, vol. 22, pp. 91-98; 1944.

mitter by a sine-wave oscillator which is operative only during line blanking and is frequency modulated by the audio signal during this interval. The center frequency and deviation are chosen so that the essential sidebands lie within the video band. These subcarrier bursts are combined with the video wave form as modulation of either the line synchronizing pulses or the post-blanking (Figure 8). From the point of view of signal-to-noise ratio, Figure 8(B) is preferable. In either case, the composite signal is applied as amplitude modulation of the radio-frequency carrier.

In the receiver, the bursts are first isolated at the video level from the picture part of the composite wave, then amplitude limited for removal of noise, and finally applied to a conventional balanced frequency-modulation discriminator centered at the frequency of the subcarrier. The output of the discriminator is an amplitude-modulated pulse wave. The audio signal is derived from the pulse output of the discriminator by removing all components in excess of one-half line frequency with a low-pass filter.

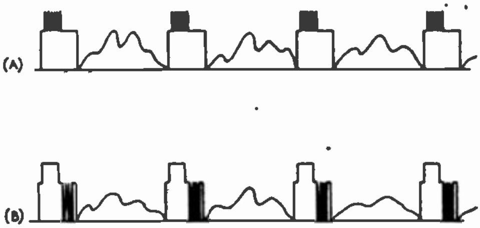


Fig. 8—Pulsed frequency-modulation duplex wave forms.
(A) in synchronizing pulse (B) in post blanking

As the result of experimental and theoretical work with pulsed frequency modulation, certain features of the technique were discovered which could escape a casual study. Such matters are treated in the following discussion.

A. Transient response of pulsed frequency-modulated circuits: If a pulsed frequency-modulated system is to function properly, the peak value of the detected pulses should depend solely on the instantaneous frequency of the subcarrier. This means that the various tuned circuits involved should complete their transients in a time which is short compared with the total duration of the wave burst. Figure 9 shows the response of a simple tuned circuit to a wave burst of constant amplitude that starts and stops with zero phase and lasts T_p seconds. The circuit has a build-up time τ, which may be adjusted by the damping resistor R and is correlated with the bandwidth b in the form

$$\tau = 2RC = 1/\pi b. \tag{2}$$

The minimum bandwidth is determined by the time allowed for the transients. If these are to be complete within p percent of the pulse time,

$$b \geqq \frac{100}{\pi p T_p}. \tag{3}$$

The total number n of cycles per pulse, as well as the number $\triangle n$ occurring before the steady state is attained, are

$$n = T_p f_s \tag{4}$$

$$\triangle n = \tau f_s \tag{5}$$

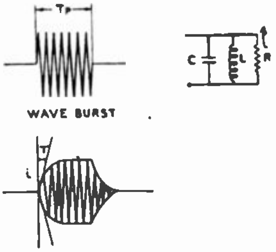


Fig. 9—Response of a tuned circuit to a pulse of frequency modulation.

Equation (5) holds regardless of the subcarrier frequency f_s. The following set of constants is representative of a typical circuit designed for pulsed frequency-modulation operation:

Pulse time T	5 microseconds
Time of build-up τ	0.5 microsecond
Circuit capacitance C	25 micromicrofarads
Circuit resistance R	10,000 ohms
Bandwidth b	400 kilocycles
Subcarrier frequency f_s	4 megacycles
Q factor	10
Cycles per pulse n	20
$\triangle n$	2 cycles

B. Generation of phasing of the subcarrier at the transmitter: According to the calculation above, a total variation of no more than

2 subcarrier cycles is sufficient to produce peak modulation in the receiver. Hence it follows that the instantaneous wave forms of the subcarrier bursts must be closely similar at the beginning. Unless the initial phase of each burst is repeated with extreme accuracy, the otherwise random initial phases may introduce audible beat notes and noise in the detected signal.

In this connection, the keying of the subcarrier for part-time modulation presents a major problem. If an independent subcarrier oscillator supplying a continuous frequency-modulated wave is used, an electronic on-off switch controlled by the main synchronizing generator must be provided. This switch cuts into the subcarrier and admits sections of its wave train for modulation of the synchronizing pulses (or blanking). It is obvious that the timing of this switch would have to be accurate within small fractions of one subcarrier cycle, or about 0.1 microsecond, in order that the initial phases of all pulses be substantially equal. The noise susceptibility of this method is high, because the subcarrier modulation is keyed on and off at full amplitude.

The problem of precise keying is further aggravated by the fact that the repetition frequency of television pulses is not constant. In all practical television synchronizing generators, the line frequency is subjected to continuous frequency control so that it constitutes, at any instant, a definite multiple of the field frequency. The field frequency is synchronized with a 60-cycle power supply which is inherently variable around a well-defined average. As a result, beat notes of variable pitch are bound to occur if a subcarrier source with continuous frequency modulation and constant center frequency is subjected to keying from a synchronizing generator unless special precautions are taken.

In the system described below, such spurious signals have been effectively eliminated. A continuous subcarrier generator is not used; instead, the bursts are supplied from a start-stop oscillator which is switched on and off by the line blanking pulses. The start-stop subcarrier oscillator shown at (4) in Figure 10 is active only when plate voltage is applied in the form of a pulse from the control tube (3). Pulses of appropriate wave shape at line frequency are derived directly from the synchronizing generator and impressed on the grid of the control tube. Hence throughout the line-scanning interval the subcarrier oscillator is inoperative, but at the end of each line it receives a plate-voltage pulse. As a result, subcarrier oscillations are built up with exactly the same initial phase conditions each time. Since the plate-power pulse is derived from the line-blanking pulse, it participates automatically in any variations of the line frequency. The power

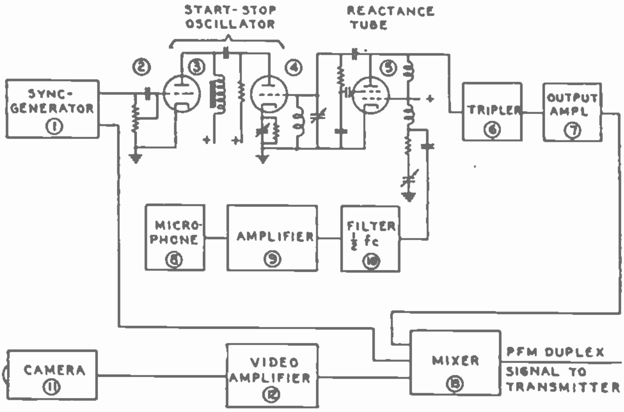


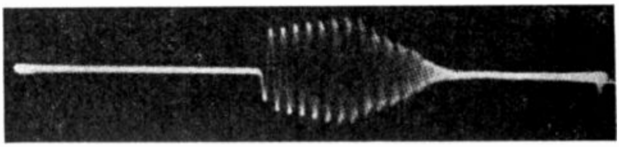
Fig. 10—Transmitter for pulsed frequency modulation.

pulses may also be preshaped in such a manner that the plate supply ceases in time to allow the subcarrier oscillations to decay within the allotted duration of sound transmission.

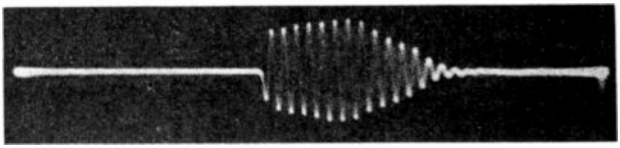
Figure 11 illustrates the subcarrier burst without and with frequency modulation.

In Figure 11(B), which shows a large number of frequency-modulated pulses in superposition, the first half of the wave burst is sharp while the wave trace appears increasingly blurred toward the end. This verifies the fact that the initial phase is substantially identical for all bursts regardless of the frequency modulation: that is, the pulse fronts are "coherent."

C. Pulsed frequency-modulation transmitter: Figure 10 shows a possible arrangement of components in a pulsed frequency-modulation



(A)
subcarrier burst



(B)
subcarrier burst, frequency-modulated
Fig. 11

transmitter. The start-stop oscillator is coupled to a reactance tube (5) which is controlled continuously by the audio signal. A low-pass filter (10) with a cutoff at one-half of the line frequency prevents the generation of overlapping sidebands of the subcarrier that would interfere with audio fidelity. In an experimental transmitter, the master oscillator generated about 10 cycles at a frequency of 2 megacycles during each burst with a deviation of ± 100 kilocycles. At the output of the doubler stage (6) the center frequency becomes 4 megacycles and the deviation ± 200 kilocycles. The subcarrier burst is amplified and combined with the video signal at (13).

From the point of view of pulsed frequency modulation the field synchronizing pulse and the equalizing pulses interfere with the regular sequence of horizontal synchronizing pulses. Some modification of the standard television wave form (Figure 1) is necessary to allow the transmission of wave bursts of constant duration. Interruptions in the sequence result in the generation of a narrow 60-cycle pulse that contains harmonics of 60 cycles extending throughout the audible spectrum.

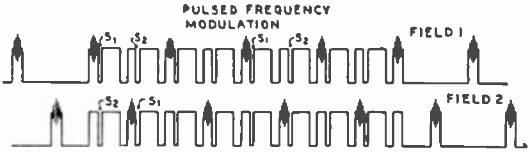


Fig. 12—Modification of television wave form for pulsed frequency modulation on line-synchronizing pulses.

If the bursts occur during postblanking (Figure 8(B)) the field-synchronizing pulse may be slotted as in Figure 6(B), but if the line-synchronizing pulses are modulated by the bursts, the modification shown in Figure 12 is desirable. Here the slots S_1 isolate the subcarrier bursts from the field signal so that separation of the sound may take place in the receiver. The slots S_2 act as equalizers for maintenance of interlacing. Figure 13 shows the modified television wave form carrying pulsed frequency-modulation duplex on the line synchronizing.

D. Pulsed frequency-modulation receiver: A complete pulsed frequency-modulation receiver is shown in Figure 14. Isolation of the frequency-modulation bursts (whether in line-synchronizing pulses as in Figure 8(A) or in post-blanking, as in Figure 8(B)) is performed by a selector such as a tube with two control grids. The selector is biased off by a suitable pulse signal generated by a multivibrator which is synchronized from the line-deflection generator or the line-synchron-

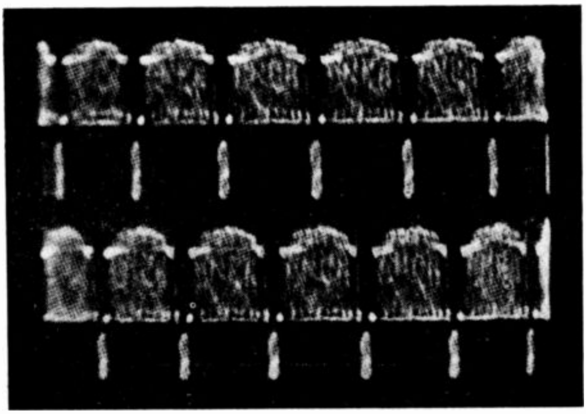


Fig. 13—Combined video signal and pulsed frequency modulation of line-synchronizing pulses.

izing circuits of the video receiver. A limiter removes the amplitude noise to substantially the same extent as in conventional frequency-modulation systems. Demodulation of the bursts is accomplished in a conventional discriminator circuit centered at the subcarrier frequency. All audio components above a frequency of one-half line frequency are removed by a low-pass filter as in the other duplex systems mentioned above. A locally generated blanking signal is required for biasing off the kinescope when the wave form of Figure 8(B) is used.

SIGNAL-TO-NOISE RATIOS OF DUPLEX AND STANDARD SYSTEMS

Formulas for the signal-to-noise ratios of duplex and standard systems are derived in Appendix IV. A comparison of the various ratios requires the assumption of a numerical relationship between the amplitudes of the respective carriers.

The usual practice in television installations is to establish the amplitude S of the standard frequency-modulated sound carrier at about 0.7 of the peak amplitude P of the picture carrier. For convenience the ratio S/P will be taken as $1/\sqrt{2}$. When amplitude modu-

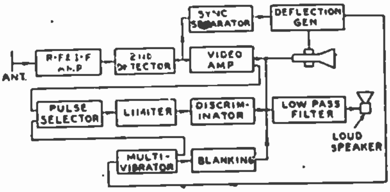


Fig. 14—Pulsed frequency-modulation receiver.

lation was standard for sound transmission prior to the adoption of frequency modulation, the same ratio $1/\sqrt{2}$ was customary.

The amplitudes in duplex transmission are fixed by the amplitude of the picture carrier.

The unmodulated amplitude h of the amplitude-modulated pulse carrier is one half the amplitude of blanking. In the standard wave form (Figure 1) blanking is three fourths of the peak amplitude of the picture carrier. Hence h may be taken as $3P/8$. The amplitude of the pulsed frequency-modulation signal during post-blanking is also $3P/8$. The amplitude H of width-modulated pulses is equal to the full amplitude of blanking or $3P/4$.

Column (1) of Table I lists the audio signal-to-noise ratios for the various methods of transmission of sound in terms of P/N and other dimensions which are associated with a particular method. P is the amplitude of the picture carrier and N is the noise factor.

Column (2) lists the critical signal-to-noise ratios below which the formulas are no longer valid. This limit exists in the case of width-modulated pulses when the peak noise is higher than one half the pulse amplitude. There is no limit in standard amplitude modulation and amplitude-modulated pulses since limiting is not applied. The limit occurs in standard frequency modulation and pulsed frequency modulation when the peak amplitudes of noise and signal are equal.

If the noise is assumed to remain constant, the signal, and therefore the signal-to-noise ratio, varies with distance from the transmitter according to the law for the propagation of television signals. Hence P/N may be replaced by K/d^2 as shown in column (3) where d is the distance from the transmitter and K is a proportionality constant. If the distance d exceeds the line-of-sight distance, a somewhat higher power of d would be appropriate.

Columns (4) and (5) show the forms taken by (2) and (3) when values are substituted.

Figure 15 illustrates the variation of the signal-to-noise ratios with distance from the transmitter. Comparisons made of the various methods of sound transmission from Figure 15 are necessarily on a relative basis since the unit of distance is $d/\sqrt{K}$.

Standard frequency modulation with a deviation of 150 kilocycles yields the most favorable signal-to-noise ratio within 0.044 units of distance. An equal ratio may be obtained over a more limited distance of 0.013 units with pulsed frequency modulation during postblanking if the maximum deviation is of the order of 1.2 megacycles. A greater deviation is required in pulsed frequency modulation for equality, because the audio signal which may be recovered from a pulsed-frequency-modulation wave is proportional to the pulse width, whereas the audio noise is proportional to the square root of the width.[12] The maximum distance from the transmitter at which limiting of a pulsed frequency-modulation signal is effective in removing noise (that is, the critical distance) is necessarily less because the noise voltage admitted to the receiver is greater as a consequence of the greater deviation.

Pulsed frequency modulation during postblanking with the customary deviation of 150 kilocycles is intermediate between standard

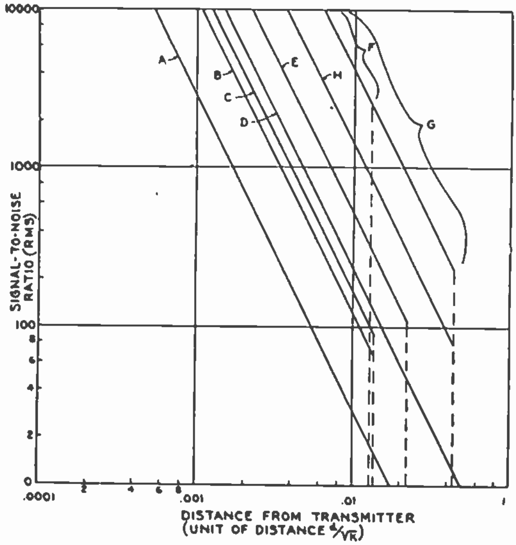


Fig. 15—Signal-to-noise ratios for sound transmission.
A = amplitude-modulated pulses
B = width-modulated pulses (symmetrical)
C = width-modulated pulses (dissymmetrical)
D = standard amplitude modulation
E = pulsed frequency modulation ($f_d = 150$ kilocycles)
F = pulsed frequency modulation ($f_d = 1200$ kilocycles)
G = standard frequency modulation ($f_d = 150$ kilocycles)
H = standard frequency modulation ($f_d = 50$ kilocycles; bandwidth = 150 kilocycles).

amplitude modulation and standard frequency modulation.

Width-modulated pulses are intermediate between amplitude-modulation pulses and standard amplitude modulation.

Amplitude-modulated pulses rank lowest, largely as a consequence of not being susceptible to limiting.

OTHER RECEIVER CONSIDERATIONS

A duplex receiver is "no better" than its sound pulse selector.

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[12] Appendix IV, equation (43).

Audible noise can be introduced into the audio system of a duplex receiver when portions of the video signal, representing picture, are selected along with the desired sound signal. This occurs when the accuracy of synchronization of the selector is reduced sufficiently by noise and interference. In this respect, the automatic frequency control of synchronization was found to be definitely superior to conventional triggered synchronization.[10] The flywheel effect of the automatic-frequency-control circuit tends to minimize the disturbing effect of noise on synchronization.

It appears that with automatic-frequency-control synchronization the major part of the total audible noise in a well-designed duplex system may be attributed to the inherent noise characteristics discussed in Appendix IV rather than to inaccurate selection of the sound signal.

The stability of duplex circuits was not studied, but it is clear that drifts in the values of circuit elements that affect the accuracy of sound selection would be detrimental.

An exhaustive study of the relative costs of a television receiver designed for duplex sound on a conventional receiver intended for reception of standard frequency modulation was not included in the scope of this project. However, an analysis of two experimental receivers constructed according to the arrangements in Figures 7(A) and 14 indicates that the cost of a commercial duplex receiver is not likely to exceed that of a standard receiver.

APPENDIX I

A rectangular-pulse wave of unit amplitude may be expressed as a cosine series

$$e(t) = r + \frac{2}{\pi}\left\{\sin \pi r \cos \omega_c t + \frac{\sin 2\pi r}{2} \cos 2\omega_c t + \cdots + \frac{\sin n\pi r}{n} \cos n\omega_c t + \cdots\right\} .$$

$$\left\{\begin{array}{l} \omega_c = 2\pi f_c \\ r = \text{pulse width per pulse period.} \end{array}\right. \qquad (6)$$

Modulation of $e(t)$ by an audio signal $M(t)$ has the result

$$f(t) = [1 + M(t)]\, e(t)$$

$$\begin{array}{ccccc} \begin{pmatrix}\text{original}\\ \text{unmodulated}\\ \text{pulse wave}\end{pmatrix} & & \begin{pmatrix}\text{audio signal}\\ \text{diminished}\\ \text{by } r\end{pmatrix} & & \begin{pmatrix}\text{sidebands of carrier}\\ \text{and its harmonics}\end{pmatrix} \\ \downarrow & & \downarrow & & \downarrow \\ = e(t) & + & rM(t) & + & \displaystyle\sum_{n=1}^{\infty} \frac{2}{\pi} M(t) \frac{\sin n\pi r}{n} \cos n\omega_c t. \end{array} \tag{7}$$

APPENDIX II

A well-known formula[13] expressing the video bandwidth required for equal horizontal and vertical resolution is

$$f_v = \tfrac{1}{2} KL^2 Na \tag{8}$$

in which

f_v = video bandwidth

L = number of scanning lines

N = frame repetition rate

a = aspect ratio

K = experimental factor often taken equal to 0.6.

Since the maximum audio frequency f_a which may be transmitted by a pulse carrier is $LN/2$, the combination of this formula and (8) gives

$$f_a = \sqrt{\frac{f_v N}{2K_a}} = K_1 \sqrt{f_v}. \tag{9}$$

Equation (9) should be regarded chiefly as an expression of proportionality between the quantities because the value of K depends upon the criterion for equal resolutions, which is not a precise concept.

APPENDIX III

FREQUENCY COMPONENTS[13] RESULTING FROM SYMMETRICAL WIDTH MODULATION OF A RECTANGULAR-PULSE CARRIER BY A SINE WAVE

The problem is the calculation of the amplitude and frequency of each component in a rectangular-pulse carrier which is width-modulated in a symmetrical manner by a sine wave. The width of a pulse is proportional to the amplitude of the modulating wave at the instant corresponding to the center line of the pulse. Hence, the width a_p of

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[13] R. D. Kell, A. V. Bedford, and M. A. Trainer, "An experimental television system, Part II. The transmitter", *Proc. I.R.E.*, vol. 22, pp. 1246-1266; November, 1934.

the pth pulse is

$$a_p = w\left(1 - \alpha \cos 2\pi p \frac{T_c}{T_o}\right) \tag{10}$$

in which

w = width of unmodulated pulse

$T_c = 1/f_c$ = period of pulse wave

$T_o = 1/f_0$ = period of modulating wave

α = modulation factor.

In the general case, the modulated carrier wave will not repeat precisely at the end of each audio cycle, but after some time greater than the period T_0 there will be repetition. Let this time be called T and the corresponding frequency, f. The equation of the modulated pulse wave may be deduced by regarding the wave as the summation of a large number of pulse waves of equal period T. Each component wave will be characterized by a certain pulse width which is constant for the particular component. Thus there is a wave starting at the origin and characterized by a pulse width a_0, a wave of width a_1 and phase T_c, a wave of width a_2 and phase $2T_c$, etc.

The pth wave has a width a_p and phase pT_c. There are $(f_cT - 1)$ waves to sum. The equation of the pth wave is

$$e_p = \left[\frac{a_p}{T} + \frac{2}{\pi}\sum_{n=1}^{\infty}\frac{\sin n\pi a_p f}{n}\cos 2\pi nf(t - pT_c)\right]. \tag{11}$$

A summation over p yields the equation of the modulated pulse wave.

$$\begin{aligned} e' &= \sum_{p=0}^{f_cT-1}\frac{a_p}{T} + \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\sum_{n=1}^{\infty}\frac{\sin n\pi a_p f}{n}\cos 2\pi nf(t - pT_c) \\ &= e_{\text{d.c.}} + e. \end{aligned} \tag{12}$$

Since the direct-current component is not of interest, it need not be considered further. If a certain frequency component of e is sought, the contributions of each of the p waves must be summed in the form

$$\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{\sin n\pi a_p f}{n}\cos 2\pi fn(t-pT_c). \tag{13}$$

The amplitude of the component (fn) is $\sqrt{A_1^2+B_1^2}$ in which

$$A_1=\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{\sin(n\pi a_p f)}{n}\cos 2\pi fnpT_c \tag{14}$$

$$B_1=\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{\sin(n\pi a_p f)}{n}\sin 2\pi fnpT_c. \tag{15}$$

The remainder of the derivation is devoted to an examination of A_1 and B_1. It is expected that only certain values of n will lead to non-zero values for A_1 and B_1. Before summation, the expression for a_p is inserted in A_1 and B_1. There results

$$A_1=\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\sin\left[2n\pi f\epsilon\left(1-\propto\cos 2\pi T_cf_0p\right)\right]\cos 2\pi fnpT_c \tag{16}$$

$$B_1=\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\sin\left[2n\pi f\epsilon\left(1-\propto\cos 2\pi T_cf_0p\right)\right]\sin 2\pi fnpT_c. \tag{17}$$

Expansion of A_1, in terms of Bessel functions, yields

$$A_1=\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\Bigg[\cos Rp\sin\sum_{s(\text{even})=2}^{\infty}2(-1)^{s/2}J_s(B)\cos sC_p+J_0(B)$$

$$+\cos Rp\cos\cdot A\sum_{v(\text{odd})=1}^{\infty}2(-1)^{(v+1)/2}J_v(B)\cos vC_p\Bigg] \tag{18}$$

and

$$B_1=\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\Bigg[\sin Rp\sin A\sum_{s(\text{even})=2}^{\infty}2(-1)^{s/2}J_s(B)\cos sC_p$$

$$+\sin Rp\cos A\sum_{v(\text{odd})=1}^{\infty}2(-1)^{(v+1)/2}J_v\cos vC_p\Bigg] \tag{19}$$

in which

$$\begin{aligned} \pi n f w &= A \qquad & 2\pi T_c f_c &= C \\ n\pi w \propto f &= B \qquad & 2\pi f n T_c &= R. \end{aligned} \tag{20}$$

A typical term in A_1 involving the summation over s is

$$\frac{4(-1)^{s/2}}{\pi n} J_s \sin A \sum_{p=0}^{f_cT-1} \cos Rp \cos sCp. \tag{21}$$

The expression

$$\sum_{p=0}^{f_cT-1} \cos Rp \cos sCp \tag{22}$$

is a finite trigonometric sum which is known to have the value

$$\frac{1}{2}\,\frac{\cos\left[(f_cT-1)(R-sC)/2\right]\sin\left[f_cT(R-sC)/2\right]}{\sin\left[(R-sC)/2\right]} + \frac{1}{2}\,\frac{\cos\left[(f_cT-1)(R+sC)/2\right]\sin\left[f_cT(R+sC)/2\right]}{\sin\left[(R+sC)/2\right]}\cdot \tag{23}$$

The above sum may be abbreviated

$$S = \frac{S_1}{2} + \frac{S_2}{2}\cdot \tag{24}$$

If the expressions for R and C in (20) are introduced, S_1 in (24) becomes

$$S_1 = \frac{\cos\pi\left[sT_cf_0 - nT_cf + n - sTf_0\right]\sin\pi\left[n - sTf_0\right]}{n\sin\pi\left(nT_cf - sT_cf_0\right)}\cdot \tag{25}$$

If S_1 is to have a nonzero solution, the denominator must be zero at least for some values of n. This follows from the observation that $\sin\pi\,(n - sTf_0)$ is always zero. From inspection, it is seen that the denominator of (25) is zero when

$$n = \pm MTf_c + sTf_0 \tag{26}$$

in which M is the positive integer. When (26) is inserted in (25), the indeterminancy may be reduced to

$$S_1 = \frac{1}{sT_c f_0 \pm M} \cdot \tag{27}$$

Similar reasoning leads to explicit forms for A_1 and B_1 in (14) and (15). Finally, (12), for the modulated wave, may be written in the form

$$e = \frac{2}{\pi} \sum_{M=0}^{\infty} \left[\sum_{\nu=-\infty}^{\infty} \{ J_{|\nu|} [\pi w \propto (Mf_c + \nu f_0)] \} \frac{\sin [\pi w (Mf_c + \nu f_0) - |\nu| \pi/2]}{M + \nu f_0/f_c} \cos 2\pi (Mf_c + \nu f_0) t \right] \tag{28}$$

FREQUENCY COMPONENTS RESULTING FROM DISSYMMETRICAL WIDTH MODULATION OF A RECTANGULAR-PULSE CARRIER BY A SINE WAVE

When each pulse of a symmetrically-modulated pulse carrier is translated to the right (or left) on the time axis by an amount equal to one half the width of the modulated pulse, the carrier becomes unsymmetrically modulated. Therefore (12) may be modified to read

$$e' = \sum_{p=0}^{f_cT-1} \frac{a_p}{T} + \frac{2}{\pi} \sum_{p=0}^{f_cT-1} \sum_{n=1}^{\infty} \frac{\sin n\pi a_p f}{n} \cos 2\pi n f \left[t - pT_c - \frac{a_p}{2} \right] \tag{29}$$

in which

$$a_p = w(1 - \propto \cos 2\pi p T_c f_0). \tag{30}$$

A mathematical process similar to that outlined in Appendix II yields the result

$$\begin{aligned} e = & \sum_{\nu(\mathrm{odd})=1}^{\infty} \frac{1}{\pi} (-1)^{(\nu+1)/2} \frac{J_\nu [2\pi w \propto (Mf_c + \nu f_0)]}{M + \nu f_0/f_c} \cos 2\pi [(Mf_c + \nu f_0)(t - w)] \\ & + \sum_{\nu(\mathrm{odd})=1}^{\infty} \frac{1}{\pi} (-1)^{(\nu+1)/2} \frac{J_\nu [2\pi w \propto (Mf_c - \nu f_0)]}{M - \nu f_0/f_c} \cos 2\pi [(Mf_c - \nu f_0)(t - w)] \\ & + \frac{2}{\pi} \sum_{\xi}^{\infty} (-1)^{(p+\beta-2)/2} \frac{J_p [\pi w \propto (Mf_c + \xi f_0)] J_\beta [2\pi \epsilon \propto (M + \xi f_0)]}{M + \xi f_0/f_c} \sin 2\pi [(Mf_c + \xi f_0)(t - w)] \\ & + \frac{2}{M\pi} \sin (\pi M f_c w) \cos 2\pi f_c M (t - w/2) \end{aligned} \tag{31}$$

$$\xi = (\rho + \beta),\ (\rho - \beta),\ (-\rho + \beta),\ (-\rho - \beta)$$

ρ and β are odd positive integers.

Other symbols have the same significance previously assigned.

APPENDIX IV

SIGNAL-TO-NOISE RATIOS

1. *Standard Amplitude Modulation and Standard Frequency Modulation*

The root-mean-square signal-to-noise ratio for 100 per cent amplitude modulation is

$$\frac{S}{\sqrt{2}} \Big/ \frac{N\sqrt{f_a}}{4} = 2\sqrt{2}\,\frac{S}{N\sqrt{f_a}} \tag{32}$$

in which

$S =$ unmodulated amplitude of carrier

$f_a =$ highest audio frequency

$N\sqrt{f_a} =$ peak amplitude of noise

$(= 4 \times$ root-mean-square noise$)$[14]

Crosby[15] and others have shown that

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{FM}} = \frac{\sqrt{3}f_d}{2f_a}\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{AM}} \tag{33}$$

in which f_d is 2 times frequency deviation. The amplitude of noise is assumed to be below the threshold value.

From (32) and (33),

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{FM}} = \frac{\sqrt{6}\,S\,\sqrt{f_d}}{Nf_a}\sqrt{\frac{f_d}{f_a}}\,. \tag{34}$$

2. *Amplitude-Modulated Pulses*

In a 100 per cent amplitude-modulated pulse system the root-mean-square value of the audio signal detected by means of a low-pass filter

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[14] Vernon D. Landon, "The distribution of amplitude with time in fluctuation noise", *Proc. I.R.E.*, vol. 29, pp. 50-55; February, 1941.

[15] Murray G. Crosby, "Frequency-modulation noise characteristics", *Proc. I.R.E.*, vol. 25, pp. 472-517; April, 1937.

in the receiver (see Appendix I) is

$$\frac{rh}{\sqrt{2}}. \tag{35}$$

in which

h = unmodulated height of the pulse

r = ratio of pulse width to period of pulse carrier.

If the assumption is made that the pulse wave is applied to the detector (low-pass filter) only during the time of the pulses, the root-mean-square noise is

$$\sqrt{r}\,\frac{N\sqrt{f_a}}{4}. \tag{36}$$

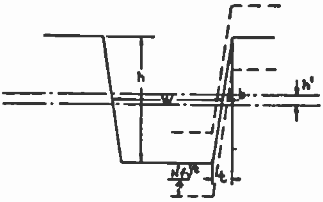

Fig. 16—Width-modulated pulses.

Hence

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{AM pulses}} = 2\sqrt{2}\,\frac{h\sqrt{r}}{N\sqrt{f_a}}.$$

3. *Width-Modulated Pulses*

Before detection, the noisy signal is clipped, or limited, top and bottom so that only a comparatively narrow section near the center of each pulse is selected. Hence, it is assumed that noise effects are introduced into width-modulated pulses chiefly by the random displacement of the sides of the pulses. The following derivation applies when the peak noise is less than one half of the amplitude of the pulses. In Figure 16, let

w = unmodulated width of pulse

H = height of pulse

t_s = time of rise of pulse side

b = displacement of side in seconds due to a root-mean-square noise voltage

f_v = video-frequency bandwidth.

Then the slope of the side of a pulse is

$$\frac{H}{t_s} = \frac{N\sqrt{f_v}}{4} \Big/ b \tag{37}$$

from which

$$b = \frac{N\sqrt{f_v}}{4} \frac{t_s}{H} \,. \tag{38}$$

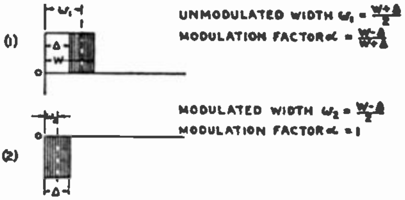

Fig. 17—Decomposition of a pulse time wave into width-modulated pulses (1) and (2) of the dissymmetrical type. Shaded lines intended to simulate an oscillogram.

The audio signal recovered from the pulse wave, whether symmetrically or dissymmetrically modulated, is

$$\frac{CwH'}{\sqrt{2}} \,. \tag{39}$$

In this expression, C is a proportional constant. H' is the new height of the pulse after the center section of the pulse has been selected out and the remainder rejected (as shown in Figure 16). Amplitude due to the random displacement of one side of a pulse in dissymmetrical modulation is

$$CbH'. \tag{40}$$

Hence the signal-to-noise ratio is

$$\frac{CwH'}{\sqrt{2}} \Big/ Cb\,H' = \frac{2\sqrt{2}\,wH}{t_s N\sqrt{f_v}}\,. \tag{41}$$

In symmetrical modulation, both sides of a pulse are subject to random displacement due to noise. The noise voltage given in (40) must therefore be multiplied by $\sqrt{2}$. The signal-to-noise ratio for symmetrical modulation is therefore

$$\frac{2wH}{t_s N\sqrt{f_v}}\,. \tag{42}$$

4. *Pulses of Frequency Modulation*

The audio signal which may be recovered from a keyed frequency-modulated wave by means of a discriminator followed by a low-pass filter (Figure 14) is proportional to the pulse width. However, the noise appearing in the audio output is proportional to the square root of the width. Hence

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{P—F—M}} = \sqrt{r}\,\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{standard F—M}}\,. \tag{43}$$

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

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* 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

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[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 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.

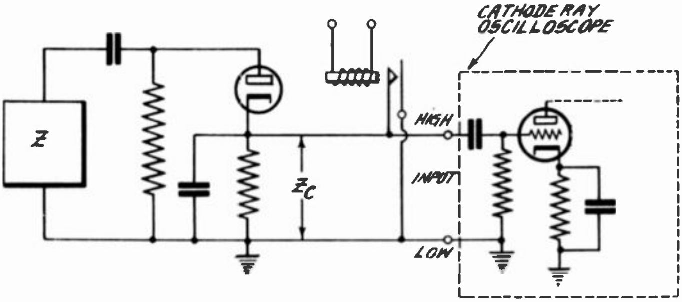


Fig. 1—Diode rectifier, "Vibroswitch," and cathode ray oscilloscope circuit arrangement.

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 Figure 3. The fundamental problem insofar as the "Vibroswitch" is concerned is to obtain a short closed contact period with clean make and break. Most of the earlier models suffered from mechanical oscillation of the swinger, causing variation in contact resistance at the

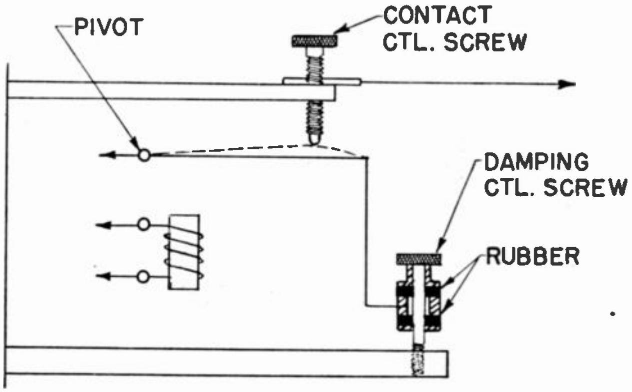


Fig. 2—Mechanical schematic diagram of the "Vibroswitch."

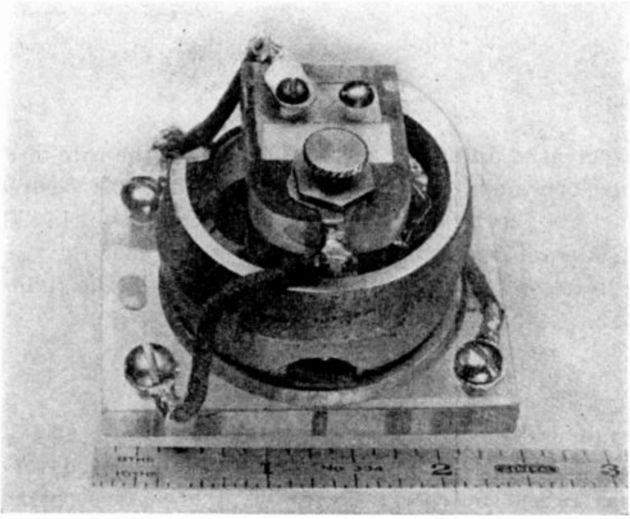
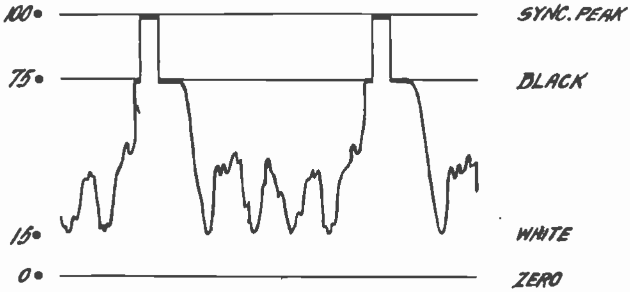


Fig. 3—A recent physical form of the "Vibroswitch."

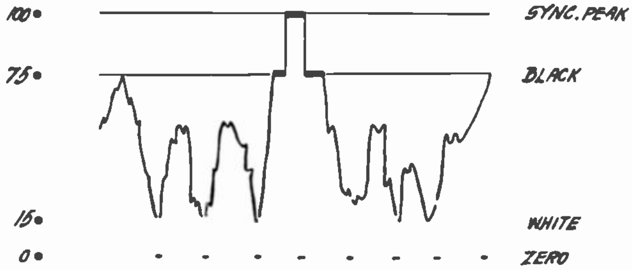
instant that the contact was closed. This led to a confused trace on the oscilloscope.

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 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



(a) Horizontal deflection rate approximately one half the field repetition rate.



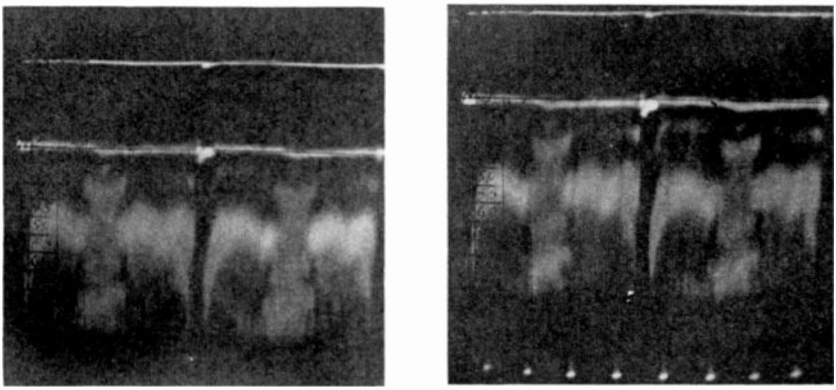
(b) Horizontal deflection rate approximately one half the line repetition rate.

Fig. 4—Representation of the expected Oscilloscope Trace:

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 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.



(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:

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.

FACTORS GOVERNING PERFORMANCE OF ELECTRON GUNS IN TELEVISION CATHODE-RAY TUBES*†

BY

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Summary—*On the basis of Langmuir's*[1] *limiting-current-density relationship, it is shown that the useful beam current in a convential television cathode-ray tube has an upper limit defined by*

$$I_s = 1.13\, i_0 E_s \frac{e}{kT} \frac{A^2}{N^2} \tan^2\Phi$$

where

I_s = *the beam current;*
i_0 = *the cathode-current density;*
E_s = *the second-anode voltage relative to the cathode;*
e = *the electron charge;*
k = *Boltzmann's constant;*
T = *the absolute temperature of the cathode;*
A = *the aperture of the final focusing system;*
N = *the number of scanning lines; and,*
Φ = *the equivalent deflection angle.*

This result is derived for the case of an ideal electron gun with no defining apertures. In practice this upper limit of beam current is not attained because of aberrations and space-charge mutual repulsion effects.

INTRODUCTION

IN DISCUSSIONS of the performance of electron guns in television cathode-ray tubes, questions frequently arise as to what will be the effect of changing this or that parameter. For example, such questions are asked as: how does the brightness of the picture depend upon the resolution?; does wide-angle deflection offer other advantages than reduction in tube length?; or, what will be the effect of increasing the operating voltage? Although the answers to these and many other questions may be derived from the fundamental principles of electron

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* Decimal classification: R583.6 × R138.3.

† Reprinted from *Proc. I.R.E.*, February, 1942.

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

[1] D. B. Langmuir, "Theoretical limitations of cathode-ray tubes," *Proc. I.R.E.*, vol. 25, pp. 977–991; August, 1937.

optics,[1–4] there is need for a simple, easily interpreted relationship correlating the various factors governing electron-gun performance. It is the purpose of this paper to present such a relationship.

THEORETICAL ANALYSIS

To formulate the problem, consider a conventional[5] electron gun of the form illustrated schematically in Figure 1. This device operates in the following manner. First, the cathode-region or first-crossover-forming lens L_1 concentrates the electron beam into a small diameter at a first crossover. Second, the electrons emerging from this first crossover are refocused to a small spot on the fluorescent screen by the final focusing lens L_2. The electron beam so formed is deflected by electrostatic or electromagnetic means to trace out the picture raster.

What factors determine the performance of this device? The analysis will be facilitated by the use of suitable nomenclature and symbols. Let

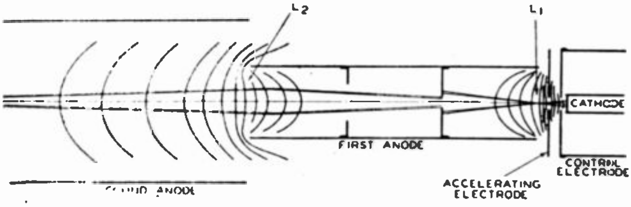

Fig. 1—Schematic representation of a conventional electron gun.

H = picture height

s = scanning-spot diameter

N = number of scanning lines to be resolved

ϕ = equivalent deflection angle

A = aperture of final focusing lens

θ = half angle of beam spread

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[2] L. Jacob, "Electron distribution in electron-optically-focused electron beams," *Phil. Mag.*, vol. 28, pp. 81–98; July, 1939.

[3] E. G. Ramberg and G. A. Morton, "Electron optics," *Jour. Appl. Phys.*, vol. 10, pp. 465–478; July, 1939.

[4] J. R. Pierce, "Limiting current densities in electron beams," *Jour. Appl. Phys.*, vol. 10, pp. 715-724; October, 1939.

[5] V. K. Zworykin, "Description of an experimental television system and kinescope," *Proc. I.R.E.*, vol. 21, pp. 1655–1673; December, 1933.

a = first-crossover to final-focusing-lens distance

b = final-focusing-lens to screen distance

d = first-crossover diameter

D = cathode diameter

i_0 = cathode-current density

i_r = first-crossover current density

E_1 = first-anode potential

E_2 = second-anode potential

e = electronic charge

k = Boltzmann's constant

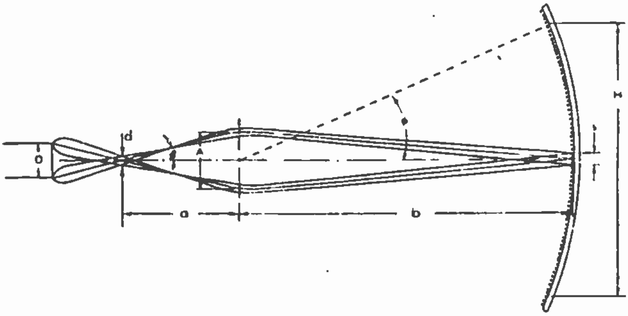

Fig. 2—Schematic representation of geometric factors determining performance of the electron gun in television cathode-ray tubes.

T = absolute temperature of the cathode

The definitions of symbols having to do with the geometric configuration of the structure are further clarified in the schematic drawing of Figure 2.

Consider the performance of this device. By definition, if the picture height is H and the number of scanning lines to be resolved is N (the meaning of resolution will be amplified later on when the question of light distribution across a beam trace is examined), the effective diameter of the scanning spot may be stated as

$$s = H/N \tag{1}$$

This scanning spot will be an image of the first crossover. If space charge is neglected, and the familiar[3] electron-optical magnification formula is applied, an ideal electron gun would produce such a scanning spot from a first crossover which had an effective diameter of

$$d=\frac{H}{N}\sqrt{\frac{E_2}{E_1}}\,\frac{a}{b}\,. \tag{2}$$

The distances between gun and screen and between first crossover and final focusing lens may be expressed in terms of the equivalent deflection angle, the aperture of the final focusing lens, and the spread of the beam as it enters the final focusing lens. If ϕ is the equivalent deflection angle,

$$b=\frac{H}{2\tan\phi}\,. \tag{3}$$

If θ is the half angle of beam spread and A is the aperture of the final focusing lens,

$$a=\frac{A}{2\tan\theta}\,. \tag{4}$$

Equation (2) may then be written

$$d=\frac{A}{N}\sqrt{\frac{E_2}{E_1}}\,\frac{\tan\phi}{\tan\theta}\,. \tag{5}$$

How much beam current may be concentrated into a crossover of given size? What will be the light distribution across a beam trace when the scanning spot is an image of this first crossover? In terms of the present nomenclature, Langmuir[1] has shown that the current density in a crossover in an ideal electron optical system is

$$i_r=i_0\sin^2\theta\left\{1+E\frac{e}{kT}\left[1+\frac{\dfrac{d^2}{D^2}\sin^2\theta}{1-\dfrac{d^2}{D^2}\sin^2\theta}\right]\right\}\epsilon^{-E\frac{e}{kT}\frac{\frac{d^2}{D^2}\sin^2\theta}{1-\frac{d^2}{D^2}\sin^2\theta}}\,. \tag{6}$$

In general $1 << E(e/kT)$, and $(d^2/D^2)\sin^2\theta << 1$. If the value of d given by (5) is substituted in this expression and it is remembered that $\sin\theta \approx \tan\theta$ for the angles commonly encountered, integration between the limits 0 and $d/2$ gives

$$I = \frac{\pi D^2}{4} i_0 \left(1 - \epsilon^{-E\frac{e}{kT}\frac{A^2}{N^2D^2}\tan^2\phi} \right) \tag{7}$$

where I is the current within a spot of effective size H/N. But $\pi D^2 i_0/4 = I_s$, where I_s is the total beam current if the system contains no limiting apertures. Equation (7) may, therefore, be written

$$\frac{I}{I_s} = 1 - \epsilon^{-E\frac{e}{kT}\frac{A^2}{N^2D^2}\tan^2\phi} \tag{8}$$

This result warrants further examination. I/I_s is the ratio of the beam current within a particular zone to the total beam current. This zone is to be of such width as to give the resolution N. But how shall resolution be defined? In the absence of defining apertures, the spot has no definite boundary, and irrespective of the spacing, the scanning lines must overlap to a certain extent. For a given resolution, how much may they overlap? To answer these questions, it is necessary to know the brightness distribution across a beam trace.

The author[6] has shown that (6) may be expressed in the form

$$i_r = C_1 \epsilon^{-B\frac{d^2}{4}}. \tag{9}$$

To determine the brightness distribution across an individual beam trace, let x be the co-ordinate expressing distance from the center of the spot perpendicular to the direction of scanning, and let y be the co-ordinate expressing distance from the center of the spot in the direction of scanning. The excitation occasioned by a single beam trace will be

$$\text{excitation} = C_2 \epsilon^{-Bx^2} \int_{-\infty}^{+\infty} \epsilon^{-By^2}\, dy = C_3\, \epsilon^{-Bx^2}. \tag{10}$$

If the light output of the phosphor is directly proportional to the excitation, this equation represents the brightness distribution also. In terms of a given resolution, how much may these traces overlap? Can the degree of overlap be expressed in terms of the ratio I/I_s?

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[6] R. R. Law, "High current electron gun for projection kinescopes," *Proc. I.R.E.*, vol. 25, pp. 954–976; August, 1937.

Figure 3 shows the resultant brightness distribution in a reproduction of a portion of a scene containing relatively large adjacent white and dark areas. The brightness distribution is shown for three values of the ratio I/I_s. When $I/I_s = 0.5$, the reproduction is substantially equivalent to that which would be obtained from a cosine-squared distribution having the same maximum value and a total width equal to twice the spacing between adjacent lines. Under these conditions the cosine-squared distribution would give a flat field.[7,8] Although the present exponential function does not give a flat field, the practical limiting resolution may be said to occur when the spot size is such that the brightness of the beam trace drops to one half its maximum value in one half the distance between the centers of adjacent scanning lines. As may be seen from Figure 3, this condition is satisfied when $I/I_s = 0.5$. But if the degree of overlap at which N lines may just be resolved is defined by the condition $I/I_s = 0.5$, (8) gives

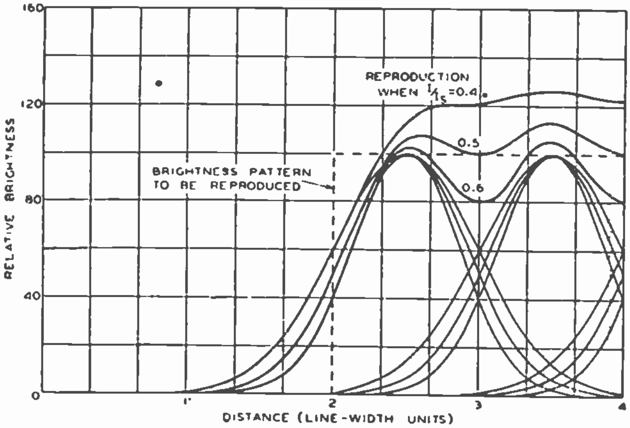

Fig. 3—Fidelity of reproduction as a function of degree of overlap.

$$I_s = 1.13 i_0 E_2 \frac{e}{kT} \frac{A^2}{N^2} \tan^2 \phi. \tag{11}$$

I_s, therefore, represents an upper limit to the useful beam current that may be obtained with an ideal electron gun having no defining apertures. With suitable defining apertures the useful beam current

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[7] P. Mertz and F. Gray, "A theory of scanning," *Bell Sys. Tech. Jour.*, vol. 13, pp. 464–515; July, 1934.

[8] H. A. Wheeler and A. V. Loughren, "The fine structure of television images," *Proc. I.R.E.*, vol. 26, pp. 540–575; May, 1938.

may be increased. This comes about because the peak current density may be maintained over the entire spot. For example, with a circular spot of uniform intensity overlapping to such an extent as to give a flat field as before, the beam current may be increased by the factor $\pi/1.13$. Inasmuch as aberrations and space-charge mutual repulsion effects operate to increase spot size, these values will not be realized in practice. For any given structure, the ratio of the measured beam current to the computed limiting value affords a figure of merit describing the performance of the gun. This ratio is ordinarily about one tenth,[2] but by minimizing the effects of space charge in the first-crossover-forming region, Pierce[9] has obtained current densities of over one half the limiting value.

DISCUSSION

By virtue of this simple relationship, describing the performance of an ideal electron gun in a television cathode-ray tube, it is now possible to answer the original questions as to what will be the effect of changing this or that parameter. Thus, if a linear relationship between picture brightness and the product of beam current is assumed, the picture brightness will vary inversely as the square of the number of lines, and directly with the square of the voltage. Wide-angle deflection does offer other advantages than reduction in tube length provided deflection introduces no defocusing; this analysis indicates that the picture brightness should vary directly with the square of the tangent of the equivalent deflection angle. In addition to correlating the various factors governing electron-gun performance, this relationship may prove useful in evaluating the performance of developmental models of electron guns by affording a direct means of computing the ideal performance of the particular structure.

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[9] J. R. Pierce, "Rectilinear electron flow in beams," *Jour. Appl. Phys.*, vol. 11, pp. 548–554; August, 1940.

TELEVISION RECEPTION WITH BUILT-IN ANTENNAS FOR HORIZONTALLY AND VERTICALLY POLARIZED WAVES*†

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

Summary—Television antennas suitable for mounting within a console receiving cabinet are described. A small loaded dipole was found to be more sensitive than a loop of equal size.

Data are given for reception in buildings on receivers with built-in antennas. Reflections caused standing waves, which affected reception of both horizontally and vertically polarized waves. The presence of people near the receiver had the most effect on the signal strength received when vertically polarized waves were utilized. Good reception in steel-frame buildings was limited to the side of the building having an unobstructed path to the transmitter. Normal obstructions in the vicinity of the antenna, such as might be encountered in residential locations, were found to attenuate vertically polarized waves more than horizontally polarized waves.

A field survey of wave propagation through normal city obstructions is recorded. A close agreement with theoretical open-country propagation characteristics was obtained.

THE loop antenna enjoyed a few years of popularity in the early days of broadcasting, but was later discarded in favor of the better performing outdoor antenna. Recently, changed listening habits of the public, higher-power broadcast stations, technical improvements in receivers, and other factors have contributed to the revival of the built-in loop antenna for standard-broadcast reception.

It is natural to ask whether the future trend of the television receiving antenna will follow the history of the standard-broadcast antenna. It seems likely that the popularity of the built-in antenna for standard broadcast will stimulate a demand for a built-in antenna for television. Factors related to this question such as the propagation of ultra-high-frequency waves through buildings and their reception on small antennas have been recently investigated. The results obtained are reported in this paper.

Before taking up these results, it may be well to review the work of others which seems most pertinent to the subject. It has been shown by Trevor and Carter[1], Norton[2], and Brown[3] that for outdoor reception free from obstructions and at ultra-high frequencies such as 50 megacycles, the field strength near the ground is substantially stronger for

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* Decimal Classification: R583.7 × R326.6.

† Reprinted from *RCA REVIEW*, April, 1942.

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

[1] Trevor and Carter, "Notes on Propagation of Waves Below Ten Meters in Length," *Proc. I.R.E.*, March, 1933.

[2] K. A. Norton, "Statement on Ultra-High-Frequency Propagation," *Television Hearing Before FCC*, Jan. 15, 1940.

vertically polarized waves than for horizontally polarized waves. Data are presented in the present paper which show substantially the same relative response, as found by the above mentioned investigators, for waves received at an outdoor location free from nearby obstructions after having been propagated through low buildings such as are found in a city residential district. Brown[3] further shows that as the receiving antenna is raised approximately 30 feet above ground, the two types of polarization yield practically identical field intensities, when the transmitting antenna is at least one wavelength above ground. Also the usual radio-noise fields in the ultra-high-frequency range are stronger in the vertical than in the horizontal plane. Therefore, in spite of the preponderance of vertically polarized field near the surface of the earth, horizontally polarized waves yield a more favorable signal-to-noise ratio for television and aural broadcast services (between 30 and 100 megacycles) where the transmitting antenna is at least a few wavelengths above ground level.

Wickizer[4] found 4.3 db higher average field strength for horizontally than for vertically polarized waves during a survey along highways with the receiving antenna 10 feet above ground. This can be explained by assuming that the normal obstructions encountered along the highway attenuated vertically polarized waves more than horizontally polarized waves. Englund, Crawford, and Mumford[5] showed that trees along the roadside absorbed and reflected vertically polarized waves. Data are presented in this present paper which indicate that trees do not materially affect horizontally polarized waves at 69 megacycles.

Jones[6] showed field-strength contours within a dwelling with reception of vertically polarized waves. Data are presented in this present paper which indicate that wood frame houses interfere more with vertically polarized waves than with horizontally polarized waves.

ANTENNA DESIGNS

A television receiving antenna, confined within a console cabinet, may be directional with means for orienting its reception characteristics or it may be nondirectional. A vertical loop may be employed as a bi-directional antenna for reception of vertically polarized waves or a horizontal dipole may be employed for bi-directional reception of

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[3] G. H. Brown, "Vertical versus Horizontal Polarization," *Electronics*, Oct., 1940.

[4] G. S. Wickizer, "Mobile Field Strength Recordings of 49.5, 83.5, and 142 Mc from Empire State Bldg. Horizontal and Vertical Polarization," *RCA REVIEW*, April, 1940.

[5] Englund, Crawford, and Mumford, "Some Results of a Study of Ultra-Short Wave Transmission Phenomena," *Proc. I.R.E.*, March, 1933.

[6] L. F. Jones, "A Study of the Propagation of Wavelengths Between Three and Eight Meters," *Proc. I.R.E.*, March, 1933.

horizontally polarized waves. For nondirectional reception a vertical dipole or a capacitive element terminated through a coupling inductance to chassis ground may be employed for vertically polarized waves or a horizontal loop or folded dipole may be employed for nondirectional reception of horizontally polarized waves.

A directional built-in antenna with means for rotating it can be employed to discriminate against interference, including undesired reflections. The nondirectional type of built-in antenna is less expensive and usually will occupy less cabinet space.

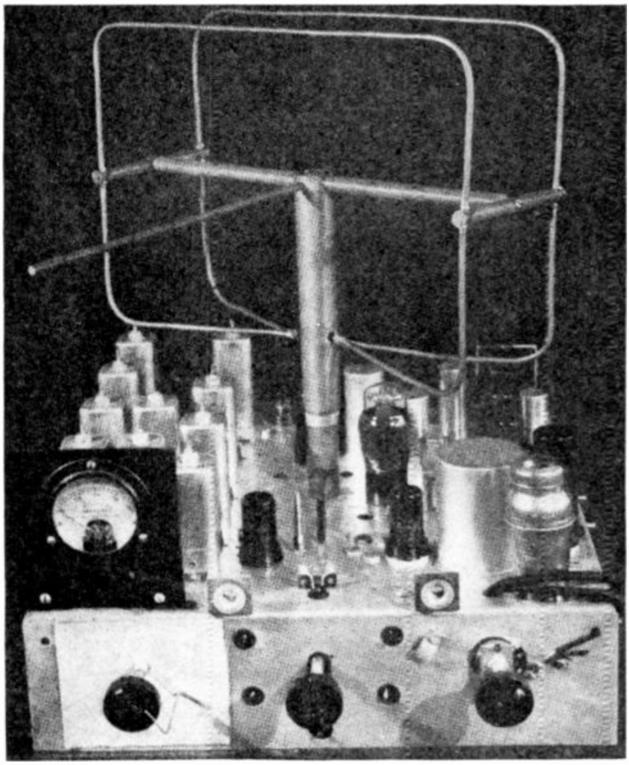
Fig. 1

Figures 1 and 2 are photographs of two experimental types of built-in antennas which were adapted to the RCA TRK-120 television-receiver chassis. Figure 1 shows a vertical loop-type antenna. The two turns are in parallel and are connected to an inductor through a wave-change switch. The antenna circuit functions as a full-wave resonant circuit and is coupled to a conventional, resonant grid circuit. The circuits are designed to give a band-pass characteristic of about 5 megacycles width. Figure 2 shows a horizontal dipole with end-capacitance load. It connects to an inductor which couples to a resonant grid circuit as in the case of the loop design. These antennas both have a figure-eight reception pattern in the horizontal plane. Both are rotatable about a vertical axis. The loop is 10 by 14½ inches. The dipole ends are 8½ inches square and are separated by 12 inches. The same cabinet space will accommodate either antenna. Provision is also made through a wave-change switch for operating the sets on conventional antennas through a transmission line.

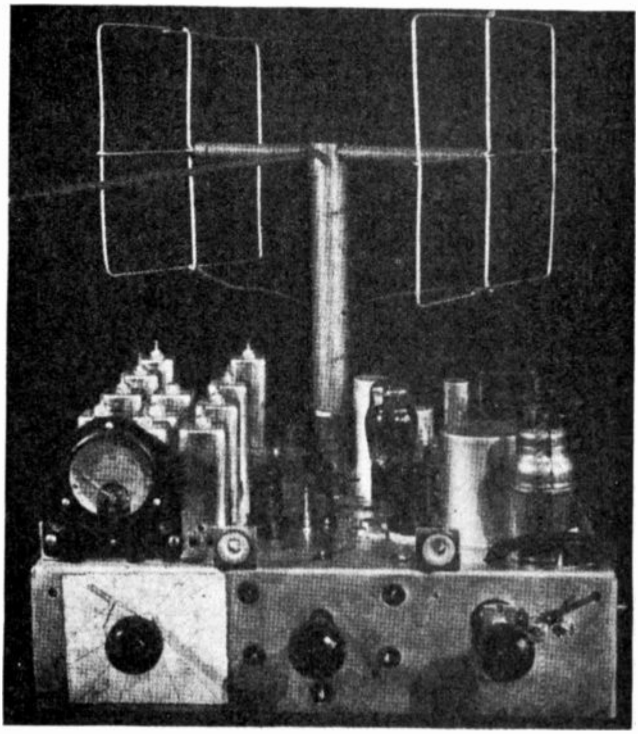

Fig. 2

The relative sensitivity of these antennas and of a half-wave dipole connected to the receiver through a short transmission line of negligible loss is given below. The measurements were made at 69 Mc in an open field with horizontal-wave polarization. The loop was in a horizontal position for this test.

Type Antenna	*Relative Sensitivity*
Half-Wave Dipole	6
Loaded Dipole	3
Loop	2

The greater sensitivity of the loaded dipole as compared with that of the loop works out as an advantage for directional reception of horizontally polarized waves (dipole in horizontal position) and as an advantage for nondirectional reception of vertically polarized waves (dipole in vertical position).

EFFECT OF WOOD-FRAME HOUSE ON RECEPTION

A survey was made comparing the reception on the two receivers of Figures 1 and 2 in a typical wood-frame dwelling. A small portable transmitter with loop antenna was set up in four different locations: 1T, 2T, 3T, and 4T adjacent to the dwelling. See Figure 3. The two television chassis, with loop and loaded-dipole antennas, were tested in three different locations within the dwelling on the first floor and in one location in the field adjacent to the house. These locations are shown as 1R, 2R, 3R, and 4R. Maximum and minimum antenna microvolts (obtained by rotating the antennas around a vertical axis) were recorded for both antennas and with both polarizations. It should be noted that in these tests the dipole was always in a horizontal position and the loop, in a vertical position. The data obtained are as follows:

FIELD TEST FROM PORTABLE TRANSMITTER ON 69 MC

Trans. Position	*Rec. Position*	*Vertical Polarization*				*Horizontal Polarization*			
		Dipole		*Loop*		*Dipole*		*Loop*	
		Max.	*Min.*	*Max.*	*Min.*	*Max.*	*Min.*	*Max.*	*Min.*
1T	1R	221	55	215	50	75	5	63	13
1T	2R	125	42	125	13	125	17	38	26
1T	3R	125	62	38	20	101	23	44	13
2T	1R	161	45	113	51	161	60	130	41
2T	2R	161	16	76	23	17	6	32	5
2T	3R	68	10	88	38	51	17	63	13
3T	1R	177	75	169	40	247	45	125	32
3T	2R	87	17	26	18	195	10	33	13
3T	3R	110	75	204	40	210	45	88	33
4T	4R			225					
1T	4R	62	62	377	44	161	29	95	38
*		189	57	155	45	163	33	93	28

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* Average for transmitter positions 1T, 2T, 3T and receiver positions 1R, 2R, 3R, corrected for 100-foot separation.

The figures in the table indicate microvolts output from the receiving antennas. The transmitter loop was 2½ feet above ground. The receiver loop and loaded dipole were 6¼ feet above ground for all tests. The field strength of the vertically polarized wave at receiver position 4R from transmitter position 1T was 3.5 times the field strength of the horizontally polarized wave. This field-strength ratio in favor of vertical polarization is abnormally high. The ratio from a normal distant transmitter would be approximately as indicated in

Figure 6. The loaded dipole is 1.5 times more sensitive than the loop at a given field strength.

The average effect of the house on reception is obtained by a comparison of the readings obtained outdoors (with the transmitter in positions 1T and 4T and the receivers in position 4R) with the readings obtained with the receivers indoors (with the transmitter in positions 1T, 2T, and 3T and the receivers in positions 1R, 2R, and 3R). The last line of the chart contains the average for all the indoor readings, with corrections for the difference in transmission distances compared to the outdoor readings for positions 1T and 4R.

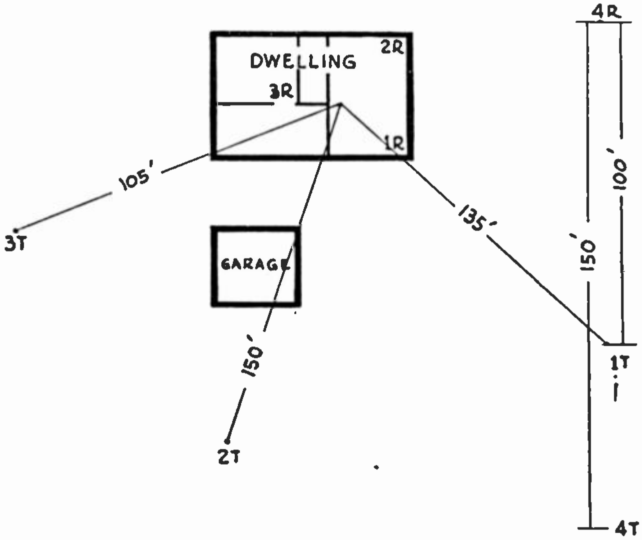

Fig. 3

The individual readings for the different transmitter and receiver positions varied widely, indicating the presence of standing waves within the house for both wave polarizations.

For horizontally polarized waves the average indoor readings were substantially the same as the outdoor readings.

For vertically polarized waves the loop maximum signal dropped from 377 microvolts outdoors to an average of 155 microvolts (40 per cent) indoors. This indicates that the polarization plane of the waves was distorted or that the waves were attenuated. The dipole maximum signal increased from 62 microvolts outdoors to an average of 189 microvolts indoors. This change indicates that the polarization plane of the waves was distorted so as to have a substantial component in the horizontal polarization plane.

The maximum voltage recorded outdoors for vertically polarized waves on the loop was 2.3 times the maximum voltage recorded for horizontally polarized waves on the loaded dipole. Indoors, the respective average maximum voltages were substantially equal. This suggests the possibility that rain pipes, electric wiring, and plumbing as they are situated in wooden frame houses may adversely affect vertically-polarized waves more than horizontally-polarized waves.

During the tests it was noted that the maximum signal on the vertical loop, for reception of horizontal waves, occurred when the loop was turned broadside to the arriving wave. Mr. A. H. Turner offered the theory that this response was due to the differences in field strength at the top and bottom of the loop, i.e., due to the vertical voltage gradient of the horizontally polarized wave. If correct, this theory would require that the response remain constant with height of loop above ground so long as the rate of change of field strength with height remains constant. This conclusion was verified by experiments which appear to confirm the voltage-gradient theory for vertical loop reception of horizontally polarized waves.

Body Effect on Reception

It was observed that persons moving about in the vicinity of the receiving antenna affected the reception, the greatest effect on the received signal strength occurring when a vertical dipole or vertical loop was being used. This result is to be expected since the body acts as a vertical dipole. The body effect was further investigated as follows:

The first tests were made in the open field with the portable transmitter located at point 4T and a half-wave receiving dipole located at point 4R of Figure 3. The receiving dipole was 6 feet, 3 inches high at its center. A vertical dipole was used for vertically polarized wave reception and a horizontal dipole rotated for normal maximum reception was used for horizontally polarized wave reception. A man 6 feet tall stood on a wooden support 22 inches above ground in positions at 10-inch intervals in front and in back of the receiving antenna. The recorded data are shown in Figure 4 for 69 megacycles and 45 megacycles. When the man's arms were raised parallel to his shoulders and parallel to the dipole, the effect on horizontally polarized wave reception was increased.

For each test the receiver gain was first adjusted to give the same arbitrarily chosen output of 100 microamperes without the presence of the man in the vicinity of the antenna. The new meter reading caused by the presence of the man was then recorded. The curves are, therefore, only an indication of the relative change in output due to the presence of the man.

The tests were also made with the man in positions along a line at right angles to the direction of wave propagation. The variations in signal recorded under this condition were never greater than those indicated for positions in line with the direction of wave propagation.

A second set of tests at 69 Mc were conducted with the receiving antenna located near the middle of the living room of the dwelling as illustrated in Figure 3. In these tests the man stood on the floor. The same horizontal dipole was used for horizontally polarized wave reception. Two vertical dipoles, spaced 40 inches apart and cross connected, were used for vertically polarized-wave reception. This type of antenna gives the same bidirectional reception for vertically polarized waves as a vertical loop.

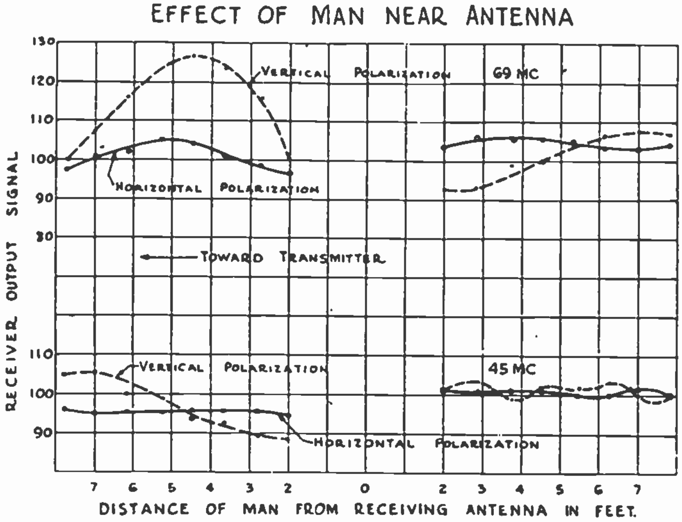

Fig. 4

When the antennas were oriented for maximum signal strength, the results were substantially the same indoors as outdoors. In one test, with the antennas rotated 45 degrees from the maximum-gain position, the response varied 2-to-1 for vertical polarization as the man walked across the room. For horizontal polarization the gain varied only 10 per cent. As the antennas were oriented towards the minimum-reception position the effect of the body became more pronounced for both polarizations.

These tests confirm the opinion that the movements of people in the vicinity of receivers operating on frequencies of the order of 70 megacycles with a built-in antenna are more likely to interfere with the reception from vertically polarized waves than with that from horizontally polarized waves. The greatest effect will be observed on the minimum response from bidirectional antennas oriented to reduce multiple images and other interferences.

RECEPTION IN STEEL-FRAME BUILDINGS

A number of field tests were conducted in New York City on television reception from Station W2XBS, Empire State Building, operating on the former No. 1 channel (44 to 50 megacycles). The results obtained at three locations on the receiver with the loaded-dipole antenna were as follows:

At 26 East Ninety-Third Street in a tenth-floor apartment, an input of 95 microvolts was obtained in a room location which gave poor results. At another location removed 15 feet, in the same room, an input of 560 microvolts gave a fair-quality picture when the antenna was oriented to reduce multiple-image responses. This location was on the side of the apartment away from the transmitter. The distance was 3 miles from the transmitter. A better picture was obtained on an outdoor antenna located on the roof.

At 75 Varick Street on the sixteenth floor facing the transmitter, an input of 1550 microvolts was recorded. This signal gave an excellent picture. Moving the receiver back towards the middle of the building gave poor results. The distance was 2 miles from the transmitter.

At the RCA Building on the fifty-third floor facing the transmitter, an input of 3000 microvolts was recorded. This gave a good picture. Another location on the opposite side of the building gave an input of 150 microvolts and a very poor picture due to multiple images. The distance to transmitter was 0.7 mile.

This survey indicates that in office buildings and apartment houses of steel construction, dependable service using built-in antennas will probably be found in locations facing the transmitter and preferably within line of sight. A bidirectional antenna is desirable to reduce multiple images.

EFFECT OF CITY OBSTRUCTIONS

The relative field strength of vertically and horizontally polarized waves passing mainly through residential areas was also investigated. For these tests a half-wave dipole receiving antenna was located remote from the receiver and buildings so as to minmize the effect of nearby obstructions.

The small test transmitter previously referred to was placed 5 feet above ground in a residential location at Haddonfield, New Jersey. The receiving dipole antenna was placed in three different locations in a field, at heights ranging from 4 to 12½ feet. At the transmitter site the ground was about 40 feet higher in elevation than the receiving locations, most of the ground rise occurring near the transmitter. The transmission distance was 0.6 mile. The receiving locations were about 150 feet from each other and about the same distance from the nearest trees and metal fences. There were eight rows of detached dwellings between transmitter and receiver. The nearest houses in line with the propagation path were 300 feet from the receiving locations.

The data obtained are recorded in Figure 5. The dots are for vertically polarized waves and the circles are for horizontally polarized waves. The solid-line curves *A* and *B* were plotted from the results of theoretical calculations[3], in which a ground dielectric constant of 15 and a transmitter height of 43 feet were assumed. With a transmitter antenna height of 5 feet, the response to horizontally polarized waves relative to vertically polarized waves would be approximately in the ratio of Curve *B′* to Curve *A*.

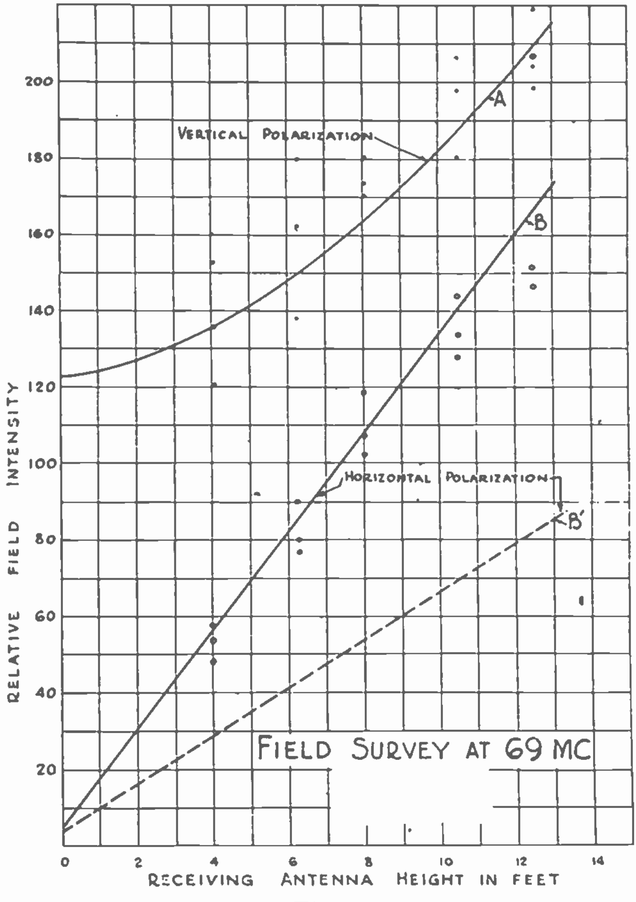

Fig. 5

Further tests were conducted with the portable transmitter located 10 feet above the roof of Building No. 5, RCA Manufacturing Company, Camden, New Jersey. The loop antenna was about 110 feet above ground. Figure 6 gives the field strengths recorded at the Camden Airport, a distance of 2.5 miles. As in Figure 5, the solid curves represent the theoretical calculations. Most of the intervening buildings along the transmission path were of brick and metal-frame construction. The terrain was practically level throughout the transmission path.

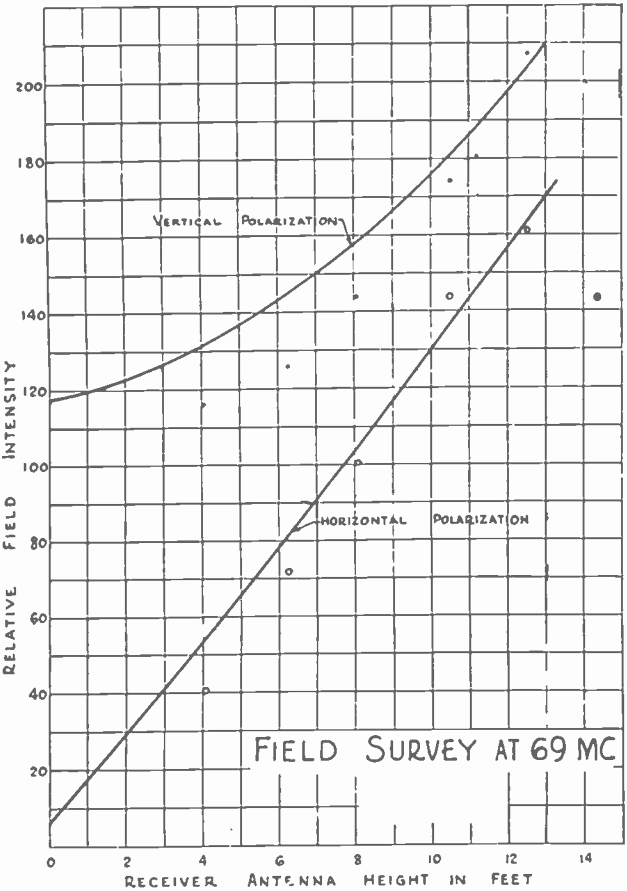

Fig. 6

A test run in back of the airport gave the same field strength for horizontally polarized waves as recorded in Figure 6. The field strength for vertically polarized waves was about the same as for horizontally polarized waves. Around the receiving location the only obstruction which might have caused this drop in vertically polarized signal was a long 6-foot high metal fence 1000 feet away from the receiver in the direction of the transmitter. A reflected wave from some distant object may have caused this result.

With the transmitter in the same location, another group of observations were made with the receiver located in Knight Park, Collingswood, New Jersey. On vertically polarized reception the field strength was normal in one location which was 200 feet remote from any obstacle, see Curve *A* in Figure 7. The field strength (Curve *A′*) was considerably reduced for the second location surrounded by trees. These observations were made in November. There was close agreement in the recorded data for horizontally polarized waves at the two locations. See recorded data *B* and *B′* in Figure 7. The terrain was fairly regular over the 3½-mile transmission path. There were numerous dwellings and miscellaneous buildings between transmitting and receiving locations.

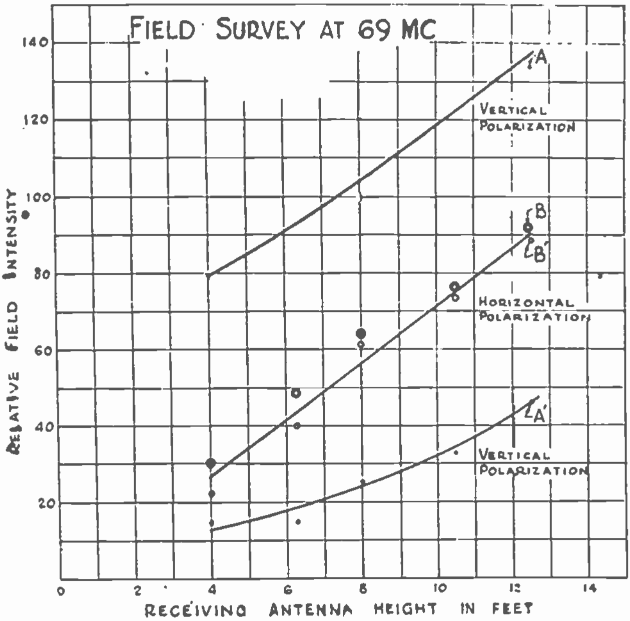

Fig. 7

The close agreement between experimental data and theoretical calculations as recorded in Figures 5, 6, and Curves *B* and *B′* in Figure 7 indicate that low buildings and other city obstructions in the transmission path do not materially affect the relative field strengths of horizontally and vertically polarized waves. The observations which did not agree with the theoretical calculations can usually be accounted for by objects in the vicinity of the receiving location which absorbed and reflected vertically polarized waves more than they did horizontally polarized waves.

The author expresses appreciation for the assistance of Dr. G. H. Brown and Messrs. E. O. Johnson, V. D. Landon, and A. H. Turner in connection with this investigation.

AUTOMATIC FREQUENCY AND PHASE CONTROL OF SYNCHRONIZATION IN TELEVISION RECEIVERS*†

BY

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Summary—*One of the problems in the reception of television images is to provide satisfactory synchronization in the presence of noise. During the past several years considerable experience has been gained with respect to this problem under various receiving conditions. The system of synchronization which has given satisfactory results up to the present time has depended for its operation on the reception and separation of individual pulses. In general, it can be said that with this system satisfactory synchronization can be obtained from those signals which will in all other respects provide an entirely acceptable picture. However, for limiting conditions of service, particularly during early operation where field strength may be low, an improvement in synchronization will be effective and desirable provided that it does not involve other complications or disadvantages.*

This paper describes a synchronizing means at the receiver that employs a new principle in the field of synchronization. The principle is automatic frequency and phase control of the saw-tooth scanning voltages. In such a system, synchrnoziation depends on the average of many regularly recurring synchronizing pulses. Noise has insufficient energy at the scanning frequencies to effect control through the direct-current link from which all but relatively long-time variations are filtered out.

Experimental receivers, in which automatic phase and frequency control of the scanning oscillators has been incorporated, have operated with high immunity to noise. The degree of immunity is of a different order of magnitude from that found in conventional synchronizing systems.

Noise cannot affect horizontal resolution or interlacing. An intrinsic property of the new system is perfect interlacing. The return line in an automatic-frequency-controlled system may start before synchronization.

Consideration of this new development indicates that its use would result in several improvements in television services: (1) when severe noise conditions occur, an improved picture is obtainable at points within the present service area; (2) under such noise conditions the useful service area is extended; (3) the maximum resolution permitted by a television channel is realizable at locations having low field strengths. It is expected that these improved results will be attained without increase in the cost of the television receiver.

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CONVENTIONAL SYNCHRONIZING SYSTEMS

IN THE operation of present commercial television receivers, the natural frequencies of the horizontal and vertical scanning oscillators, in the absence of a synchronizing signal, are lower than the line or field frequencies, respectively, at the transmitter. The application of a transmitted pulse initiates or "triggers" a new cycle of the oscillator before one would otherwise occur. The period of the horizontal-scanning oscillator is shortened to conform to line frequency and the period of the vertical oscillator to field frequency. Thus, triggering is required for each successive horizontal and vertical scan. This is the basic principle of operation of conventional synchronizing systems.

Figure 1 shows a scanning oscillator of a typical commercial television receiver. A cycle of operation in the absence of a synchronizing signal is shown in Figure 2. As the grid potential e_g of the tube T_1 reaches the cutoff point as a consequence of leakage of charge through resistance R from the previously charged capacitance C, plate current i_p begins to flow. A short time later, the induced voltage e_t causes the capacitance C to take a large charge which, in turn, lowers e_g to a high negative value. Plate current does not flow again until sufficient current has leaked through R. The excursions of e_g above the cutoff point of the tube T_2 are responsible for the generation of a saw-tooth wave e_s in the plate circuit of that tube.

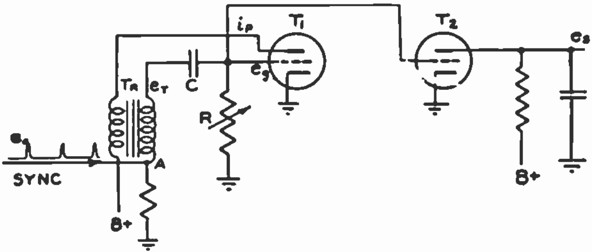

Fig. 1—Conventional triggered scanning oscillator.

Assume now that a synchronizing pulse e_0 is applied between point A and ground in Figure 1. The effect is a premature rise of the potential e_g to the cutoff voltage of T_1 as shown in full lines in Figure 3a. A pulse of plate current i_p occurs earlier than in the absence of a synchronizing signal. The dotted wave in Figure 3a shows the variations of currents and voltages in the absence of a pulse as in Figure 2. Wave e_s may represent current variations in the coils of an electromagnetically deflected tube or voltage variations across the plates of an electrostatically deflected tube.

The behavior of the triggered oscillator when noise is present in the signal is the primary interest here. Hence, we shall wish to determine how the frequency or phase of the scanning voltages are affected when the picture signal is accompanied by noise peaks which exceed black level and therefore appear in the synchronizing signal as shown in Figure 3b. Noise peak "*a*" superimposed on the normal grid potential e_g curve is insufficient to raise the potential above the cutoff of tube T_1; hence, the peak is ignored by the oscillator. Noise peak "*b*", however, does have sufficient amplitude to cause e_g to rise above the cutoff potential, and therefore initiates a new cycle of oscillation prematurely. The legitimate synchronizing pulse at "*c*" would have caused the normal cycle shown in dashed lines. The deflection signal e_s shown in full lines corresponds to the premature synchronization. The dashed lines represent the desired signal.

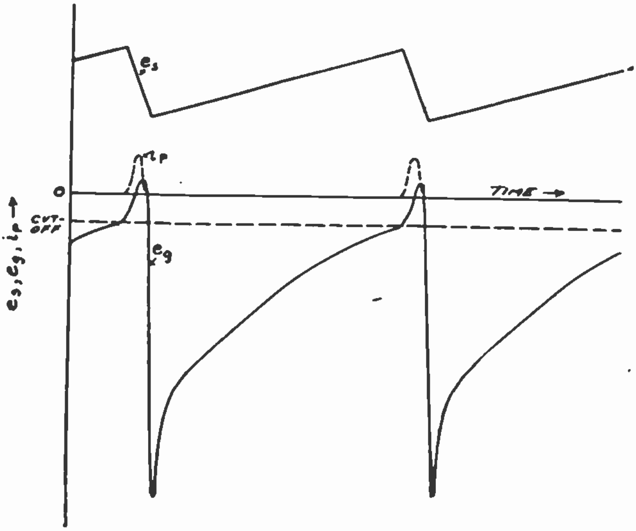

Fig. 2—Operation of the scanning oscillator.

If e_s represents the horizontal-deflection signal, the observer interprets the misplacement of e_s as a line out of the normal position on the viewing screen. If e_s represents the vertical deflection signal, he observes a vertical movement of the picture.

It will be realized that the immunity of the system to noise is least when e_g is near cutoff because lower noise peaks are sufficient to initi-

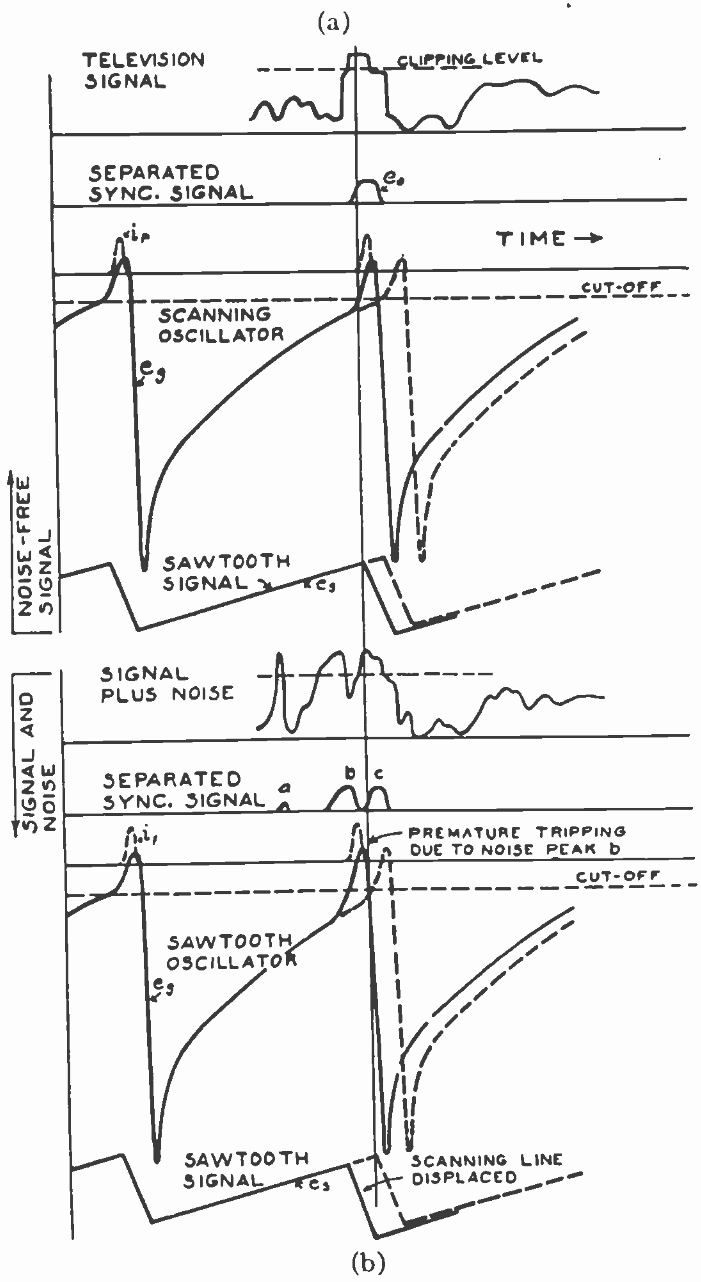

Fig. 3

(a)—Operation of scanning oscillator.
(b)—Operation of scanning oscillator when noise is present.

ate a new cycle of the oscillator. In the event that a synchronizing pulse is obliterated by noise, the oscillator may remain inactive until e_g reaches the cutoff potential of the tube and thus initiates a new cycle which is late relative to the normal position. It is clear, therefore, that triggered synchronizing as described above is subject to noise limitations that are inherent in the principle of operation.

AUTOMATIC FREQUENCY- AND PHASE-CONTROLLED SYNCHRONIZING SYSTEMS

Figure 4 is a block diagram of the essential components of an automatic frequency- and phase-controlled synchronizing system. Since the same principle is involved in the operation of the horizontal and vertical circuits, it is unnecessary to specify a particular circuit. The phase detector receives the synchronizing signal at *A* and a saw-tooth wave at *B* taken from the output of the scanning oscillator. A control voltage produced at *D* by the phase detector contains information regarding the phase of the saw-tooth wave relative to the synchronizing pulses. The phase detectors described below respond to changes in relative phase that may exist at the time of arrival of each pulse. However, only the slowly varying components of the control voltage are passed by the filter following the phase detector. Rapid variations corresponding to rapid or erratic changes in relative phase are eliminated. Thus, the control voltage at *F* may be regarded as a direct voltage which is applied to the scanning oscillator in order to restore the phase of the oscillator relative to the synchronizing pulses when

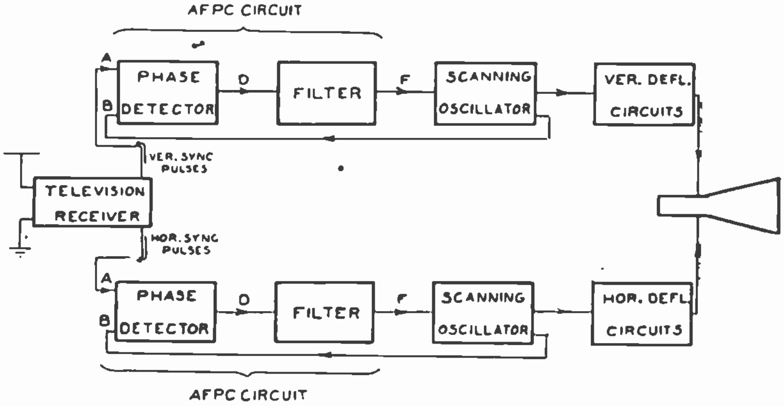

Fig. 4—Block diagram of an automatic frequency- and phase-controlled system.

there is a long-time trend in phase away from the equilibrium state established by the speed control of the scanning oscillator. Such changes in phase and frequency of the saw-tooth as occur as a result of the action of the control voltage are conducted back to the phase detector through the feedback path in order to provide further correction.

In the presence of noise of sufficient magnitude, the phase detector may register the relative phase of a noise peak and the corresponding saw-tooth cycle. Such spurious components in the control voltages at *D* usually lie in the range of frequencies beyond cutoff of the filter and are therefore effectively removed from the voltage at *F*. The noise immunity of automatic frequency- and phase-controlled circuits is a consequence largely of the action of the filter. Further insight into the theory of automatic frequency- and phase-controlled synchronization may be obtained from a more detailed account of the operation of specific circuits.

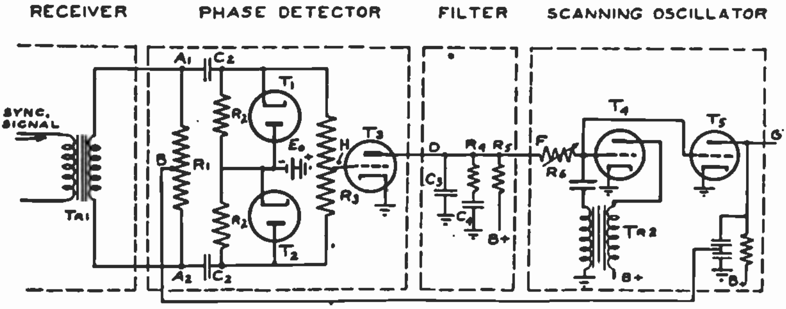

Fig. 5—Automatic frequency- and phase-controlled circuit.

Figure 5 shows a circuit which may be used for automatic frequency and phase control of a horizontal or a vertical oscillator. Here, synchronizing pulses are supplied to the terminals A_1-A_2 of the phase detector by means of a balanced circuit. A fraction of the output of the scanning oscillator is introduced at point *B* of the phase detector in order to form the composite signals shown in Figure 6(a) and (b). In practice, when the automatic frequency- and phase-controlled system is in equilibrium, the synchronizing pulse must occur sometime during the return line, that is, during the steep portion of the saw-tooth wave. This restriction is necessary for the viewing of a television picture in the correctly framed position on the screen of the cathode-ray tube.

If it is assumed that a state of equilibrium is attained, the condition is maintained in the following manner. Tubes T_1 and T_2 are diode rectifiers which may be idealized for simplicity of explanation. We shall assume that the circuit composed of the resistance R_2 and the capacitance C_2, associated with the diode T_1, maintains a potential variation (Figure 6(c)) at the cathode of the diode that resembles the wave in Figure 6b in every respect except that the peak amplitudes of the pulses are definitely located at $-E_0$ volts. In popular terms,

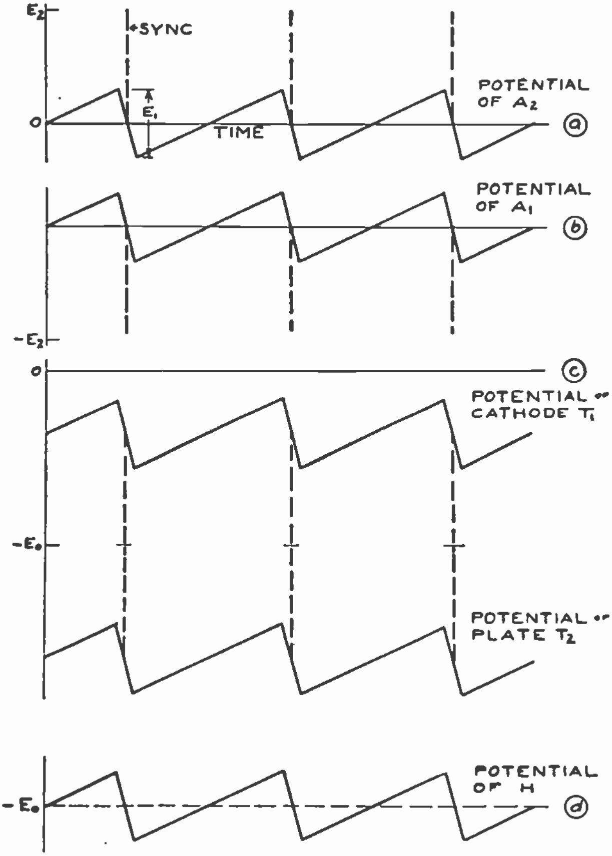

Fig. 6—Composite signals for possible equilibrium position.

the diode is said to "set direct current." The values of R_2 and C_2 must be chosen with the view of causing T_1 to act as a direct-current setter or peak rectifier.

In a similar manner, the diode T_2 in combination with its associated elements R_2 and C_2 maintains the potential variation at the plate of T_2 as shown in Figure 6c in which the peak amplitude of the synchronizing pulse is maintained at a potential of $-E$ volts with respect to ground. The potential with respect to ground of the mid-point of the resistance R_3 shown in Figure 6(d) is the average of the potentials at the end point of R_3. An important observation to be made in Figure 6(d) is that the synchronizing signal is balanced out. This leads to the conclusion that the waveform and the direct-current component of the signal at point D are independent of the amplitude assumed for the synchronizing signal, but that the direct-current component is dependent upon the phase relation of the synchronizing signal and saw-tooth wave.

The low-pass filter in the plate circuit of the amplifier tube T_3 transmits the direct-current component of the signal at point D and greatly attenuates the alternating-current components. The amplified direct-current component or control signal at point F is applied as a positive bias to the grid of the scanning oscillator tube T_4. The frequency of the oscillator is a function of the grid bias. Consequently, the saw-tooth wave generated by tube T_5 has a frequency controlled by the resistance R_6 and the control voltage at point F. This signal is applied to the phase detector at point B by way of the feedback path.

The capability of the circuit for controlling the frequency and phase of the saw-tooth wave may be understood with the aid of Figure 7. Assume, as was done above, that Figure 6 expresses a state of equilibrium in the circuit and that the frequency and phase of the saw-tooth will remain indefinitely as shown if the circuit is not disturbed. Let Figure 7(a) represent a departure from the equilibrium condition as a result of some disturbance such as drifting in the values of circuit constants or voltages. The relative phase of the synchronizing signal and the saw-tooth wave differs from the equilibrium phase relation (Figure 6(a)) by an amount ΔT. As before, the peak rectifiers hold the peaks of the synchronizing signal at a potential $-E_0$. Hence, the alternating-current axis of the control signal at point H (Figure 7(d)) is lowered by an amount ΔE. The direct-current component at point H therefore amounts to $-(E_0 + \Delta E)$ volts or an increment of $-\Delta E$ volts over the equilibrium value in Figure 6(d). This increment tends to increase the frequency of the oscillator and thus to shift the saw-tooth wave toward the position of equilibrium. Similarly, a departure from equilibrium shown in Figure 8(a) and (b) gives rise to an increment of $+\Delta E$ volts in the control voltage which acts to decrease the frequency of the oscillator and thus again restore the equilibrium of the circuit.

The amount of control or hold-in power available for overcoming phase and frequency deviations of the saw-tooth wave is proportional to the gain of the direct-current amplifier and to the difference between the direct-current components of the control signals at *H* corresponding to the two extreme phase conditions. Extreme conditions occur when a synchronizing pulse occurs either at the maximum or minimum points of the saw-tooth wave. This difference is equal to the amplitude of the saw-tooth signal introduced at point *B* except for negligible voltage drops in the phase-detecting circuit.

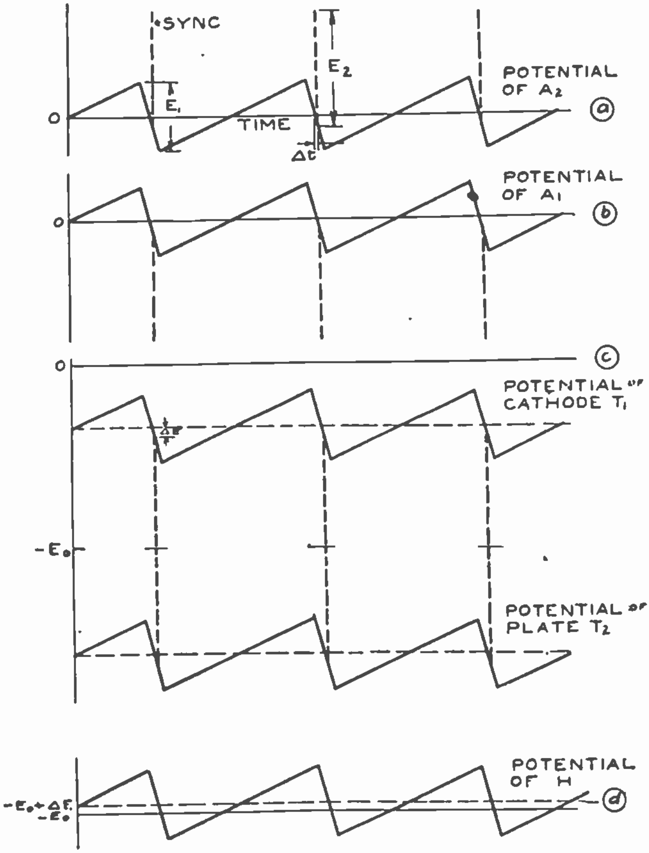

Fig. 7—Phase of saw-tooth delayed from equilibrium position of Fig. 6.

When noise is present in the synchronizing signal, the voltage across the terminals A_1-A_2 contains noise pulses which have been passed by the synchronizing separators as sketched in Figure 9(a). The peak amplitude of some noise pulses exceed the bias voltage of the diodes and cause diode current to flow. Hence, the signal at point *H* resembles the erratic saw-tooth wave shown in Figure 9(b) in which the noise pulses themselves are balanced out. The filter in the plate circuit of T_3 transmits only the direct-current component and the slowly varying components of the grid voltage. All components above a few cycles per second in frequency are effectively suppressed. The slowly varying components represent a persistent trend in the alternating-current axis away from the equilibrium position and give rise to a change in the control voltage applied to the oscillator. The resulting deviation in the phase and frequency of the oscillator is automatically minimized by the restoring effect of the automatic frequency- and phase-controlled circuit.

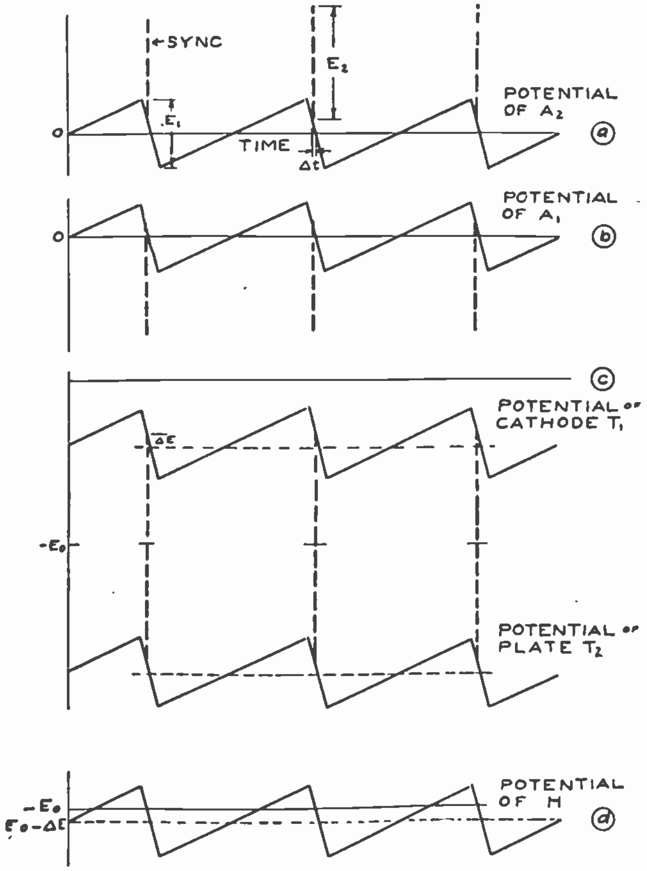

Fig. 8—Phase of saw-tooth advanced from equilibrium position of Fig. 6.

Almost identical constants have been used in the filters for the horizontal and vertical circuits. The response of the filter is reduced to about one third for a sine wave of 1 cycle per second and the response to 60 cycles is practically zero. Therefore, individual scanning lines cannot be perceptibly displaced with respect to neighboring lines. That is, the horizontal oscillator cannot respond to noise fast enough to impair horizontal resolution. In general, a triggered oscillator is sensitive to noise to an extent that horizontal resolution is decreased. Likewise the response to 30 cycles is so low that interlace is essentially perfect, even in the presence of severe noise or an imperfect vertical-synchronizing signal.

Another form of phase detector is shown in Figure 10. This circuit uses four diodes and is inherently balanced, except for second-order effects. The circuit is more complicated than the two-diode circuit of Figure 5 but is somewhat easier to set up and adjust. The principle of operation is the same as the two-diode circuit, although the details are different, and a description can be given more easily in another manner. The four diodes may be considered as a single-pole single-throw switch which connects the output capacitance C_3 to the input circuit resistance R_2 during the synchronizing pulse interval. This is accomplished as follows: the four diodes may be considered as a bridge in which synchronizing pulses are applied in push-pull across the diagonal AB with polarities such as to cause current conduction in each diode. A direct-current biasing voltage, that maintains all diodes in the nonconducting stage during the intervals between pulses, is built up across the combination R_1C_1. This circuit is complete in itself; no battery is needed. The two corners C, D of the bridge are connected

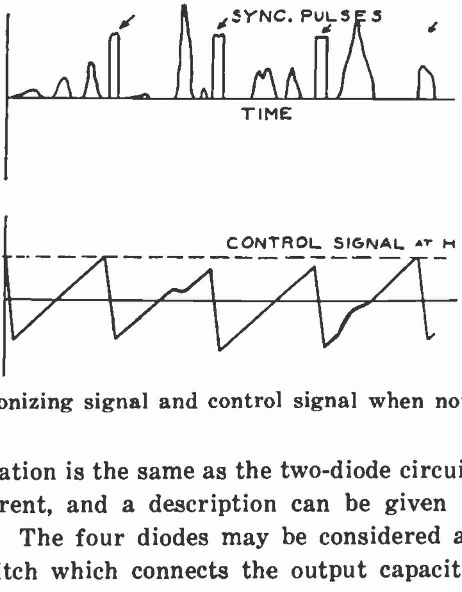

Fig. 9—Synchronizing signal and control signal when noise is present.

to the input and output circuit, respectively. It will be noted that each of these corners connect within the bridge to both a cathode and an anode. Thus, when the diodes are in a conducting state, current may flow in either direction between the input and output circuits. Capacitance C_3 in the output circuit receives a charge which brings the potential at point *C* nearly to the value existing at the input point *D* during the synchronizing pulse interval. This voltage will, of course, depend upon the phase of the synchronizing pulse and the saw-tooth signals.

In the circuit of Figure 10, the signal at *C* contains neither of the input signals but only a direct-current component which is corrected once during each pulse in accordance with the phase relation between the input signals. This voltage controls the output of the direct-current amplifier which acts through the feed-back loop and causes the phase of the saw-tooth wave to vary until an equilibrium is reached. This equilibrium occurs on the positive slope of the saw-tooth. In order that the phase relation for a properly framed picture shall exist, the saw-tooth applied at point *D* must have the polarity for which the return-line portion has a positive slope. The reason for this requirement may be seen by tracing the operation of the circuit. Assume that the local oscillator is out of equilibrium in such a direction that its frequency is low. The return line (positive slope) of the saw-tooth will occur late, as shown dotted in Figure 11 and the synchronizing pulse will occur at a more negative point as shown at *B*,

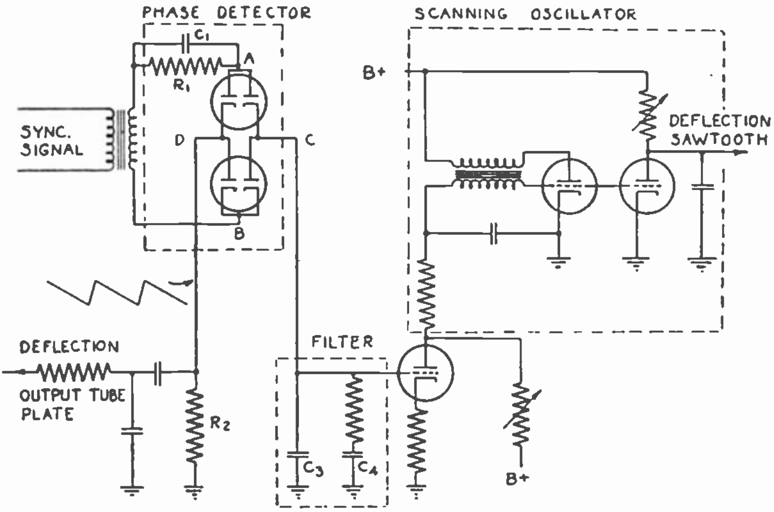

Fig. 10—Four-diode automatic frequency- and phase-controlled circuit.

rather than at A. The potential of capacitance C_3 then becomes more negative and the plate of the amplifier tube becomes more positive with the result that the frequency is increased and the equilibrium restored.

The circuits shown in Figures 5 and 10 require the application of saw-tooth signals of opposite polarity to the phase detector. Furthermore, the signal applied to the direct-current amplifier in Figure 5 contains a saw-tooth component whereas the circuit of Figure 10 does not. Either circuit could be changed to operate the same as the other in these two respects by reversing the input and output connections of the phase detector. There are many possible variations in phase detectors. However, the two described have been used satisfactorily and serve to illustrate the important characteristics of this portion of an automatic frequency- and phase-controlled system.

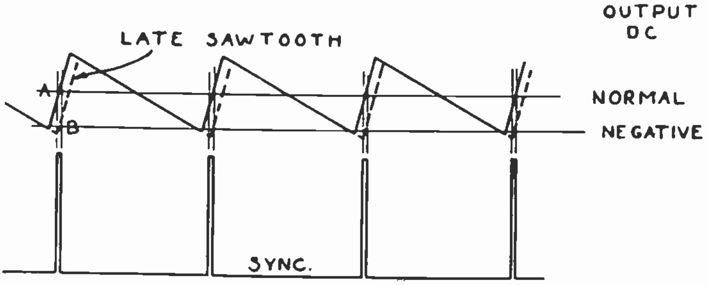

Fig. 11—Equilibrium and off-equilibrium conditions for four-diode circuit.

In an automatic frequency- and phase-controlled system there is an approximate equivalent to the degree of lock-in of a tripped oscillator. The tripped oscillator is locked in tighter when the amplitude of the synchronizing signal is increased. Greater susceptibility to noise is noticed in the tightly locked-in oscillator. In the automatic frequency- and phase-controlled system the amplitude of the synchronizing signal is relatively unimportant but the alternating-current gain from the phase detector to the oscillator influences the speed with which the oscillator may be changed in frequency, that is, the tightness of lock-in. If the automatic-frequency and phase-controlled oscillator can be shifted rapidly, relatively few pulses are required to obtain control and noise is averaged out over a short period. Conversely, if the gain is low, many consecutive pulses are necessary to obtain control and noise is averaged out over a long period, a condition which is obviously desirable. A low alternating-current gain, unfortunately, has another effect; i.e., when the receiver has just been turned on, or for some other reason is completely out of synchronism, the time for pulling into synchronism may be excessively long.

The mechanism of pull-in in an automatic frequency- and phase-controlled system may be outlined as follows: Assume that when no signal is received, the speed-control setting is such that the frequency of the oscillator deviates from synchronous frequency by a small amount. When a synchronizing signal is received, the phase detector generates an alternating-current wave, the frequency of which is the difference frequency of the synchronizing signal and the locally generated saw-tooth wave. If the difference frequency is attenuated strongly by the filter, there is little or no tendency for the oscillator to pull in since no direct-current control signal is generated. If the filter is such that some components of control voltage are passed without excessive phase shift and attenuation, the frequency of the oscillator tends to follow the instantaneous value of the control signal.

When the oscillator is pulled toward synchronism, the momentary difference frequency is decreased and the oscillator tends to remain longer in this phase than in the opposite phase during which the difference frequency is increased. This amounts to a distortion of the control signal which produces a new axis of the control voltage in the direction to pull the oscillator toward synchronism. The oscillator may be said to take three steps toward synchronism and two away, the sequence continuing until the deviations become sufficiently small and the oscillator falls into synchronism. Low phase shift through the filter and appreciable alternating-current gain are favorable for rapid pull-in. However, since the system employs a feedback loop, care must be taken to avoid self-oscillation. These requirements have led to the unusual filter, shown in Figures 5 and 10.

An automatic frequency- and phase-controlled system, unlike a triggered system, is not limited to a single phase relationship between synchronizing and deflection. That is, the blanking bar may be caused to occur, tightly locked in, anywhere on the screen. This means that the deflection-return line may start at the beginning of the "front porch,"[1] ahead of the synchronizing pulse, thus greatly easing the return-line time requirements in the deflection system. Shifting of the blanking bar is accomplished as follows: Equilibrium, as previously stated, occurs with the synchronizing pulse located on the return line of the saw-tooth signal applied to the phase detector. This saw-tooth,

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[1] The term "front porch" refers to the part of the synchronizing signal between the beginning of the blanking signal and the front edge of the synchronizing pulse.

however, need not be taken directly from the oscillator. The steep edge of the original saw-tooth wave may easily be delayed and, since the synchronizing pulse occurs in equilibrium on this delayed edge, the blanking bar must necessarily occur after the oscillator has tripped. An appropriate delay for horizontal deflection occurs automatically if the saw-tooth is made by integrating the pulse on the plate of the deflection tube, the delay being caused by the deflecting coil itself.

EXPERIMENTAL RECEIVERS

Several experimental receivers incorporating automatic frequency- and phase-controlled circuits have been constructed. These have been tested both in the laboratory and the field and have given remarkably superior performance over conventional receivers for limiting conditions of service. Some pictures were taken in an attempt to show this difference but obviously it is impossible to convey accurately information regarding accuracy of synchronization by means of a still photograph. Figure 12 shows a test pattern received by an automatic-frequency and phase-controlled receiver when operating with a signal above the hiss level but with interference from a high-frequency buzzer. The interference can be seen only as short black lines, and the figure serves mainly to show the resolution and interlace obtained essentially in the absence of noise. Exposure for this and the other test-pattern

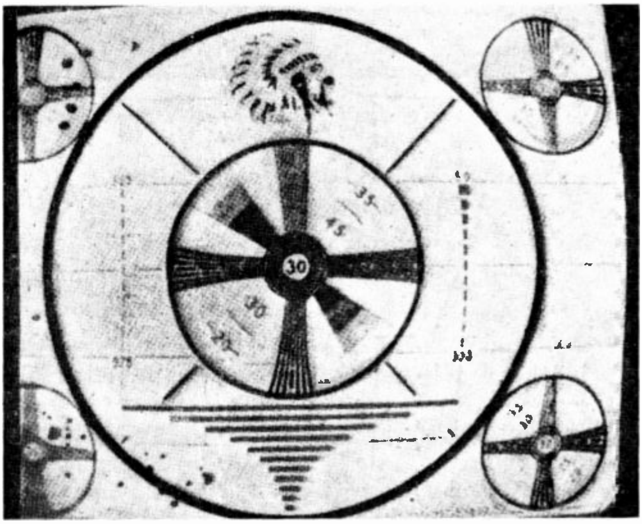

Fig. 12—Automatic frequency- and phase-controlled receiver—interference from high-frequency buzzer.

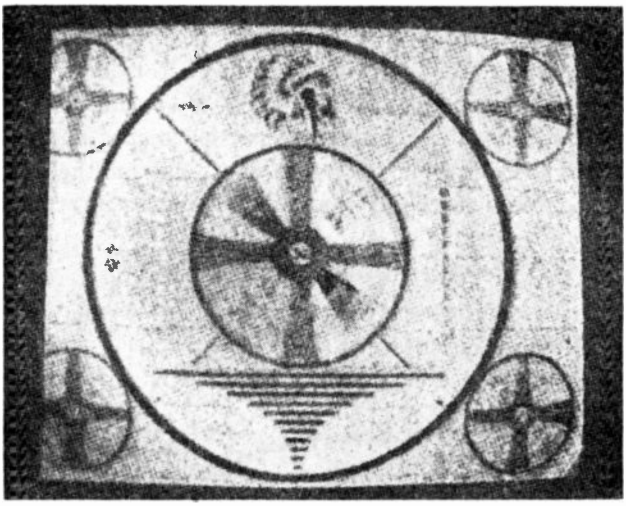

Fig. 13—Conventional receiver—hiss noise.

pictures was approximately 1/15 second, or 2 frames.

Figures 13 and 14 are test patterns showing the relative synchronizing capabilities of a conventional receiver and an automatic frequency- and phase-controlled receiver in the presence of about equal amounts of hiss noise. The loss of horizontal resolution and lack of interlace are plainly evident in Figure 13. Resolution in Figure 14 is limited only by the modulation of the kinescope by noise. This obser-

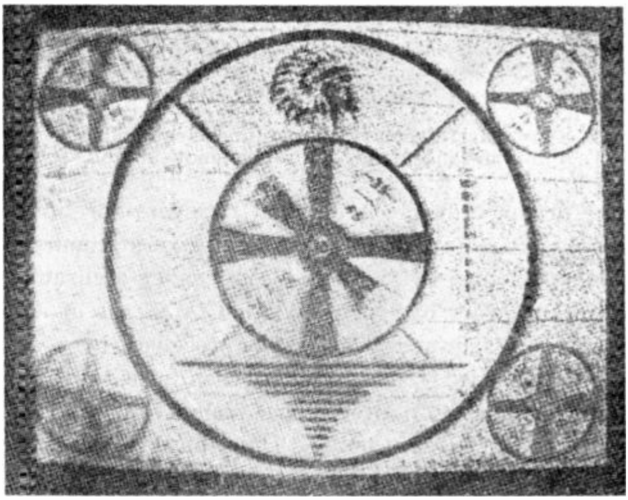

Fig. 14—Automatic frequency- and phase-controlled receiver—hiss noise.

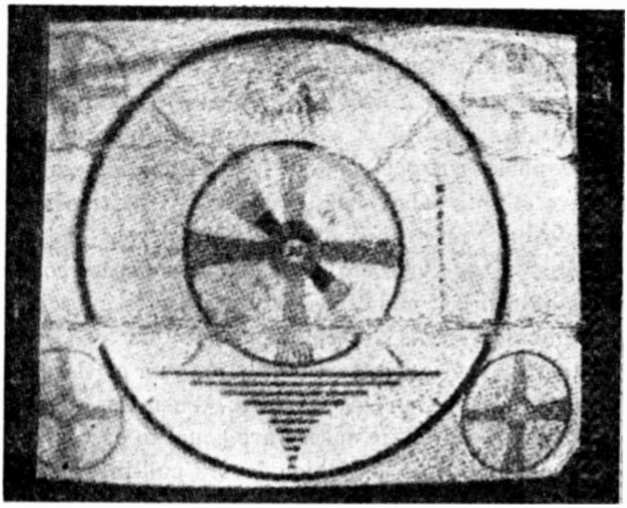

Fig. 15—Conventional receiver—interference from electric razor.

vation was made when a noise-free driven synchronizing signal was substituted but the picture signal left unchanged. Interlacing in Figure 14 is not perceptibly affected by noise.

The interference in Figures 15 and 16 was caused by an electric razor. Synchronization in Figure 15 (conventional receiver) is lost entirely during severe noise peaks. Horizontal resolution is also seriously affected. Figure 16, received on an automatic frequency- and

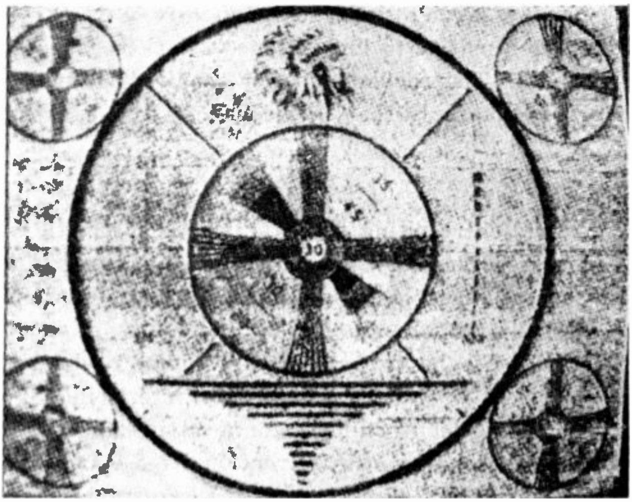

Fig. 16—Automatic frequency- and phase-controlled receiver—interference from electric razor.

phase-controlled receiver, does not exhibit the loss of resolution seen in Figure 15.

The operating characteristics of an automatic frequency- and phase-controlled receiver and of a conventional receiver are quite different. During severe noise conditions a picture synchronized by automatic frequency and phase control remains together as a whole but may appear to move slightly about the equilibrium position in a random manner. Single lines or groups of lines cannot tear out horizontally because the filter in the horizontal circuit does not pass components of the control signal which would cause abrupt changes in oscillator speed. When synchronizing signals are obliterated for an appreciable length of time, the vertical and horizontal oscillators run at the free-running speed until synchronizing is re-established. When the receiver is properly adjusted, the free-running speeds are equal to or very close to the synchronous speeds, a condition which favors pull-in.

Automatic frequency- and phase-controlled synchronization is not entirely without disadvantages. For instance, reasonably good stability must exist in the synchronizing generator at the transmitter. However, the tentative standard recommended by the Federal Communications Commission is deemed entirely adequate and present synchronizing generators are well within the recommended standard. Also, since the system is slow to fall out of synchronism, it is likewise slow to pull into synchronism. For instance, during a local thunderstorm, when a multiple lightning stroke obliterates the signal for a considerable portion of a second, the oscillators may fall out of synchronism and require as much as a second to resynchronize.

The system is sensitive to line-voltage variations unless glow-tube regulation is used. The regulated power required is small, however, and the chief objection is that occasioned by the extra tubes and sockets.

CONCLUSIONS

Superior reception resulting from the use of automatic frequency- and phase-controlled synchronizing has been experienced in field tests under conditions of severe noise such as may exist occasionally even within the normally useful service area of a television station. Horizontal resolution is found to be limited only by modulation of the kinescope by noise. Noise does not destroy interlacing of scanning lines. Tearing of the picture in horizontal strips and rapid vertical movement which may occur in a conventional receiver during severe noise bursts are essentially eliminated by the long time constants of an automatic frequency- and phase-controlled system.

RADIO-FREQUENCY-OPERATED HIGH-VOLTAGE SUPPLIES FOR CATHODE-RAY TUBES*†

BY

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Summary—The operation of tuned step-up transformers in self-excited oscillator circuits as high-voltage sources for kinescopes is analyzed. General information and data are given for optimum radio-frequency-transformer design and operating conditions with specified rectifier loads. Practical high-voltage supplies are illustrated ranging from 1 to 50 kilovolts with power-output values of one-quarter watt to 50 watts, respectively. The performance of these supplies in television equipment is discussed.

INTRODUCTION

THE operation of cathode-ray tubes for television requires high-potential direct-current sources, ranging in voltage from less than 1 kilovolt for iconoscopes to 30 kilovolts and higher for projection kinescopes.

The conventional high-voltage supply consists of an iron-core step-up transformer energized from the 60-cycle power line, and a rectifier circuit with smoothing filter. Mechanical and insulation problems make it difficult to construct small 60-cycle transformers with tightly packed windings for voltages exceeding approximately 5 kilovolts. Practical transformers, therefore, are relatively large and heavy and can furnish currents considerably in excess of the usual requirements.

The use of high-frequency-power sources permits a substantial reduction in transformer inductance and results in a relatively simple transformer construction. The input power is generated by vacuum-tube oscillators, which automatically limit the possible power output. This characteristic and the low-energy storage in the small smoothing reactances permit the construction of safe supplies provided the current requirements are not too high.

The theory of tuned step-up transformers points out the necessity of constructing unusually high-impedance secondary circuits to obtain

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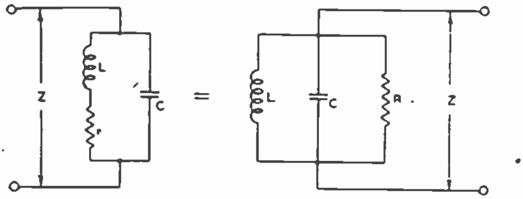
efficient operation. The design of optimum high-voltage coils is, therefore, of prime importance in the construction of practical radio-frequency-operated supplies.

A brief analysis of tuned step-up transformers in self-excited oscillator circuits with rectifier loads will furnish design data for the various circuit components and show their influence on the performance of the high-voltage supply.

THE TUNED STEP-UP TRANSFORMER

The exciting current of a transformer is determined by the reactance of the primary winding and its power factor. The power factor is expressed at radio frequencies by its reciprocal value, the Q value of the reactance. The loss component may be represented as a series resistance r or a shunt resistance R (Figure 1). For Q values greater than 5,



$\tan \phi = \frac{r}{\omega L} = \frac{\omega L}{R} = \frac{1}{Q}$

at resonance: $Z_0 = Q\omega L = R$

Fig. 1—Power factor and impedance of tuned circuits.

$$\left.\begin{array}{l} r = \dfrac{X}{Q} \\ R = QX \end{array}\right\} \begin{array}{l} Q > 5 \text{ and} \\ X = \omega L \text{ or } X = 1/\omega C. \end{array} \qquad (1)$$

The magnetizing current of the transformer is canceled with respect to the power source by the operation of tuning the transformer primary. The resonant impedance, hence, of a tuned circuit is

$$Z_0 = R. \qquad (2)$$

The secondary of the transformer is tuned by the natural circuit capacitances consisting of distributed coil capacitance, diode capacitance, and stray capacitance. The secondary circuit has, therefore, a natural frequency ω_{02} which determines the operating frequency of the transformer.

A high-voltage radio-frequency transformer is a special case of two coupled tuned circuits. The method of coupling is in general immaterial; the circuit however, must be suitable for stable self-excited oscillations, maintain a substantially constant secondary voltage under considerable external load variations, and load the oscillator efficiently.

The use of critical coupling

$$K_c = 1 \sqrt{Q_I Q_{II}} \tag{3}$$

furnishes a maximum voltage step-up for the no-load condition

$$E_2/E_1 = \sqrt{Z_{II}/Z_I} \tag{4}$$

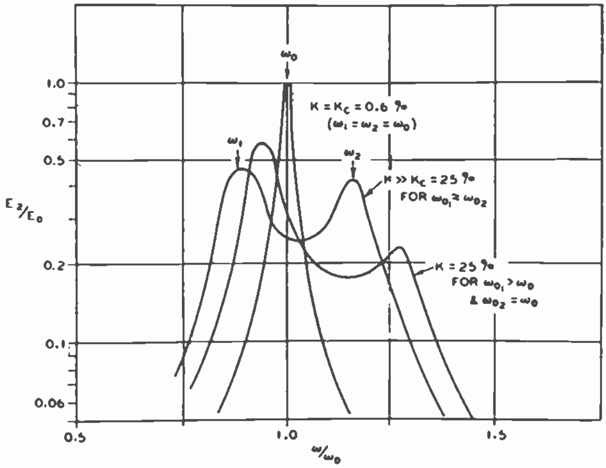


Fig. 2—Frequency characteristics of coupled circuits.

but it is not suitable for variable loads, because of its dependance on the Q value and impedance of the secondary circuit (equations (3) and (4)). The maximum energy transfer into the secondary circuit is limited to 50 per cent of the power input to the primary circuit. The stability of the secondary voltage can be greatly improved by increasing the coupling to a value $K >> K_c$ for a fully loaded circuit. The theoretical efficiency limit then can increase to 100 per cent.

The overcoupled circuit has two coupling frequencies ω_1 and ω_2 which cause a double-hump resonance curve as shown in Figure 2. The spread of the peaks depends on the coupling,

$$K = \frac{1 - (\omega_1/\omega_2)^2}{1 + (\omega_1/\omega_2)^2}. \tag{5}$$

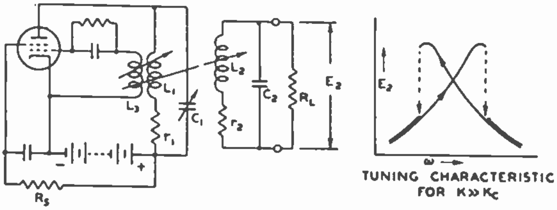


Fig. 3—Oscillator circuit for $K \leqq K_c$, and unsuitable tuning characteristic when $K >> K_c$.

The relative amplitude of the peaks depends on the relative frequencies ω_{01} and ω_{02} to which the circuits are originally tuned before coupling. It is, hence, possible to control the secondary voltage E_2 by changing the primary tuning without change of coupling or of secondary tuning.

The best voltage stability for variable loads is obtained by operation at the lower coupling frequency, and maximum energy is obtained for a tuning adjustment $\omega_{01} \approx \omega_{02}$. The latter adjustment, however, is not critical. It is, therefore, the desirable operating condition of the circuit. The voltage step-up is reduced to approximately one half of the maximum obtainable in order to provide high efficiency and good voltage regulation. The latter is in the order of 7 to 15 per cent from no load to full load when the output is measured at the direct-current terminals of practical kinescope supplies and includes oscillator performance. A coupling of $K \geqq 20\ K_c$ is required at full load.

REQUIREMENTS FOR SELF-EXCITATION, INDUCTANCE, AND Q VALUES

Self-excitation with feedback from the primary winding causes an unstable tuning characteristics as indicated in Figure 3. A stable oscillation characteristic requires coupling of the grid-circuit inductance L_3 to the secondary circuit L_2 as shown in Figure 4. The circuit oscillates at the lower frequency peak ω_1 when the winding directions between L_1 and L_3 are as in normal oscillator circuits. Reversal of L_1 or L_3 causes stable oscillation at ω_2.

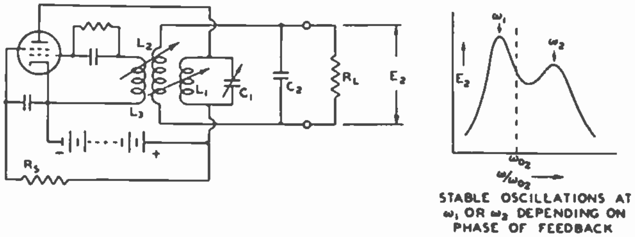


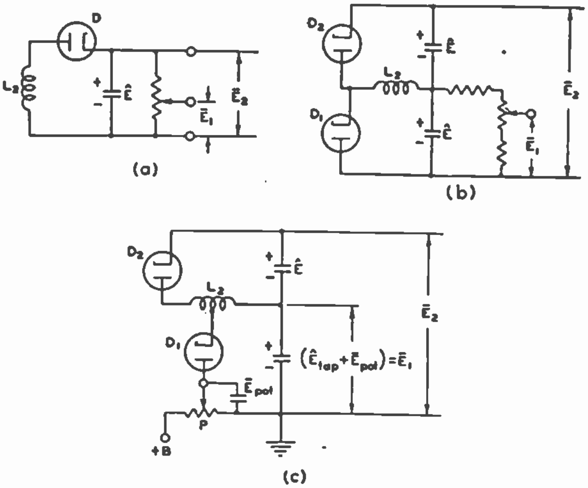
Fig. 4—Oscillator circuit for all values of K, and stable characteristic when $K >> K_c$.

The full-load Q values of primary and secondary circuits should be high to obtain a large degree of over-coupling ($K \approx 20\ K_c$) with moderate values of K which cannot be made very large because of insulation requirements ($K \approx 25$ per cent).

Desirable values are

$$Q_I \geqq 10 \text{ when transformer is shunted by the reflected plate load } R_p$$
$$Q_{II} \geqq 20 \text{ when transformer is shunted by the equivalent rectifier load } R_L. \tag{6}$$

Corresponding inductance values are



(a) (b) (c)

Fig. 5—Rectifier circuits for kinescope supply voltages $\bar{E}_1$ and $\bar{E}_2$.

$$\omega L_1 \leqq 0.1 R_p$$
$$\omega L_2 \leqq 0.05 R_L. \tag{7}$$

The no-load Q values should of course be considerably higher than the full-load values. A loss of 10 per cent per circuit requires ten times the Q value given in (6); i.e.,

$$Q_{L1} = 100; \qquad Q_{L2} = 200. \tag{8}$$

THE EQUIVALENT RECTIFIER LOAD

The equivalent rectifier load R_L depends on the rectifier circuit, three types of which are shown in Figure 5. The alternating-current load R_L is determined by the direct-current load resistance $\bar{R}$, the direct-current output voltage $\bar{E}$, and the alternating peak voltage $\hat{E}_2$ applied to the rectifier tubes.

$$R_L = \frac{(\hat{E}_2)^2 \bar{R}}{2\bar{E}^2} \tag{9}$$

$$R_L = 1/2\bar{R} \text{ for half-wave rectifiers}$$

$$R_L = 1/8\bar{R} \text{ for voltage-doubling circuits.} \tag{9a}$$

The direct-current load resistance $\bar{R}$ of a supply furnishing 1 milliampere at 10 kilovolts is $\bar{R} = 10$ megohms. The secondary circuit feeding a half-wave rectifier must, therefore, have an impedance $Z_0 = R = 10R_L$, i.e., 50 megohms for a secondary loss of 10 per cent. Secondary circuits of such high impedance are too expensive and large for practical use and efficiency is, therefore, sacrificed in favor of size and cost as shown later. Equation (9a) points out the advantage of a doubling circuit, from an efficiency standpoint, because it requires only one fourth the circuit impedance. The circuit of Figure 5(c) is similar to a doubling circuit, except that D_1 rectifies only part of the coil voltage. The voltage $\hat{E}$ tap is made slightly lower than the desired focusing potential $\bar{E}_1$ for electrostatic types of kinescopes. $\bar{E}_1$ is adjustable by means of the poteniometer P, which allows the addition of B-supply voltage to the radio-frequency voltage. This circuit has a high efficiency, because it does not dissipate power in a bleeder resistance. It maintains also a very stable voltage at increased first-anode current.

THE REFLECTED PLATE LOAD

The primary-circuit constants are determined after the secondary coil has been designed from the operating frequency, the total power output $P.O.$ to be supplied by the oscillator, and the oscillator peak voltage swing $\hat{E}_p$. From these, the reflected load,

$$R_p = \frac{(\hat{E}_p)^2}{2P.O.}. \tag{10}$$

The primary reactance ωL_1 is then obtained from (7). The problem of designing an optimum high-voltage coil and determining its operating frequency may be approached in the following manner.

HIGH-VOLTAGE COIL DESIGN

Physical Dimensions

The physical size of the coil depends on the required minimum sparking distances and the power which must be dissipated. The latter is at first unknown. A coil of desirable size for the particular purpose is hence chosen and given a copper cross section consistent with voltage requirements and high-Q values. The winding is subdivided into pies (Figure 6) with approximately 5 turns per layer, the pie spacing being somewhat less than the pie height in order to maintain the same potential gradient between coils as inside the winding. The coils should be supported by strips of insulating material or by an inpregnated paper tubing, which is perforated to permit free circulation of air and to reduce dielectric losses. The coil size indicated in Figure 6 will dissipate approximately 6.5 watts in a horizontal position.

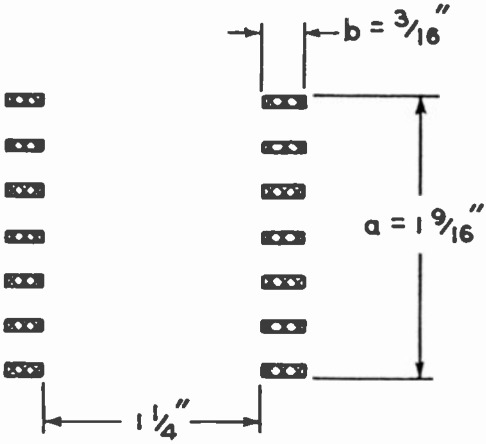


Fig. 6—Dimensions of high-voltage coil for 10 to 15 kilovolts.

DESIGN FOR OPTIMUM ELECTRICAL CHARACTERISTICS

The power loss in the coil is given by

$$P = (E_2)^2/R. \tag{11}$$

The equivalent shunt-loss resistance R at resonance (see Figure 1 and (1)) can be written $R = L/rC$. Thus,

$$P = (E_2)^2 \frac{rC}{L} \tag{11a}$$

For given values of secondary voltage E_2 and tuning capacitance C, which should be as small as possible, a minimum for the power loss requires a high L/r ratio. At low frequencies, this ratio has a constant value, depending only on the total copper cross section of the coil and its shape. At higher frequencies, the coil resistance r increases because of eddy currents, as follows:

$$r = r_0(1 + k^2) \tag{12}$$

where k is the eddy-current factor expressed as

$$k = \frac{0.04N'd^3}{l} f$$

and r_0 = direct-current resistance, ohms
0.04 = constant for particular coil shape
N' = total number of insulated wire strands in cross section of coil
d = strand diameter, inches
l = effective length of coil ($a + b$ in Figure 6), inches
f = frequency, cycles per second.

It is apparent from (12) that operation at high frequencies requires a small wire or strand diameter d. If it is desired to use Litz wire, we may select No. 41 enamel wire as the smallest desirable wire for strands, but are at liberty to use a single wire or parallel wires (Litz) per turn, thus being able to vary L and f without affecting the copper cross section or any of the remaining factors which determine k in (12). The coil in Figure 6 contains 4200 strands of No. 41 wire in its cross section; i.e., it may be given as 4200 turns of single No. 41 wire or 2100 turns with 2 parallel strands of No. 41 wire, etc. The tuning capacitance C is estimated to be $C = 7$ micromicrofarads (coil capacitance = 3 micromicrofarads).

The lowest operating frequency of the circuit with $N = 4200$ turns ($L = 387$ millihenries) is 96 kilocycles, at which the eddy-current factor k^2 has still a negligible value ($k^2 = 0.037$). The equivalent shunt resistance R has, therefore, the optimum value obtainable with this coil size: $R = 22.5$ megohms but the value of Q is only 97. Figure 7 shows the results of paralleling strands to vary L and f as explained above. The shunt resistance R decreases to 50 per cent of its optimum value at the frequency where Q goes through a maximum.

A good compromise between efficiency and voltage regulation, which depends on coupling and Q values as explained, indicates 1400 turns of 3-strand Litz wire with $L = 43$ millihenries, a resistance $R = 17.5$ megohms, $Q = 227$, and an operating frequency $f = 288$ kilocycles. Other factors, such as winding time, cost of wire, etc., may influence this choice. The maximum peak voltage for $P = 6.5$ watts is, hence, $E_{\max} = 15$ kilovolts for this coil.

TUBES AND CIRCUIT ASSEMBLY

Efficient operation of the oscillator tubes requires class C excitation and low plate-voltage loss. Beam power tubes such as the 6L6 and 6Y6G are, therefore, especially suitable for use at low supply voltages.

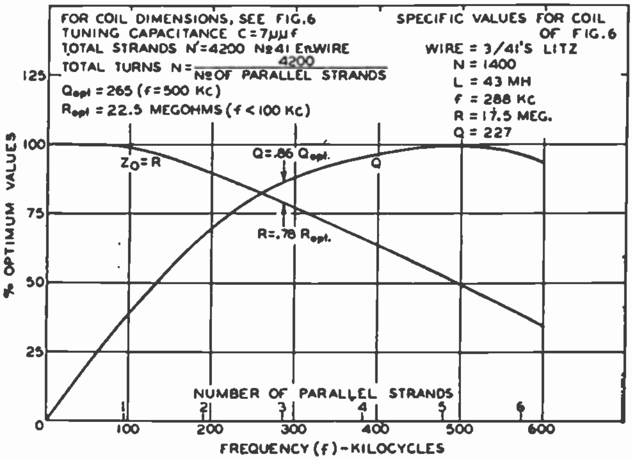


Fig. 7—Efficiency and Q values of coil (Fig. 6) versus number of strands per turn for a fixed product (strands × turns).

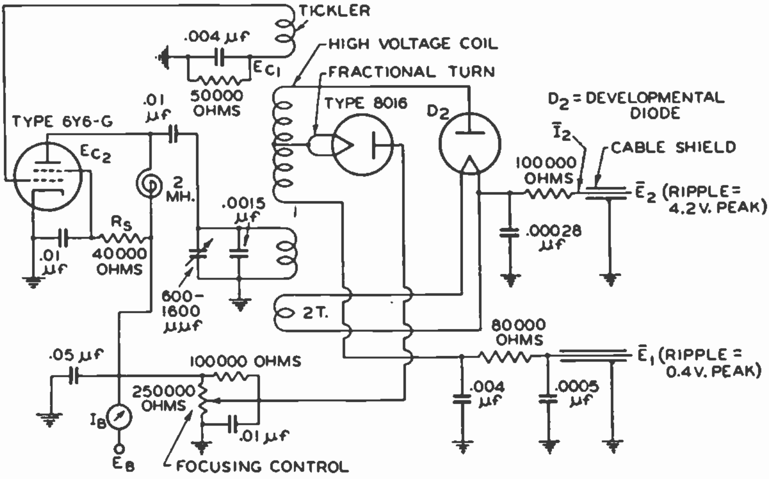
The 6Y6G can furnish 15 watts power with 75 to 85 per cent efficiency at voltages between 300 and 375 volts. The grid-leak bias should be $E_{c1} = 2E_{cc0}$. The screen-grid voltage is made self-regulating by a series resistance R_s (Figure 8). It varies from approximately 65 volts at no load to 120 volts at full load and, thus, aids the voltage regulation of the supply. Larger output powers require parallel operation of tubes.

HIGH-VOLTAGE RECTIFIER TUBES

Standard high-voltage rectifiers such as the 2X2 or 2V3-G require considerable heater power and are not designed for high-frequency operation. The development of special diodes for rectification of high radio-frequency voltages was therefore indicated. The RCA-8016 requires a cathode power of only one-quarter watt and thus permits economical radio-frequency heating from the oscillator source.

Smoothing Filter Requirements

The filter capacitances have small values because of the high operating frequency (300 kilocycles for a 10-kilovolt supply). In contrast to conditions with 60-cycle operation, the ripple voltage is determined substantially by the ratio of the sum of diode and stray capacitances to the filter-condenser capacitance. The actual ripple percentages must be of considerably lower value than in 60-cycle filters to avoid capacitive coupling and interference with receiver operation. Typical values are given in Figure 8 for a 10-kilovolt kinescope supply.



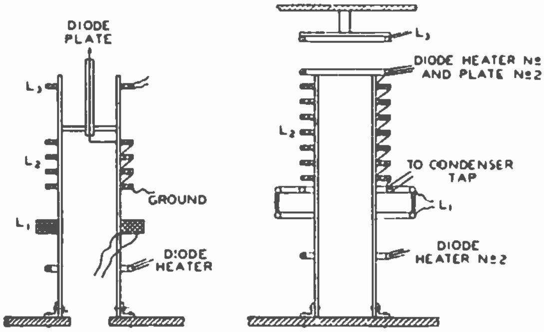
OPERATING CONDITIONS (APPROX.)

E_B VOLTS	I_B MA.	E_{C_2} VOLTS	E_{C_1} VOLTS	$\bar{E}_1$ KV.	$\bar{E}_2$ KV.	$\bar{I}_2$ MA.
+325	31	+68	−36	2	10.6	0
+325	62	+115			9.4	0.95

Fig. 8—Circuit and operating conditions of the 10-kilovolt supply for kinescopes.

Circuit Assembly

The particular form of the transformer assembly depends on the type of circuit and the required sparking distances. Typical assemblies are shown in the sketches of Figure 9. The operation at high radio-frequency voltages emphasizes corona effects because of increased



COIL ASSEMBLY FOR HALF-WAVE DIODE CIRCUIT

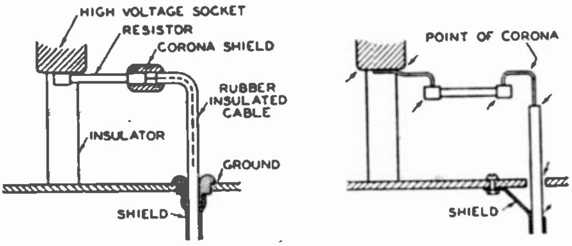
COIL ASSEMBLY FOR VOLTAGE-DOUBLING CIRCUIT

Fig. 9—Typical high-frequency transformer assemblies.

dielectric losses in ionized air. The fine-wire, high-potential ends of transformer windings, must, therefore, be protected against power loss and destructive effects due to corona by guard rings or conductors of sufficient radius of curvature as illustrated in Figures 9 and 10. This requirement also includes diode terminals and filter circuit.

Developmental Voltage Supplies

A very small voltage supply for iconoscopes is shown in Figure 11. It was built several years ago for battery operation and is housed in a coil shield 2½ inches in diameter. The 955 oscillator tube takes 8 milliamperes at 180 volts to supply 1 kilovolt to a bleeder circuit and iconoscope. The operating frequency is 1.2 megacycles. The small diode is an experimental tube. The larger kinescope supply shown in Figure 12 operates between 7 kilovolts and 12 kilovolts and measures 7¾ × 4¼ × 9 inches. The supply includes the oscillator tube, which is separated by a heat shield from the transformer assembly. The housing is ventilated at the oscillator but otherwise closed, to prevent



GOOD CABLE CONNECTION

POOR CABLE CONNECTION

Fig. 10—Corona shielding of cable connections.

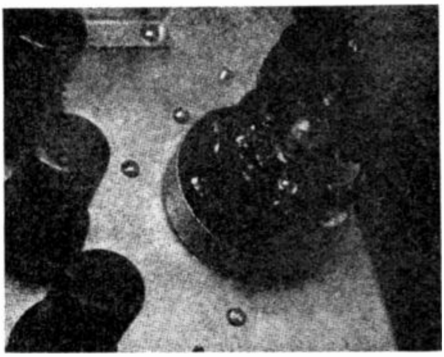


Fig. 11—A 1-kilovolt high-voltage supply for iconoscopes.

dust precipitation on the high-voltage conductors. Operating data are given on the circuit diagram in Figure 8.

A 30-kilovolt projection-tube supply with separate oscillator for the focusing voltage is shown in Figure 13. Transformers and rectifier assembly are housed in dust-tight shields. The outside dimensions of the second anode supply are 11 × 11 × 12 inches high. The focusing voltage can be varied from 4 to 7 kilovolts by tuning the primary of its oscillator circuit. The main second-anode supply employs a voltage-doubling circuit energized by three parallel 6Y6G oscillator tubes. Both supplies are operated in series to maintain a desired voltage ratio under varying load conditions. Circuit and performance are shown in Figures 14 and 15.

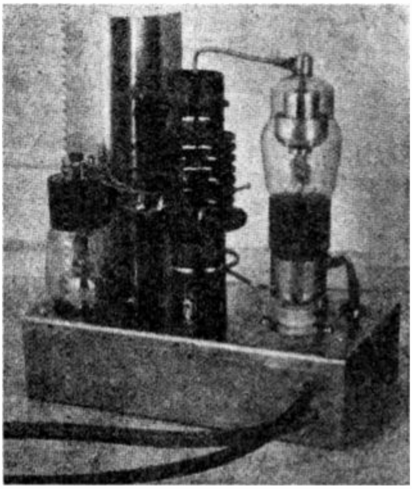


Fig. 12—A 10-kilovolt high-voltage supply for kinescopes.

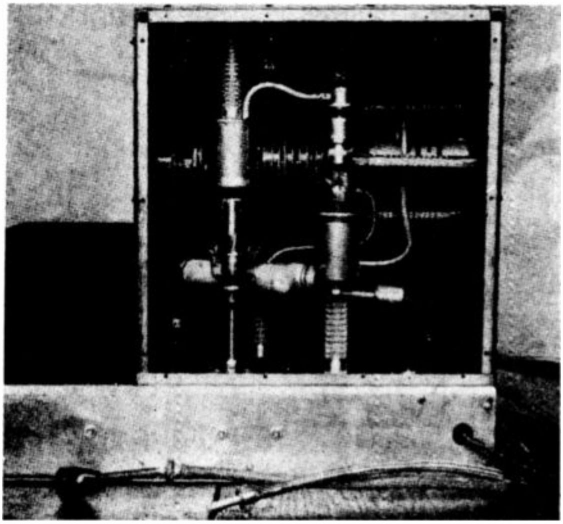


Fig. 13—Arrangement of a 30-kilovolt voltage-doubling circuit.

A number of radio-frequency-operated supplies for various voltages have given trouble-free service in the laboratory and in television equipment. Voltage stability and focus regulation under actual operating conditions are quite satisfactory. Little difficulty was experienced

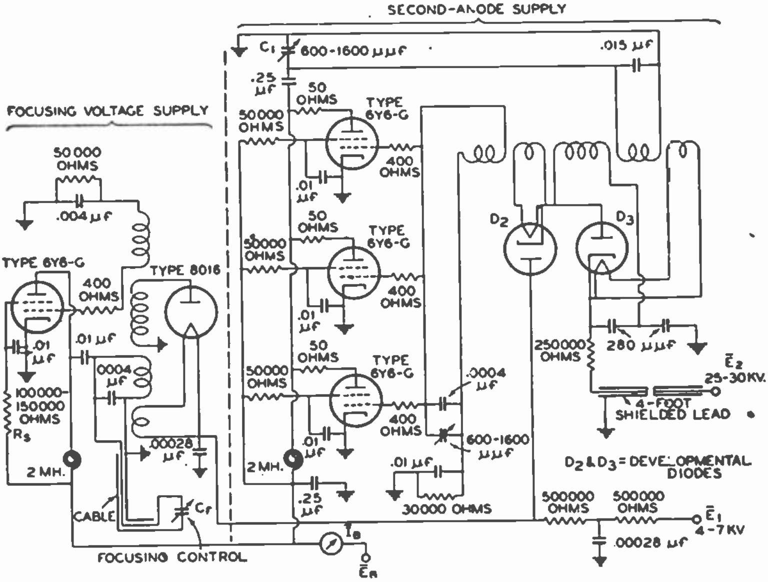


Fig. 14—Circuit of the 30-kilovolt supply for projection kinescopes.

in preventing oscillator interference with television equipment but isolating resistors or chokes may be required when a single source of filament or B supply is used.

CONCLUSIONS

Actual performance has proved that high-voltage supplies energized by vacuum-tube oscillators at high frequencies are practical as kinescope and iconoscope supplies. The obtainable power output is limited by the oscillator power. This method permits the construction of safe supplies, where the current requirements are not too high. This low-power reserve also protects the kinescope and rectifier in case of spark-over or accidental short circuits because of the small short-circuiting current. It must be remembered, however, that currents of dangerous magnitude are obtainable, depending on the voltage step-up and the oscillator power. For such conditions, due precautions for safety must be taken.

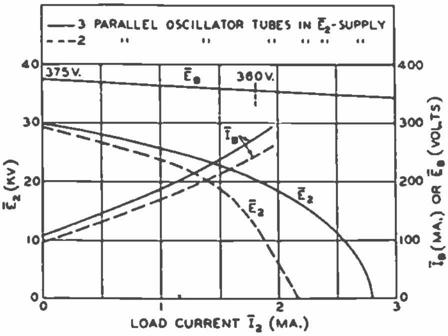


Fig. 15—Regulation characteristics of the 30-kilovolt supply.

The cost of high-frequency-operated supplies compares favorably with 60-cycle supplies when the oscillator power is moderate and when, consequently, small oscillator tubes can be used. Kinescope supplies for voltages up to 30 kilovolts and approximately 50 watts output are in this range.

A TYPE OF LIGHT VALVE FOR TELEVISION REPRODUCTION*†

By

J. S. Donal, Jr.# and D. B. Langmuir‡

Summary—The desirability of a light valve for the reproduction of television pictures is discussed, and the use of a suspension of opaque platelike particles for this purpose is shown to offer the particular advantages that the electron beam would be only a control mechanism and the picture brightness would be limited only by the light source and lens system.

The theory of operation of such a suspension is described and it is demonstrated that inertial effects may be neglected and that the rate of orientation of the particles is independent of particle size and is a function of the viscosity and dielectric constant of the suspending medium and of the square of the applied voltage. The contrast ratio obtained may be made very high, although the optical efficiency will decline as the contrast ratio rises.

It is found that suspension resistivity must be considered in practical application of the light valve, for if the field is applied through an insulating wall the valve will respond only to changes in potential of the outside of this wall, since leakage will prevent a constant wall potential from maintaining a field across the suspension.

From the results of tests, the conclusions are drawn that the fundamental optical behavior of the suspensions considered is in accordance with the predictions of a theory based on simple assumptions, and that the suspensions fulfill the basic requirements of a television light valve.

Introduction

THE development of television has seen progress in the reproduction of images, from the early scanning disks to the modern kinescopes, or cathode-ray tubes. Because of the limitations of kinescopes in the production of very large images, alternative reproduction devices have been studied.[1] This paper presents a description

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* Decimal classification: R583 × R388.

† Presented at the Fifteenth Annual I.R.E. Convention, in Boston, Massachusetts, on June 29, 1940. Reprinted from *Proc. I.R.E.*, May, 1943. This paper reports work carried on, prior to 1940, as part of the television-development program of the RCA Manufacturing Company, Inc.

Research Department RCA Laboratories Division, Princeton, N. J.

‡ Formerly, Research Laboratories, RCA Manufacturing Company, Inc., Harrison, New Jersey; now the Office of Scientific Research and Development, Washington, D. C.

[1] Dr. Rosenthal has described an alternative system, "the Skiatron", which depends upon the development of opaque areas in microcrystalline layers of ionic crystals under the action of electron bombardment. See A. H. Rosenthal, "A system of large-screen television reception based on certain electron phenomena in crystals", *Proc. I.R.E.*, vol. 28, pp. 203-213; May, 1940.

of a type of cathode-ray-controlled light valve which appears to offer promise in this field.

Since a new type of television reproduction device is described, it may be helpful to compare its principles with those of other reproduction systems. Three examples will be considered: the combination of a scanning disk with a single light source modulated by a video signal; a nitrobenzene Kerr cell combined with a scanning disk; and a kinescope.

The system using the scanning disk has two major disadvantages. The light which produces the picture must be generated by power modulated at video frequencies. Also, light can be delivered to only one picture element at a time and the resulting picture must suffer in respect to brightness by a factor of many thousands relative to the limiting brightness which can be obtained with a lantern slide where, for example, light from every element reaches the observer's eye continuously.

The conventional Kerr cell has the theoretical advantage that the video signal is used as a control rather than as a generator of light. However, it suffers from the same unfavorable efficiency factor as does the scanning disk in that light reaches the observer's eye from only one picture element at a time.

In the kinescope, the light must be generated by modulated power, and it can be generated at only one picture element at a time, so that both of the disadvantages of the scanning disk are present. The success of the kinescope in television reproduction is due chiefly to certain special features of the device. Cathode-ray beams make it possible to concentrate power into small areas very effectively. The combination of such large power densities with phosphors which can efficiently transform the power into light produces a satisfactory result in spite of the theoretical disadvantages.

It is worth while to consider systems of television reproduction which are free from both of the stated handicaps. An ideal device, which may be called a television light valve, can be conceived as a lantern slide which at every point has an opacity that can be controlled instantaneously. Such a slide might be scanned so that the transparency varies from point to point in accordance with the picture signal. If the transparency remained constant at the value set by the signal until the return of the beam during the next scan, light would reach the observer's eye continuously from all of the desired elements of the picture. The brightness of the picture would be limited only by the light source, the lens system, and the maximum transparency of the light valve; the electron beam would be only a control mechanism.

In exploring the groundwork for such a hypothetical system the variety of possibilities is large and it is difficult to limit them in any preconceived manner. However, confining the scanning systems to electron beams, and the controlling effects to voltages developed by the beams, certain almost inescapable features of a practical light valve can be defined in advance. First, the layer of optically controllable material must be thin, of the order of a few thousandths of the width or length. Second, the light-control effect must result from an electric field which is parallel to the direction of light transmission. While neither of these requirements is absolute or final, the complexities introduced by deviating from either of them militate against the practicability of a light valve which fails to meet them.

It is illustrative to consider the conventional type of nitrobenzene Kerr cell in the light of these conditions. Since such a cell must be at least several millimeters thick in the direction of light propagation in order to obtain useful transmission with the maximum electric field permitted by the dielectric strength of nitrobenzene, and since the lines of force must run perpendicular to this direction, neither basic requirement is fulfilled. The classical Kerr effect seems unsuited to the television light-valve problem.

In respect to frequency response the requirements for the optical medium in the present problem are much less stringent than those for a light valve used in conjunction with a scanning disk. Since any given picture element need respond only once during each frame period, the time of respose of the light valve may be 1/100 second or even longer. The system may be considered to consist of a large number (one for each picture element) of parallel communication channels of very narrow bandwidth. Taken all together these transmit the same amount of intelligence as a single wide-band channel.

An optically sensitive medium which satisfies the basic requirements outlined and which has been found to possess many advantageous qualities consists of a suspension of small, flat, opaque particles in an insulating fluid. If the suspended platelike particles have a dielectric constant greater than that of the liquid, an electric field will cause them to align themselves parallel to the field. The shadow cast by the particles and, hence, also the light transmission of the suspension will therefore be subject to control by the field.

Theoretical and experimental investigations of such suspensions are described in this paper. A separate paper[2] treats the methods by which cathode-ray beams can be made to control a light valve and

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[2] J. S. Donal, Jr., "Cathode-ray control of television light valves", *Proc. I.R.E.*, this issue, pp. 195-208.

describes a cathode-ray-controlled suspension light valve with which a high-definition television picture has been projected on a screen.

THEORY

Consider a suspension in which there are n absorbing particles per cubic centimeter, each particle having a projected area a on a plane perpendicular to the x direction. Light traveling in the x direction will vary in intensity L according to the familiar exponential absorption law

$$\frac{dL}{dx} = -naL. \tag{1}$$

Integrating, we have

$$\frac{L}{L_0} = e^{-nax}. \tag{2}$$

Here L is the light intensity after traversing the distance x in the suspension, while L_0 is the incident light intensity at x equals zero.

The quantity nax is equal to the sum of the projected areas of all particles in a suspension thickness x. Letting this total area equal A, we have

$$\frac{L}{L_0} = e^{-A}. \tag{3}$$

Now let A_1 be the total projected area of the particles in the unoriented condition and let A_2 be their projected area after having been aligned by the field, with L_1 and L_2 representing the corresponding light intensities after transmission through a suspension thickness x. Then, from (3),

$$\frac{L_2}{L_1} = e^{-(A_2-A_1)}.$$

The ratio L_2/L_1 is equal to the maximum contrast ratio obtainable while L_2/L_0, the ratio of emergent to incident light when all particles are lined up, may be called high-light transmission or optical efficiency of the suspension. Designating these by C and E, respectively, and substituting, we obtain

$$C = e^{-A_2(1-(A_1/A_2))} \tag{4}$$

$$= E^{(1-(A_1/A_2))}. \tag{5}$$

This relationship between contrast, optical efficiency (or high-light transmission L_2/L_0), and the shape factor A_1/A_2 of the individual particles is presented graphically in Figure 1. It is seen that the contrast ratio is by no means limited to a value equal to the shape factor, and that a contrast ratio of any desired magnitude can be obtained simply by increasing the total area of the suspended particles per unit area of the valve. This can be done, for example, by increasing the thickness of the fluid layer, or by increasing the concentration of the particles. The optical efficiency will decline as the contrast ratio rises at a rate depending upon the shape factor of the particles.

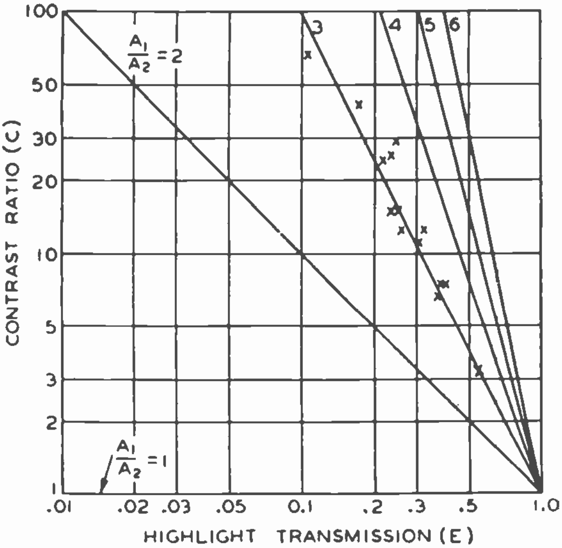

Fig. 1—Curves showing the contrast ratio obtainable from a suspension as a function of its light transmission in the completely oriented state for various values of the shape factor of the particles. The lines are theoretical. Experimental points for a suspension of graphite in castor oil are also shown.

The rate at which the particles are oriented by an applied field will depend upon their size, shape, and moment of inertia, and upon the fluid viscosity, dielectric constant, and density. An exact theory would be complicated. It can be shown, however, that inertial forces in the motion are very small compared to viscous ones, and need be considered only when the Reynolds number[3] R has a value of the order of magnitude of unity. The Reynolds number appropriate to this problem is

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[3] The significance of the Reynolds number in problems of this nature is stated at length by Sir Horace Lamb in his "Hydro-dynamics", sixth edition, Cambridge University Press, Cambridge, England, 1932.

$$R = \frac{\rho\omega r^2}{\mu}$$

where (in the centimeter-gram-second system of units)

ρ = fluid density
r = equivalent particle radius
ω = angular velocity of particle
μ = coefficient of viscosity

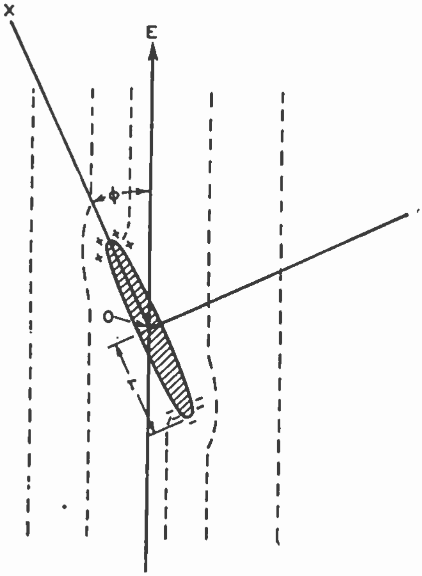

Fig. 2—A platelike particle is viewed edge on in an electric field. The field component parallel to the surface induces charges on the particle and the field component perpendicular to the surface acts upon these induced charges to produce a torque which tends to rotate the particle so as to place its long axis parallel to the electric field.

For the range of values of the four variables which might be practicable in a television light valve R is much less than 1, so that viscous effects predominate overwhelmingly.

Since inertial effects may be neglected, the rate of orientation of the particles can be calculated quite simply if conducting plates, such as those of graphite, are considered. Such a particle is shown in an electric field in Figure 2. The plane of the flat body is shown perpendicular to the page and its periphery may have any shape whatever.

Consider the field E resolved into components $E \cos \phi$ along OX

and $E \sin \phi$ along OY. The former may be considered to build up polarization charges on the plate so as to give it a double moment of strength M. The latter may be regarded as exerting a torque upon this dipole without affecting the value of M. If the size of the plate (proportional to r) is varied, keeping the particle shape factor and the electric field E constant, the surface charge density at corresponding points will remain constant. The total polarization charges will therefore be proportional to the area of the plate and, of course, to ϵ, the dielectric constant of the medium. Since the separation of positive and negative charges will be proportional to r, it is clear that

$$M = K_1 \epsilon r^3 E \cos \phi$$

where K_1 is a constant depending upon the shape of the periphery of the plate.

The torque exerted on the plate by the field will be

$$\begin{aligned} L_E &= M\,E \sin \phi \\ &= K_1 E^2 \epsilon r^3 \sin \phi \cos \phi \\ &= K_2 E^2 \epsilon r^3 \sin 2\phi. \end{aligned}$$

This torque will be opposed by a torque due to viscous drag, for if the same plate as is shown in Figure 2 is considered to be rotating in a viscous fluid, a torque will be exerted on it by the drag of the fluid. Since the inertial forces may again be neglected, it can be shown from dimensional reasoning that

$$L_\phi = K_3 \mu \frac{d\phi}{dt} r^3$$

where μ is the coefficient of viscosity.

The reason for the third-power variation with r can be summarized qualitatively as follows: As r increases with constant μ and $d\phi/dt$, the area upon which shearing or pressure forces are exerted increases as r^2, while the lever arms by which the forces must be multiplied to give the torque increase as r. With constant $d\phi/dt$ the rate of shear at corresponding points in the fluid remains constant as r increases since the increased velocity of parts of the system is counterbalanced by the increased scale of size of the system. Therefore, the net result is a torque increasing as r^3.

Equating these torques and rearranging, we obtain

$$\frac{d\phi}{dt} = K \frac{\epsilon}{\mu} E^2 \sin 2\phi.$$

This equation can be integrated, yielding

$$\tan \phi = \tan \phi_0 e^{-2ct} \tag{7}$$

where $c = K(\epsilon/\mu) E^2$, t is the time, and ϕ_0 is the value of ϕ when $t = o$.

Since both the torques involved increase as the cube of the particle size, the rate of orientation is independent of the particle size. The time response of the suspension light valve depends therefore upon the square of the applied voltage and upon the viscosity and dielectric constant of the suspending medium.

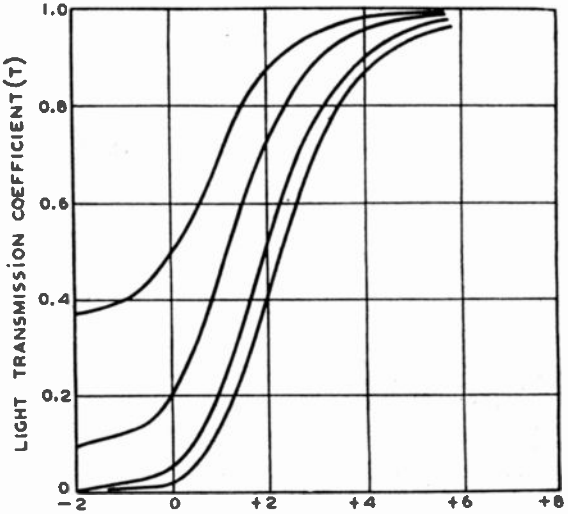

Fig. 3—Theoretical curves of light transmission as a function of time t for a suspension of particles of infinite shape factor, under a constant orienting field. The constant $c = E^2(\epsilon/\mu)$. At $ct = 0$ all particles lie at 45 degrees to the field.

The projected area of the suspended particles is equal to $A_0 \sin \phi$, where A_0 is the total area presented by the deoriented particles in 1 square centimeter of valve area. It is assumed that this projected area equals zero when alignment is complete. Therefore, the transmission coefficient T of the suspension as a function of time will be given by

$$T = e^{-A_0 \sin \phi} \tag{8}$$

where $\tan\phi = \tan\phi_0\, e^{-2ct}$ from (7).

The curves of Figure 3 show the relations, derived from (8), between the transmission coefficient and the time for various initial values of light transmission in the deoriented condition.

For comparing theory with observed data, it is convenient to eliminate ϕ from (7) and (8). When this is done the equations may be put into the form

$$\ln\ln\frac{1}{T} - \ln A_0 = \tfrac{1}{2}\ln\left(1 + e^{4ct}\right). \tag{9}$$

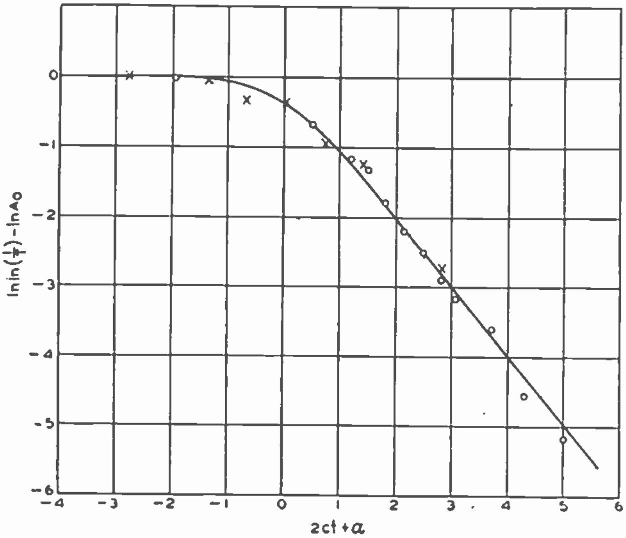

Fig. 4—Comparison between test and theory of light-transmission-versus-time curves. The two adjustable constants used in fitting the experimental points to the curve are discussed in the text.

The constant ϕ_0 is arbitrary and has been set equal to $\pi/4$ in this case.

This equation is shown plotted in Figure 4, together with observed points from two suspensions. The constant c, which is indeterminate because of its dependence upon the shape factor of the particles, has been adjusted to give the best fit. In addition the points have been shifted horizontally by means of the constant α so as to make the curves coincide. The latter procedure is necessary since the time scale for the theoretical curve is essentially a relative one. An absolute origin for the time is indeterminate because we have no way of measuring the value of ϕ when the electric field is first applied. The theory is at best approximate because of the adjustable constants and because of the assumption (not consistent with actual suspension characteristics) that the transmission coefficient approaches unity with complete orientation. The general behavior of the suspensions is seen, however, to be in conformity with that predicted by simple calculations.

TESTS

Observations of suspension characteristics were made using cells with glass walls. The thickness of the fluid layer was varied from a few millimeters down to about 0.1 millimeter. Potentials were applied by means of semitransparent sputtered metal coatings in direct contact with the suspension. Light intensities were measured with a phototube. When transient effects were studied this tube was connected to the input of a direct-current amplifier which controlled an oscilloscope having a long-persistent screen on which single traces could be studied.

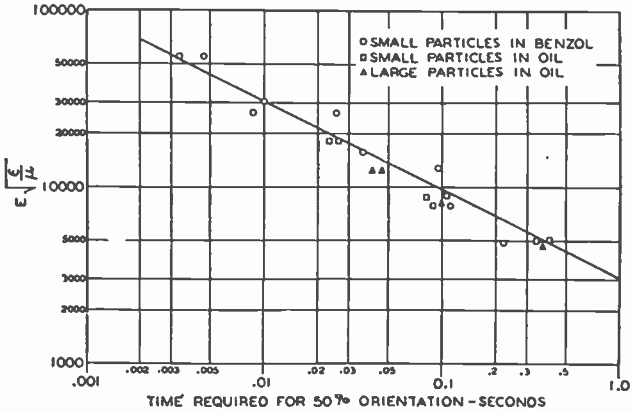

Fig. 5—Time required for light transmission to rise to half of its final value as a function of the parameter $E\sqrt{\varepsilon/\mu}$. This curve checks the square-law response to field strength, the effect of dielectric constant and viscosity, and the independence of particle size upon rate of orientation.

In one series of tests the light transmission in the deoriented and completely oriented states was measured for several different values of opacity controlled by the concentration of particles or the thickness of the cell. The results for a suspension of graphite in castor oil are shown as the crosses in Figure 1. The theoretical relationship between contrast and optical efficiency is seen to be verified. The effective shape factor of the particles is apparently between 3 and 4. A contrast ratio of 25 with an efficiency of 0.2 is seen to be obtainable.

Figure 5 is a plot of the time required for the transmission of the suspension to reach half of its final value against the parameter $E\sqrt{\epsilon/\mu}$. The parameter was varied not only by applying different field strengths to the suspension, but by using fluids of different viscosity and dielectric constant. Suspended particles of differing sizes were also used. The time of orientation is seen to be proportional to the inverse square of $E\sqrt{\epsilon/\mu}$ and to be independent of the particle size, as predicted by theory.

Figure 4 showed a check of the detailed form of the transmission-versus-time curves. The theoretical curve as represented by (9) is shown by the solid line. Points measured experimentally with two different suspensions are also shown, and the agreement with theory is satisfactory.

POLARIZATION EFFECTS

In the experiments just described the potential difference was applied to the suspension from electrodes in contact with the fluid. The behavior observed was not a function of the conductivity of the suspension, the various values of fluid conductivity resulting merely in the passage of a greater or smaller conduction current. However, in important practical cases, as already described,[1] the potential differences may not be applied directly to the fluid. One wall of the light valve might consist, for example, of mica or of thin glass which is charged on its outer face by an electron beam. In that case the conductivity of the suspension will have a marked effect upon the performance; therefore, it seems worth while to discuss the broader aspects of this phenomenon at this point.

Consider the behavior of a cell of which one electrode is in direct contact with the suspension and is grounded while the other electrode is the charged outer surface of a perfectly insulating wall. Two cases will be discussed. In the first, the potential difference between the two electrodes is held constant for a time long compared to the relaxation period of the suspension, while in the second case the charged electrode is allowed to float after bringing its potential to a certain value. In both cases the general behavior of the suspension will be the same; that is, when the charge is first applied a field will be established across the valve causing it to light up, but the potential drop across the leaky fluid will decline with time. If the particles are now deoriented the valve will remain dark until the charge is removed from the electrodes, at which time another flash will occur due to the depolarization field set up when the charge on the inner wall of the valve leaks off.

The differences in detail, depending upon whether the outside of the insulating wall is held at a constant potential during the relaxation period, may be analyzed quantitatively in terms of the circuit of Figure 6. The condenser C_1 corresponds to the insulating wall, while the leaky condenser C_2 represents the valve fluid.

Suppose a voltage is applied suddenly to the circuit (switch in position 1), held constant for a time long compared to the relaxation period, which in this case is $R\ (C_1 + C_2)$, and is then removed suddenly by changing the switch to position 3. The resulting changes of the

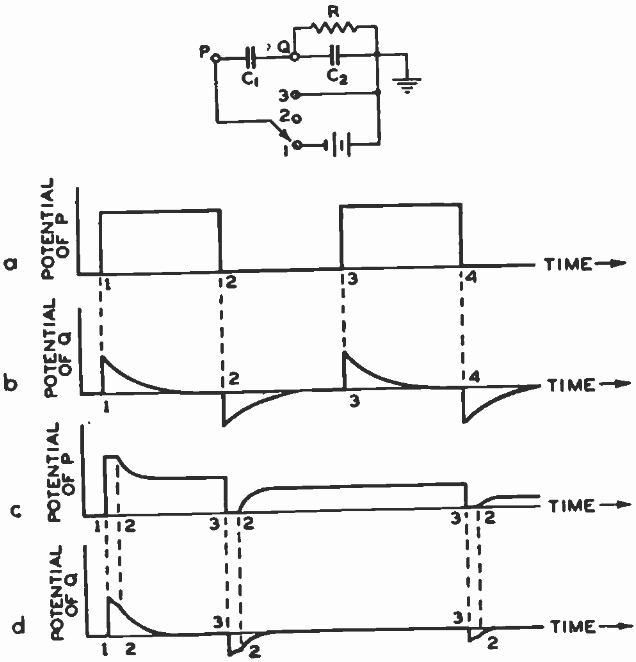

Fig. 6—Analogue to the behavior of a leaky suspension.
(a) and (b) show the potential changes (assumed to be with respect to ground) when a constant potential difference is applied through an insulating wall and then removed; (c) and (d) show the potential changes when the suspension is actuated by charging the outside of an insulating wall which is then allowed to float until the charge is removed. The numerals on the abscissas indicate the positions of the switch in the diagram above.

potential of the point P, with respect to ground, are represented in Figure 6(a). The potential differences across C_2 may be represented by the variations of the potential of the point Q with respect to ground shown in Figure 6(b).

After the initial potential difference across C_2 has fallen to zero, the full applied voltage will be supported by the insulating wall. At the removal of the applied potential the voltage across C_2 due to depolarization will be equal and opposite to that resulting from the initial voltage application. This reversed voltage falls to zero by leakage and the valve returns to its original potential distribution. The single removal of the applied potential by grounding point P completely discharges the valve provided P is kept grounded until the reversed potential across the suspension is reduced substantially to zero by leakage. Regrounding of point P at a later time would have no additional effect if there is no recharging in the meantime. Also, a subsequent application of a potential difference across C_1 and C_2 of a value equal to that applied initially (as shown in the second cycles of Figure 6(a) and Figure 6(b)), or of a different value, would produce proportionate potential differences across the valve suspension, uninfluenced by the potential differences of any preceding cycle.

The second of the cases mentioned above will now be considered. If instead of holding the voltage across the cell constant, we place the switch in position 2 after bringing the voltage to a steady value, the potential differences across the suspension will differ, as shown in Figure 6(c) and Figure 6(d), from those considered above. This case may be realized in practice if an area of the light valve is charged by an electron beam which then goes on to scan other valve areas. If an amount of charge sufficient to cause an initial potential difference V_0 between P and ground is put on condenser C_1 at P, the initial potential across C_2 is of course $V_0C_1/(C_1 + C_2)$ as before. However, this will decay exponentially at a rate more rapid (Figure 6(d)) than in the case considered above since C_1 and C_2 are no longer effectively in parallel between point Q and ground and the time constant will now be reduced to RC_2. Instead of the entire initial potential difference being impressed across C_1 as a result of leakage through the suspension, the potential difference across C_1 will remain unchanged by the leakage through R, and, therefore, the potential of P above ground will fall from V_0 to the asymptotic value $V_0C_2/(C_1 + C_2)$ as indicated in Figure 6(c). If now the valve is discharged by bringing the point P to ground potential suddenly and then allowing it to float again, a reversed potential difference will appear across C_2, (Figure 6(d)), but of a magnitude smaller than that which occurred in the first direction, since the change in potential of P by grounding it is now only $V_0C_2/(C_1 + C_2)$ instead of V_0, and the fraction $C_1/(C_1 + C_2)$ of this appears across C_2. This potential difference is then reduced to zero by suspension leakage.

It will be remembered that in the first case considered, the changes in potential produced by subsequent charging or discharging of the point P were independent of the first cycle just considered because the point P was kept grounded until the reversed potential difference across C_2 had disappeared. Now, however, the situation is different, for the potential of P drifts away from ground (Figure 6(c)) as the charge on C_2 associated with the reversed potential difference leaks off. Thus, if the point P is again carried to ground potential (Figure 6(c)) without an intermediate recharging, a second although still smaller reversed potential difference appears across C_2 (Figure 6(d)). On repeated discharge of P to ground, these reversed potential differences across C_2 would approach zero asymptotically. Furthermore, if before this has occurred and before the potential of P is in equilibrium at ground potential, a new potential difference V_1 (not shown in the figure) is applied between P and ground, the change in potential of P will be less than V_1 and, hence, the new potential difference developed across C_2 will be less than $V_1C_1/(C_1 + C_2)$.

From the practical standpoint, this phenomenon results in a delay in the response of the light valve coming into equilibrium with the impulses applied. This subject is discussed at greater length elsewhere.[1]

DISCUSSION

From the foregoing it will be seen that the suspension of opaque platelike particles fulfills to a considerable degree the requirements of the ideal light valve in which the opacity of a thin sheet may be varied from point to point to reproduce the lights and shades of a picture.

The change in opacity of the suspension under the action of an applied potential difference, which limits the contrast obtainable, may be made as high as desired by increasing the concentration of the suspended particles. Although the optical efficiency declines with increasing contrast, the results show that with particles of easily obtainable shape factor and with sheets of suspension thin enough to afford the possibility of reasonable resolutions, high contrast can be obtained with an optical efficiency which is not unreasonably low.

Since any given point of the layer of suspension must respond only once in each frame time, the rate of response of the suspensions investigated appears to be fully adequate for television purposes. This is particularly true since the sheet of suspension must be thin, of the order of a line width in thickness, if adequate resolution is to be obtained. This requirement means that high fields will result from relatively moderate potential differences across the suspension layer.

Since the rate of orientation was found to be independent of the particle size, the principle limitation on particle size is that the individual particles shall not be evident in the reproduced picture. In general this condition will be satisfied if the particles are small compared to the size of a picture element.

Although polarization effects have been shown to result in a delay in the attaining of an equilibrium between the response of the valve and the potential differences applied, the effects of these residual potential differences may be expected to be reduced in importance by the fact that the suspension responds to the square of the applied difference in potential.

REFLECTIVE OPTICS IN PROJECTION TELEVISION*†

By

I. G. MALOFF# AND D. W. EPSTEIN‡

Summary—Development of a process for molding large aspherical correcting lenses from clear plastics now makes projection television techniques economical and practical for home receivers as well as theater systems. Optical principles, mechanical mounting problems, design of correcting lenses, molding methods and a receiver console arrangement are presented.

IT HAS been known for a long time that aspherical surfaces in combination with either spherical or aspherical mirrors may be arranged into optical systems of high aperture and high definition. Astronomers made use of this principle in an arrangement consisting of a spherical mirror and an aspherical lens; however, high costs and difficulties in constructing such systems prevented their general utilization.

In searching for efficient optical systems for projecting television images originating on screens of cathode-ray tubes, the principle of reflective optical systems has been made a subject of concentrated study and experimentation. This has resulted in the development of a number of reflective optical systems suitable for projecting television images with diagonals ranging from 25 inches to 25 feet. RCA systems consist of a spherical front surface mirror and an aspherical lens, positive in the central portion and gradually changing into negative near its periphery. The gain in illumination on the viewing screen with the new systems is about six or seven to one when compared with a conventional $f/2$ lens. The quality of the images obtained is comparable with images produced by conventional projection lenses.

The main handicap of the new system, the high cost of the aspherical lens, has been overcome by the development of machines for making aspherical molds and by development of a process for molding aspherical lenses from plastics. RCA reflective optical systems are designed for a fixed image distance and require cathode-ray tubes hav-

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Home Instrument Department, RCA Victor Division, Camden, N. J.

‡ Research Department, RCA Laboratories Division, Princeton, N. J.

ing face-curvatures fixed in relation to the curvature of the mirrors in the system. The last two factors, while limiting the versatility of a given system, appear to be a small price to pay for the manifold gain in light. The design, manufacturing, installation and servicing of the RCA reflective optical systems have been improved and simplified to such a point that these systems can be considered as proven tools in television and oscillographic techniques. Reflective systems designed for infinite throw have been already applied successfully to television outdoor pickup cameras with the same manifold gain in light.

ANALYSIS OF THE PROBLEM

The problem of projecting images originating on the screens of cathode-ray tubes has received a great deal of attention from investigators here and abroad over a period of years. It has been shown[1] that the space distribution of light emitted by the screen of a cathode-ray tube follows very closely the cosine or Lambert law of perfectly diffusing surfaces. When a lens such as the conventional motion-picture projection lens is used to project a cathode-ray tube image onto a viewing screen, the overall efficiency of such a system is extremely low.

In motion-picture projection most of the light striking the film is delivered to the viewing screen, except of course for the light absorbed by the darkened portions of the film, thus creating the picture itself. However, when projecting light from a perfectly diffusing surface onto a viewing screen by means of the same lens, much of the light is lost. For large magnifications the following relation holds:

$$\frac{\text{(lumens on viewing screen)}}{\text{(lumens on tube)}} 100\% = K \frac{1}{4f^2} 100\%$$

where K is the transmission coefficient of the lens and f is the $f/$ number of the lens.[2] Good, commercially available, treated projection lenses having a relative aperture of $f/2$ and a transmission coefficient of nearly 100 percent, collect from the tube and deliver at large magnification to the viewing screen only 6.25 percent of the light generated.[3]

The image on the face of the cathode-ray tube is obtained at a

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(1) Orth, R. T., Richards, P. A., and Headrick, L. B., Development of Cathode-Ray Tubes for Oscillographic Purposes. *Proc. I.R.E.*, 23, p. 1316, Nov., 1935.

(2) Maloff, I. G., and Epstein, D. W. ELECTRON OPTICS IN TELEVISION, McGraw-Hill Book Co., New York, N. Y., 1938.

(3) Maloff, I. G., and Tolson, W. A., A Resume of the Technical Aspects of RCA Theatre Television, *RCA REVIEW*, 6, p. 6, July, 1941.

relatively high cost in equipment, effort and power. Any increase in the brightness of this image may be obtained only at great cost from the standpoint of design and operation. For this reason, the problem of providing a more efficient optical projection system has received a great deal of attention. Improvement of a few percent was of no interest. A manifold increase in the percentage of light delivered to the screen was sought. The answer was finally found in modifying a principle known to astronomers and adapting it to the problem on hand.

For quite a long time,[4, 5, 6] astronomers and opticians have known that optical systems combining spherical and aspherical mirrors and surfaces are capable of working at very high relative apertures and at the same time are remarkably free from optical defects. Schmidt[6] applied this principle to astrophotography. The so-called Schmidt camera is an optical system (Figure 1) comprising a spherical mirror *A* and a weak aspherical lens *B* at the center of curvature of the mirror. Images of distant objects are formed on an image plate *C*, which in itself is part of a sphere of radius slightly larger than half the radius of the mirror and located at the focal point of the system.[7]

SYSTEM USED IN ASTRONOMY

Of the outstanding defects of the images formed by optical systems (spherical and chromatic aberation, coma, astigmatism, curvature of the field and distortion), only spherical aberration is distributed uniformly over the whole image field; all other defects increase with the distance from the axis. A spherical mirror has no axis and is, of course, achromatic. If a small aperture is placed at the center of curvature of a spherical mirror, then any narrow beam of parallel light coming from any direction through this aperture onto the mirror will focus at a point located on a sphere whose radius is equal to half the radius of curvature of the mirror. If the aperture is increased, spherical aberration becomes apparent and the quality of the image deteriorates.

The correcting lens in the Schmidt arrangement (shown in Figure 1) introduces into the incident beam an amount of spherical aberration which is equal to that introduced by the mirror but is opposite in sign. Thus, by placing a suitably shaped correcting lens at the center

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(4) Schwarzschild, K., "Theorie der Spiegeltelescope," *Gottingen Abhandlungen*, 2, 1905.

(5) Kellner, G. A. H., U. S. Patent 969,785, granted Sept. 1910.

(6) Schmidt, Bernard, Mitt, Hamburger Sternwarte in Bergedorf, 7, No. 36, 1932.

(7) Hendrix, D. O., and Christie, W. H., Some Applications of Schmidt Principle in Optical Design, *Sci. Amer.*, 161, p. 118, Aug., 1939.

of curvature of the mirror the non-aberration condition for all rays arriving at the mirror from distant objects may be retained. The system is then free from the spherical aberration, while coma, astigmatism and chromatic aberration introduced by the correcting lens are minimized by proper shaping of this lens.

SYSTEMS FOR TELEVISION PROJECTION

When a reflective optical system is used for projecting images originating on luminescent screens of cathode-ray tubes, the requirements which the optical system must fulfill are considerably different from those of the Schmidt camera. The most important difference is that the light from a point on the luminescent screen does not emerge from the optical system as a bundle of parallel light. On the contrary, it emerges as a bundle converging to a point or focus at a definite distance. This finite throw system is radically different from that of the infinite throw. The other difference is that the thickness of the glass face plate of the cathode-ray tube introduces a certain amount of spherical aberration, which has to be taken into account when balancing the spherical aberration of the correcting lens against that of the mirror.

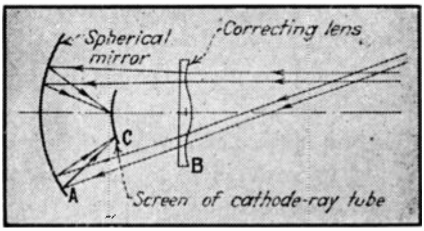

Fig. 1—Optical system of the so-called Schmidt astronomical camera, adapted by RCA for use in projection television systems.

The outstanding advantage of an optical system such as that shown in Figure 1 over a more conventional optical system is its ability to focus a large field (large tube diameter) with a large relative aperture. As was mentioned already, such a system possesses this property primarily because a spherical mirror with an aperture located at the center of curvature of the mirror suffers from only two aberrations, spherical aberration which is uniform all over the field, and curvature of the field. This may be seen from Figures 2 and 3, where C is the center of curvature of the mirror and O_1 and O_2 are object points located on the axis and off the axis respectively.

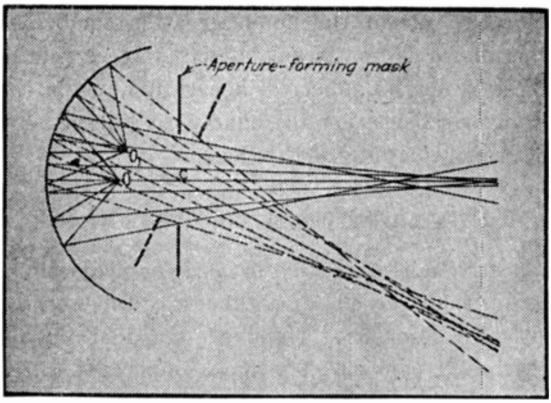

Fig. 2—Spherical mirror with an aperture at its center of curvature.

Figure 2 shows the ray paths for these two object points with the aperture located at the center of curvature. It is seen that the image or rather the circle of least confusion, since spherical aberration is present, is practically of the same size and symmetry for both object points. The reason for this is that the principal ray, i.e., the ray passing through the object point and center of the aperture also passes through the center of curvature of the mirror, and is therefore also an axis of symmetry for the sphere. The only difference is that the circular aperture mounted perpendicular to the principal axis and therefore symmetrically located with respect to the principal axis is non-symmetrically located with respect to the auxiliary axis. This causes some non-symmetry in the light distribution of the circle of least confusion but this non-symmetry becomes of importance only in the case of very large fields (large objects).

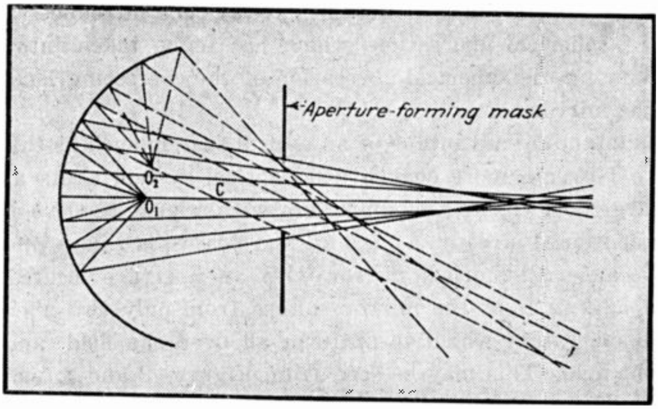

Fig. 3—Spherical mirror with an aperture that is not at the center of curvature.

Figure 3 shows the imaging properties of a mirror with the aperture located not at the center of curvature. It is seen that there is barely any sign of image formation for the off-axis object point.

PURPOSE OF CORRECTING LENS

The object of the correcting lens is to correct for the spherical aberration of the mirror without introducing any serious aberrations of itself. This is accomplished by making the correcting lens as weak as possible and locating it in the plane of the aperture at the center of curvature. In this way, the symmetry property of the spherical mirror is least disturbed. The curvature of the field is not corrected as it is actually used to good advantage in cathode-ray tube projection.

The spherical aberration of the mirror may be interpreted as focusing by means of zones, each zone having a different focal length. The correcting lens has to be such that each zone of the lens has a different focal length, compensating for the various focal lengths of the mirror and resulting in a focusing system with all zones of the same focal length.

The shape of the correcting lens will thus depend upon the zonal focal length of the mirror one chooses as the focal length of the optical system (mirror plus correcting lens). Since theoretically there are an infinite number of zones on the mirror, there are theoretically an infinite number of correcting lens shapes that will produce a system in which all zones have the same focal length.

Since the mirror with an aperture at the center of curvature has no extra-axial or chromatic aberrations, such aberrations are caused by the correcting lens itself, i.e., by the power or slopes on the correcting lens. From the standpoint of these aberrations, therefore, that shape should be chosen whose maximum slope is the least. Thus if the paraxial (central) focal length of the mirror is chosen as that of the system, then the central focal length of the correcting lens is infinite and the shape of the curve is concave. Alternately, if a zonal focal length of the mirror is chosen as that of the system there will be a zonal focal length of the correcting lens which is infinite and the shape of the curve is convex at the center and concave past this zone. If a peripheral focal length is chosen, the required correcting lens is convex. The maximum slope is least for a convex-flat-concave curve.

The shape and size of the correcting lens depend upon the throw or magnification for which the system is to be used. For a given focal length and relative aperture, the correcting lens aperture decreases as the magnification decreases. That this must be so, may be surmized from the fact that for unity magnification the plate aperture is zero, since object and image coincide at the center of curvature.

Figure 4 shows the variation of correcting lens aperture and mirror aperture with magnification. Thus, a different correcting lens is required for each throw or magnification. If a high relative aperture astronomical Schmidt camera is used for projection at a throw only a few times the focal length, the resulting image is of poor quality. The reason is that a high relative aperture optical system can be well-corrected for only one position of object and image. The throw or magnification tolerance for a given correcting lens decreases with increased relative aperture for a given resolution.

To obtain a flat image field, i.e., focus on a flat viewing screen, it is necessary that the object field or tube face be curved. Calculations show that in general the shape of tube face depends on the throw—a sphere for infinite throw and an ellipsoid for finite throw. The eccentricity of the ellipsoid is sufficiently small, however, so that even for finite throw the tube face may be made spherical with a radius of curvature equal to that of the focal length of the system.

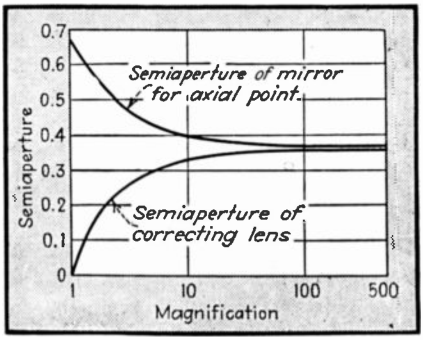

Fig. 4—Manner in which the semi-apertures of the mirror and correcting lens vary with magnification.

DESIGN OF CORRECTING LENS

The shape of the correcting lens must be such that all rays emanating from an object point O, and reflected by the mirror, shall meet at the image point I located at a distance S from the correcting lens. Figure 5 shows three rays emanating from O and striking the mirror at different apertures. Without the presence of the correcting lens, rays 1, 2, 3 would intersect the axis at distances q_1, q_2, and q_3 from the center of curvature. The slopes on the correcting lens have to be such (approximately as shown on Figure 5) that all three rays intersect at

I; hence, the correcting lens has a flat zone at the point where ray 2 passes, negative slope where ray 1 passes and positive slope where ray 3 passes.

Considered from the point of view of spherical aberration, if the zone where ray 2 strikes the mirror is taken as a reference, then the mirror has negative spherical aberration for smaller apertures and thus requires a positive lens for correction, and positive spherical aberration for larger apertures and thus requires a negative lens.

The shape of the curve of the correcting lens for any throw may be calculated to about the same accuracy as that obtained with the equation given by Hendrix and Christie[7] for infinite throw, from the formula

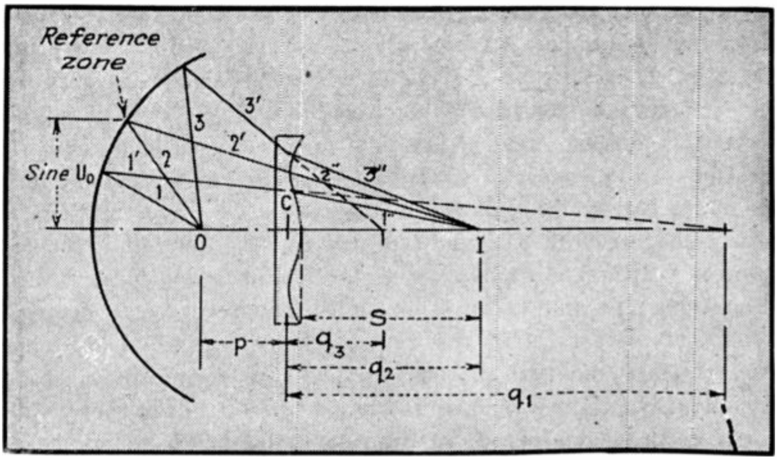

Fig. 5—Diagram illustrating how a suitably designed correcting lens makes the required corrections for spherical aberration.

$$d = \frac{1}{N-1}\left[\frac{1}{4}\left(\frac{1-p}{p}\right)^2 x^4 - \frac{1}{2}\left(2+\frac{1}{s}-\frac{1}{p}\right)x^2\right] \qquad (1)$$

or from its equivalent

$$d = \frac{1}{N-1}\left[\frac{1}{4}\left(\frac{m-S}{S}\right)^2 x^4 - \frac{1}{2}\left(\frac{2S-m+1}{S}\right)x^2\right] \qquad (2)$$

where d is the depth of the curve at the zone of radius x, p is the distance between object (tube face) and center of curvature of the mirror.

S is the distance between image (viewing screen) and center of curvature of the mirror, and m is the magnification. The relation between the quantities p, S, m and the focal length f of the system is given by

$$S = mp \tag{3}$$

$$p = f\ (m - 1)/m \tag{4}$$

$$S = f\ (m - 1) \tag{5}$$

All distances in the above equations are measured in terms of the radius of curvature of the mirror, i.e., the radius of curvature is taken as the unit of length.

In applications such as projection television, the light emitted by the luminescent screen first passes through a thickness of glass constituting the tube face. Although the effect of the tube face is small in cases of high f/number, it becomes quite appreciable for a low f/number system. The fact that the tube face is curved endows it with some power and actually alters the magnification of the system slightly. However, the largest effect of the tube face is caused by its spherical aberration. The presence of the tube face necessitates a change in the shape of the correcting lens. For a convex-concave correcting lens, the spherical aberration of the tube face calls for greater correction from the convex portion and smaller correction from the concave portion.

Equations (1) and (2) are not sufficiently accurate to determine the shape of the correcting lens for systems with high relative apertures. It was found that the best method of determining accurately the shape of the correcting lens is the old reliable and rather tedious, but very accurate method of tracing rays through the system consisting of the tube face, mirror and correcting lens.

PROJECTION EFFICIENCY

The projection efficiency of any optical system will be defined as the percentage of the total light flux, in lumens, emitted in a forward direction by an axial element of a perfectly diffusing source, such as a luminescent screen of a cathode-ray tube, which the optical system accepts and focuses on the corresponding image element, assuming 100 percent mirror reflection and 100 percent lens transmission.

The efficiency, e, in percent as defined above is given by: $e = 100 \sin^2 U$, where U is the semi-apex angle shown on Figure 6. Hence, to determine the efficiency of a lens for a perfectly diffusing source, it is merely necessary to know the angle that the lens, or entrance pupil, subtends at the source. As may be seen from Figure 6,

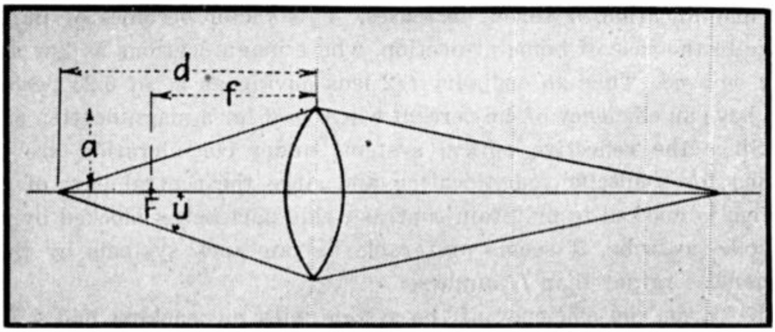

Fig. 6—Efficiency of a simple lens decreases as the lens is moved away from its source to decrease the magnification.

the farther a given lens is from a source, i.e., the small the magnification, the lower is the efficiency of the lens.

It has become customary to rate a lens by its *f*/number for infinite magnification, i.e., object located at the focal point of the lens. The *f*/number is defined as

$$f/\text{number} = 0.5 \sin U = 0.5 \sqrt{e_\infty}$$

where e_∞ is the efficiency (a fraction, not percent) for infinite magnification. The smallest *f*/number possible is 0.5, since at 0.5 the efficiency is unity and all the light emitted by the object element in a forward direction is concentrated in the image element. Figure 7 shows the efficiency e_∞ of a lens as a function of *f*/number. It is seen that the efficiency of most lenses is very low.

As already mentioned, the efficiency of a given lens decreases when

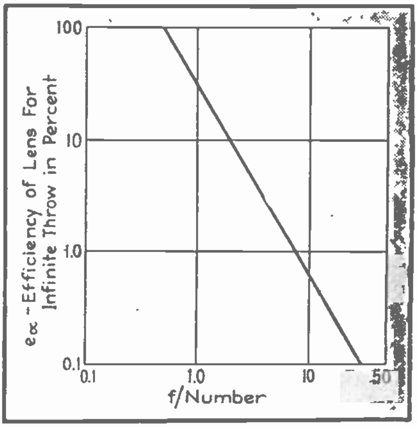

Fig. 7—Variation of lens efficiency with its *f*/number

the magnification or throw decreases. This factor becomes of importance in the case of home projection, where magnifications as low as 5 may be used. Thus an ordinary $f/2$ lens having an e_∞ of 6.25 percent will have an efficiency of 4.6 percent when used for a magnification of 6.

Since the reflective optical systems under consideration are designed for a specific magnification and since the central part of the system is masked to maintain contrast, this part being blocked by the cathode-ray tube, it seems preferable to rate such systems by their efficiencies rather than $f/$number.

Let e_0 be the efficiency of the system with no masking and e_1 the efficiency of the central part of the system that is masked. The efficiency e of the masked system is then simply

$$e = (e_0 - e_1)\ 100\%.$$

Here e_0 and e_1 (fractions, not percent) may be calculated approximately from the equations

$$e_0 = \frac{h_c^2}{p^2} = \frac{h_c^2}{f^2}\,\frac{m^2}{(m-1)^2}$$

$$e_1 = \frac{h_t^2}{f^2}\left(\frac{m^2}{m^2-1}\right)^2$$

where h_c is the semiaperture of the correcting lens and h_t is half the diameter of the tube face. For high-efficiency systems e_0 will be above 40 percent and e_1 approximately 10 percent so e, the efficiency of the system with blocking, will be about 30 percent. Neglecting losses in the system, about 30 percent of the light emitted by an axial point will be focused into an image point. This corresponds to the efficiency of an $f/0.8$ lens used at a magnification of 6.

ALIGNMENT REQUIREMENTS

The center of the correcting lens must be located at the center of curvature of the mirror and, for uniform illumination over the field, the axis of symmetry of the correcting lens should preferably coincide with the axis of symmetry of the periphery (circle) of the mirror. The tube face must be located so that the center of curvature of the tube face lies on the axis of symmetry of the correcting lens. For uniform illumination over the field, the axes of symmetry of periphery of tube face and correcting lens should preferably coincide.

The tube face should, of course, be located at the correct axial distance from mirror or correcting lens for focusing. The viewing screen should be normal to the axis and at the correct throw.

The most critical alignment items are: (1) Lateral displacement of the center of the correcting lens from an axis of symmetry of the mirror, i.e., a line passing through the center of curvature of the mirror; (2) Lateral displacement of the center of curvature of the tube from the axis of symmetry of the system. For good resolution these displacements should be kept within 0.001 R, where R is the radius of curvature of the mirror. The permissible tolerances on other alignments are about 10 times greater.

There are two distinct applications for projection television, namely, in theater television equipment and in television receivers for home use.

PROJECTION RECEIVERS FOR HOME

In a self-contained projection television receiver[8] the optical system can be mounted near the floor with its axis vertical, projecting the image straight up and onto a flat mirror inclined at 45 degrees to the incoming beam of light, and throwing the image on a translucent screen. Such an arrangement presents the advantages of compactness, relatively small depth of the cabinet and can be styled along the familiar lines of a radio console.

A number of such reflective projection systems suitable for home receivers of the type described have been designed, built and operated in actual receivers. The smallest of these was built for use with a cathode-ray tube having face diameter of 3 inches, and consists of a spherical mirror 9 inches in diameter and a correcting lens 6 inches in diameter. The largest has tube, mirror and lens diameters of 5, 14 and 9.5 inches respectively. A number of systems in sizes intermediate between the two just described have been built. The throw or distance between the correcting lens and the viewing screen varies between 36 and 54 inches and the optical efficiencies are between 18 and 35 percent. In resolution and contrast these systems compare favorably with well-corrected conventional projection lenses, and do not limit the performance of present television systems.

SYSTEMS FOR THEATERS AND CAMERAS

A description of the RCA theater television system was published several years ago.[3] The optical system consists of a 30-inch mirror, 22.5-inch correcting lens and operates with a cathode-ray tube 7.5

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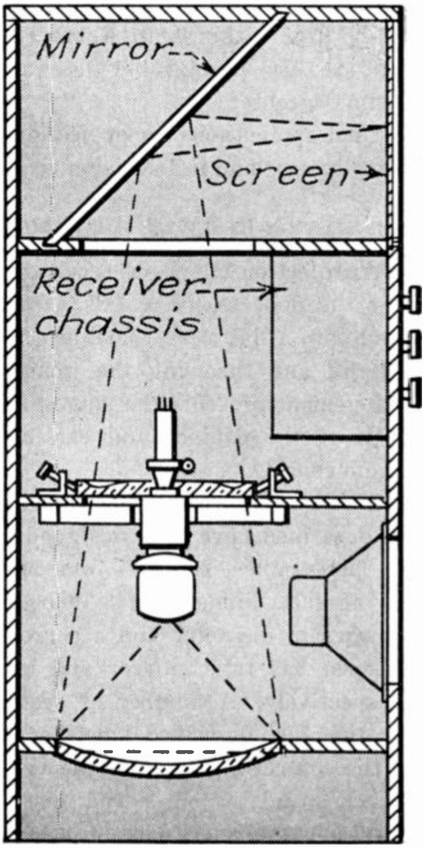
(8) Landis, D. O., U. S. Patent 2,273,801, filed Dec. 30, 1938.

inches in diameter. Figure 8 shows the optical system with the cathode-ray tube in place. The control console may be seen in the background.

Reflective optical systems built for infinite throw find useful application in television pickup cameras under conditions of low illumination, such as during the last minutes of a football game or in direct pickup from a theater stage. The great light-gathering power of these optical systems is demonstrated in Figure 9. An optical system with infinite throw was pointed from a window in Camden, N. J., toward the Philadelphia skyline. The bright image of the skyline can be seen inverted on the dummy tube face, undestroyed by the full daylight illumination.



Arrangement of optical system for a home television receiver employing reflective optical principles. This design gives a large-screen picture with a console cabinet no deeper than that of an ordinary home radio receiver.

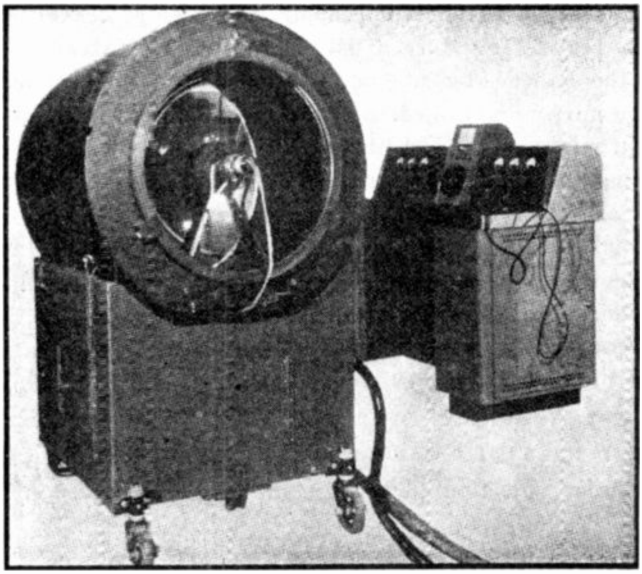


Fig. 8—RCA theater television projector, with control corsole in background.

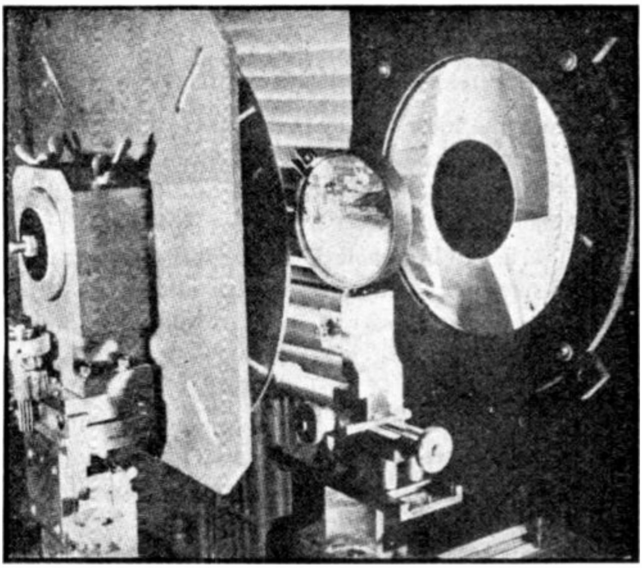


Fig. 9—Image of Philadelphia skyline as formed on the face of a dummy tube by a reflective optical system.

An interesting modification,[9] applicable to all systems described, is shown in Figure 10. Here a flat mirror is inserted about half-way between the cathode-ray tube and the spherical mirror. Since the center of the mirror is blacked out to increase contrast, the opaque back of the flat mirror cuts very little of the useful light coming from the tube facing the spherical mirror, but the flat mirror permits placing another cathode-ray tube back of the spherical mirror and facing the flat mirror. Such an arrangement may be used in theater work since both tubes can operate singly with roughly the same optical efficiency. If one tube goes bad the other may be turned on by a flip of the switch. With some technical difficulty both tubes may be operated at the same time, the problem arising in the exact super-position of two scanning patterns.

Cost Factors in Reflective Optics

The major objection to the use of reflective optics in television receivers has been the high cost of the aspherical correcting lens. The spherical mirror, while quite large, is an old and familiar item to the well-established optical industry, as most of the conventional optical surfaces are spherical and are easily made. The aspherical correcting lens, similar to a figure of revolution developed by rotating a shallow letter S around one of its ends, presents an altogether different problem. Unlike the spherical mirror, such a figure is not a naturally-generated surface and there are no machines on the market for straightforward production of such surfaces. True enough, astronomers, with their traditional patience and lack of hurry, produced excellent aspherical lenses on machines used for making astronomical instruments, but only by tedious step-by-step methods.

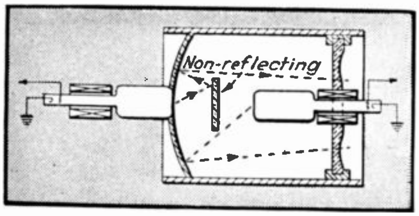


Fig. 10—Bi-reflective optical system, employing two projection television cathode-ray tubes.

In the early stages of the development, RCA used methods and machines based upon astronomical technique. Exceedingly high cost of experimental reflective optics resulted. The gain in light over the

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(9) Epstein, D. W., U. S. Patent 2,295,779, filed Aug. 17, 1940.

conventional projection lens was very attractive, but the cost of such individually produced lenses was prohibitive. The apparent solution to the cost problem was that of molding the aspherical lenses from a suitable transparent material.

PLASTIC CORRECTING LENSES

A special development project was undertaken and soon concentrated on investigation of a clear thermoplastic material known under the name of methyl methacrylate, and sold under the registered trade names of Lucite and Plexiglas.

A new set of difficult problems came to the foreground. The most formidable of these was that of making molding surfaces of metal in shapes of the negative replicas of aspherical lenses. Almost as serious was the problem of obtaining optical finishes on metals. Both of these problems have been successfully solved.

A flat disk of hardenable stainless steel is first profiled with the aid of a template. The template itself is filed according to a theoretically calculated curve. The profiling machine has a five-to-one lever action which calls for a template five times deeper than the final curve.

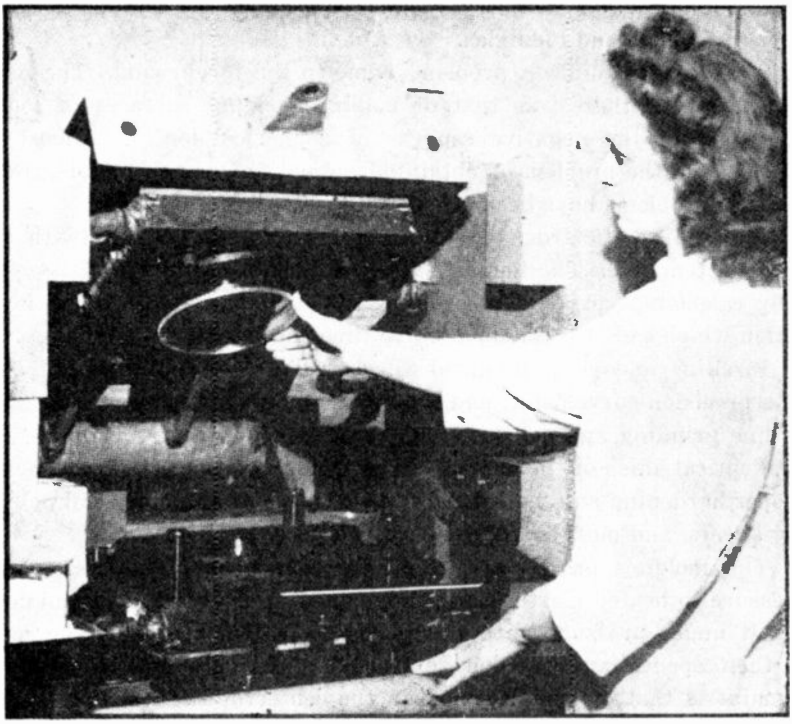
Profiling is done by diamond wheels. The resultant curve is tested on a precision curvemeter, and final adjustments of the curve are done by fine grinding and polishing on a precision polishing machine. The final optical finish of the surface is the result of proper choice of metal, proper hardening and tempering, proper choice of abrasives and polishing agents, and most of all, patience and perseverance.

The molding process is essentially that of applying very high pressure to heated plastic material confined in a heated mold and cooling it under pressure until it reaches room temperature. The mold is then opened and the lens extracted. The only operation which remains is that of boring a hole in the center of the lens for accommodating the protruding neck of the cathode-ray tube. The lens is then ready for use, with no polishing or finishing of any sort required.

Molded correcting lenses for reflective optical systems possess very good optical properties, including slightly better transmission and slightly lesser scattering of light than glass. They do not possess the surface hardness and scratch resistance of glass, but even without any special care or protection they have stood up under laboratory operation for more than three years. The cold flow under operating conditions of three years was found to be negligible. The cold flow depends on the operating temperature, which for the plastic lens of a television receiver is not far from room temperature. Should design considerations call for higher operating temperatures, the new boilable methyl methacrylates can be used.

MOUNTING PROBLEMS

From a practical standpoint, the use of reflective optics in television receivers calls for careful consideration in the mechanical construction of the mounting which supports the optical system and the cathode-ray tube. This mounting, combining "the barrel" and "tube support," has to fulfill a number of requirements: (1) Since the position of the correcting lens with respect to the mirror is rather critical, the mount must provide for positive and simple alignment at the factory; (2) It must be dustproof, since accumulation of dust on the mirror and correcting lens reduces both the contrast and the illumination, while frequent dusting would be detrimental to the plastic lens and the front surface mirror; (3) It must be electrically shock-proof since in some cases final optical focusing of the picture on the viewing



Molding press used for producing plastic correcting lenses for projection television systems.

screen must be done with a picture and consequently with high voltage on the cathode-ray tube; (4) The barrel should preferably be made of metal, to cut off x-rays generated by the cathode-ray tube. These rays are very soft and weak; nevertheless, they are measurable and should be screened in; (5) It must provide for positive and convenient initial adjustments of the tube face position along three rectangular coordinates, one of which coincides with the optical axis of the system. These initial adjustments may be carried out by the factory and by experienced servicemen; (6) It must provide for easy tube replacement by people unfamiliar with optics, such as the average serviceman and the customer himself; (7) It must provide for easy and safe focusing after tube replacement; (8) It must be designed to lend itself to such inexpensive manufacturing processes as stamping or die casting, involving a minimum of machining; (9) It must not deform in transportation and during years of service.

TYPICAL MOUNTING

A layout of a mounting satisfying the requirements discussed is shown in Figure 11. Here the correcting lens fits into a recess on the top of a metal barrel, this recess being counterbored for a snug fit with the correcting lens. The spherical mirror is mounted on the bottom cover of the barrel by means of a collar and nut through the center hole of the mirror.

The tube support consists of an arm of insulating material anchored on the side of the barrel and a metal ring supporting the face edges of the cathode-ray tube. The tube face is held tight against this ring by suitable springs. The high voltage is brought to the second anode of the tube through a dust-tight hole in the wall of the barrel. The metal ring holding the tube is at high potential, and several inches of Micalex insulate it from ground. The high-voltage cable has a grounded shield on the outside and the barrel itself is grounded.

The tube support arm is arranged to slide back and forth, providing for tube adjustment in a direction perpendicular to the optical axis of the system, say, along a rectangular coordinate x. The support of the arm is arranged to slide along an intermediate supporting plate in a direction y, perpendicular to both the x coordinate and the axis of the optical system. The intermediate supporting plate is made to slide up and down the barrel by means of a screw, providing a focusing means along the axis of the optical system or coordinate z.

The deflecting yoke is supported by the neck of the cathode-ray tube and is equipped with dust-proof gaskets. The top of the barrel may be equipped with a cardboard shield reaching to the upper part of the television cabinet and preventing dust from settling on the upper side of the correcting lens.

The arrangement described satisfies the requirements enumerated more or less completely and allows for variations governed by the individual preference of the designer.

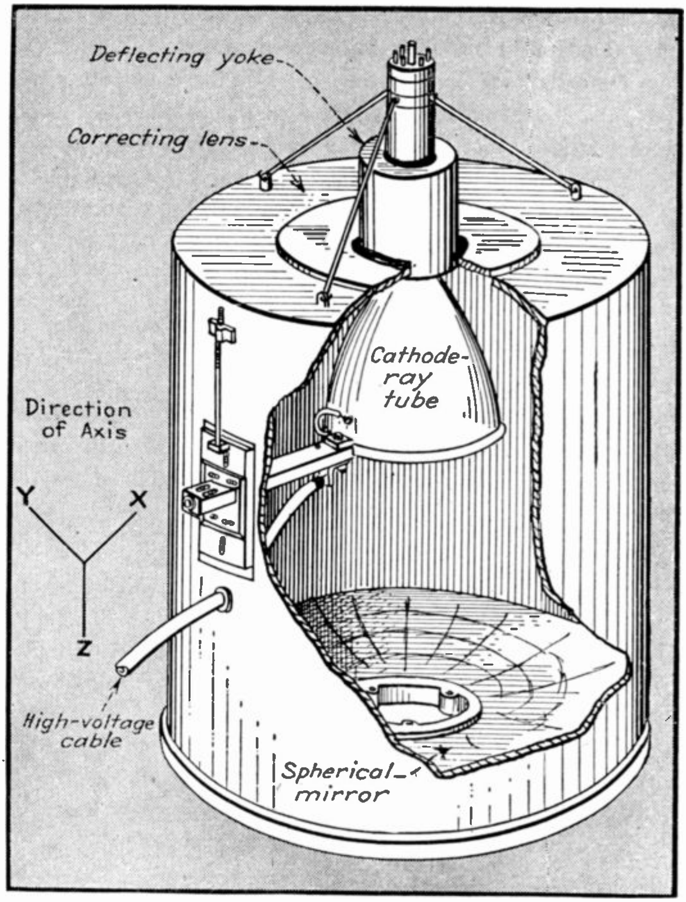


Fig. 11—Method of mounting optical components of a projection television system to give rigidity while permitting adjustments when required.

APPARENT DETAIL

If one wants to place an enlargement of a given picture on the wall of a room of a given size, he can find by experiment a size of enlargement that will give an "optimum effect." This size will give a picture that is not unduly blurred and does not require squinting to see the detail. In television with its intrinsic or absolute detail governed by the bandwidth of the channel of the transmitter or the wire channel, the subject of optimum size for a given application is of major importance.

The amount of apparent detail needed for a pleasing television picture will determine how much magnification the picture will stand in any particular application. For a given amount of absolute detail the picture size will be larger for hotel lobby applications than it will be for home use, still larger for auditorium use and much larger for theater use. The exact sizes may vary somewhat but it is believed that the buying public will soon find out what value of apparent detail is the most acceptable for a given use. Consequently, the apparent detail will determine the size of the projected television picture to be preferred for each application.

CATHODE-COUPLED WIDE-BAND AMPLIFIERS*†

BY

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Summary—A general analysis indicates that, in wide-band amplifiers, stable operation is possible with triodes in circuits using the cathode as a signal terminal. The amplification, however, is approximately equal only to the square root of that available with grounded-cathode amplifier, and therefore twice as many tube units are required to obtain the same amplification. In certain applications, however, the utility of such circuits outweighs the loss of gain.

A simple radio-frequency amplifier was designed for television receivers, using a cathode-input circuit. By combining a cathode-output and a cathode-input stage using one single twin-triode tube, a circuit was devised which compares favorably with pentode stages with respect to gain, stability, and economy, while it has far superior noise characteristics. The new circuit, called the "cathode-coupled twin-triode" amplifier, provides greater flexibility than conventional amplifier circuits, and can be used for radio-frequency, intermediate-frequency, video, converter, or detector services. Since the same tube type can also be used for synchronizing and deflection circuits, the number of tube types can be materially reduced, and greater standardization with further economical advantages may be obtained. An interesting application of the new circuit is a novel bidirectional amplifier.

I. INTRODUCTION

FOR approximately two decades, screen-grid tubes were used almost exclusively for amplification of high-frequency signals. The screen grid acting as a shield reduced the effect of the output circuit on the signal circuit, and provided a high-impedance output. When, with the advent of the video art, in the case of extremely wide-band amplifiers, the external circuits had lower impedances, the advantage of the high plate impedance became less significant. In the particular case of the cathode-output (cathode-follower) circuit, for instance, multigrid tubes were used as triodes purely for the reason that they had higher transconductance than the commercially available triodes.

As the operating frequencies of radio communication increased, the transit-time effect became more and more significant. In order to reduce the effect, the spacing of the tube electrodes was reduced, and it became increasingly difficult to align several grids in extremely close

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* Decimal classification: R583 x R363.4
† Reprinted from *Proc. I.R.E.*, October, 1945.

proximity. Thus, in lieu of the screen grid, the grid was used as a shield between the input and output, and in certain cases the cathode-input (grounded-grid or inverted) amplifiers provided superior results in performance and economy. As engineering knowledge about noise sources expanded, the multigrid tubes were avoided in stages where the signal was small.

It is the purpose of this report to give a comparative analysis of vacuum-tube circuits using multigrid and triode tubes in wide-band circuits. In the case of the triodes, circuits using the cathode as a signal electrode are emphasized, and a new cathode-coupled circuit is introduced. This circuit surpasses the advantages of pentode circuits with respect to economy and stability, and possibly permits a broader tube-standardization program.

II. DEFINITIONS

As it appeared above, the nomenclature applied to the various amplifier circuits is not too well standardized. As compared with the conventional amplifier, in which the cathode is substantially grounded with respect to high-frequency current, two other configurations are possible when the cathode is not grounded. In the first, the cathode serves as output terminal, and is called the cathode follower, or grounded-plate amplifier. The second uses the cathode as the input terminal, and is called the inverted, or grounded-grid amplifier. In the present paper we propose to regard the cathode as the reference point, since it is the primary electrode of a vacuum tube (the source of electrons), and we shall use the terms of grounded-cathode (Figure 1), cathode-output (Figure 2), and cathode-input circuits (Figure 3). For the circuit shown in Figure 4, we adapted the term of "cathode-coupled twin-triode" stage. All circuits in which the cathode is not at ground, but serves as an input or output terminal, will be designated as cathode-coupled circuits against the conventional grounded-cathode amplifier.

III. WIDE-BAND GROUNDED-CATHODE AMPLIFIERS

The basic circuit, and its equivalent network, are shown in Figure 1. This familiar circuit is designed such that, at frequencies of $f_0 \pm \Delta f/2$, the amplification is 0.707 times that of the amplification at f_0, where f_0 is the resonant frequency and Δf is the bandwidth. If this stage is preceded by a similar stage, the amplification is then $A_{gc} = g_m/(\Delta\omega\sqrt{C_1C_0})$ (see Appendix I (7)) where g_m is the transconductance, $\Delta\omega = 2\pi\Delta f$, C_1 is the input, and C_0 is the output capaci-

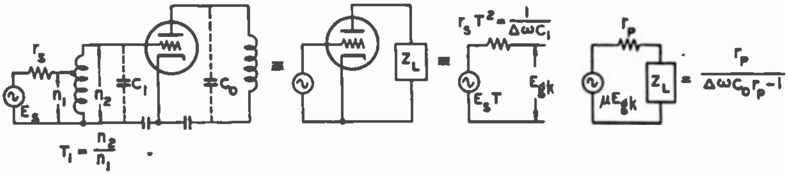

Fig. 1—Grounded-cathode amplifier circuit and equivalent network.

tance. The grid-to-plate capacitance and other sources of feedback are assumed to be negligible. Since the last assumption is generally untrue, in order to reduce the grid-to-plate capacitance a screen is placed between the grid and the plate.[1] In this and the following equations, the value of $\Delta\omega = 2\pi\Delta f$ may be taken around any center frequency, and, accordingly, they are equally valid for video, intermediate-frequency, or radio-frequency amplifiers (see Appendix II). The formula given above is for a simple tuned circuit, as shown in Figure 1. With a coupling circuit of more complex nature, greater gains may be obtained, as was shown by Wheeler.[2] For purposes of simple comparison, only the simple coupling circuit is considered here, but the same factors of improvement apply in all the cases when more complex coupling circuits are used.

IV. WIDE-BAND CATHODE-OUTPUT AMPLIFIERS

The basic circuit and its equivalent network are shown in Figure 2. This circuit is shown in a form to work into a high-impedance circuit, such as the input of another similar stage. A circuit of this type was proposed in 1925 in order to reduce feedback in radio-frequency amplifiers.[3] A more important application of this circuit became popular in recent years when it was applied to output loads of low impedance, such as transmission lines.[4] This latter type of operation of this cir-

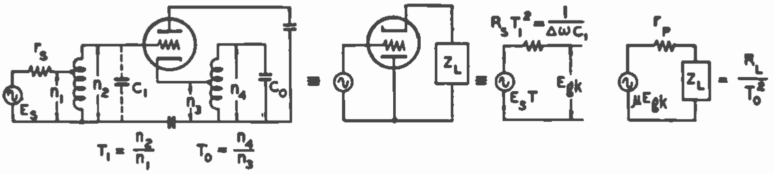

Fig. 2—Cathode-output amplifier and equivalent network.

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[1] W. Shottky, United States Patent No. 1,537,708.

[2] H. A. Wheeler, "Wide-band Amplifiers for Television," *Proc. I.R.E.*, vol. 27, pp. 429-438; July, 1939.

[3] A. Winther, United States Patent No. 1,700,393.

[4] A. D. Blumlein, United States Patent No. 2,178,985.

cuit has been frequently analyzed in the literature.[5,6] It was shown that the input capacitance of the stage is reduced by a factor of (1 — amplification from grid to cathode) providing greater permissible impedance for the previous circuit. In general, the circuit behaves as if the tube had an amplification factor and plate resistance divided by $(\mu + 1)$. For our particular case the amplification is $A_{co} = \sqrt{g_m/(\Delta\omega C_1)}$ (see Appendix I (16)). This is (provided the stage is preceded by a similar stage) the square root of the amplification obtainable from a pentode with the same g_m and capacitances, thus indicating that two cathode-output stages in cascade are required to provide gain in the same order as that of one pentode stage.

V. WIDE-BAND CATHODE-INPUT AMPLIFIERS

The basic circuit and the equivalent network are shown in Figure 3. In this circuit, the input and output circuits are shielded by the grid. This method of shielding although disclosed[7] as early as 1927, became popular only during the last few years.[8-10] If the stage operates from a source of matching impedance, as is the case when the source is predominantly resistive, the amplification is $A_{ci} = (1/2)\sqrt{\mu/[\Delta\omega C_0 r_s(\Delta\omega C_0 r_p - 1)]}$ (see Appendix I (23)). It may be observed again that the amplification is less than the square root of the amplification of the grounded-cathode triode amplifier, thus requiring twice as many stages for the same over-all gain. When the signal sources is predominantly reactive, the expression changes to

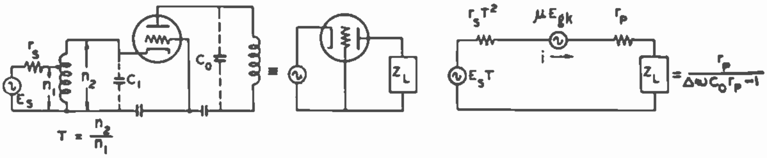

Fig. 3—Cathode-input amplifier and equivalent network.

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[5] A. Preisman, "Some Notes on Video Amplifier Design," *RCA REVIEW*, vol. 2, pp. 430-432; April, 1938.

[6] A. A. Barco, "An Iconoscope Preamplifier," *RCA REVIEW*, Vol. IV, pp. 102-107; July, 1939.

[7] E. F. Alexanderson, United States Patent No. 1,896,534.

[8] C. E. Strong, "The Inverted Amplifier," *Electronics*, vol. 13, pp. 14-56; July, 1940.

[9] M. Dishal, "Theoretical Gain and Signal-to-Noise Ratio of the Grounded-Grid Amplifier at Ultra-High Frequencies," *Proc. I.R.E.*, vol. 32, pp. 276-284; May, 1944.

[10] M. C. Jones, "Grounded-Grid Radio-Frequency Voltage Amplifiers," *Proc. I.R.E.*, vol. 32, pp. 423-429; July, 1944.

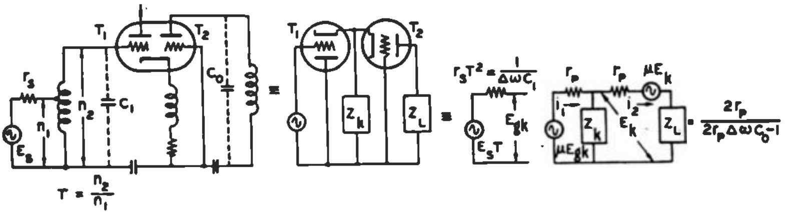

Fig. 4—Cathode-coupled twin amplifier and equivalent network.

$A_{ci} = \sqrt{g_m/(\Delta\omega C_0)}$ (see Appendix I (26)).

While this amplifier has lower gain than the grounded-cathode amplifier, it finds its greatest utility as a radio-frequency stage between the antenna and the converter stage because of the fact that it is stable without the need of a screen grid or neutralization. This type of amplifier generates considerably lower noise currents than a pentode would in the same service. It provides great improvement, for instance, in receiving television signals. Since the impedance appearing across the tube input for high g_m is low, $Z_1 = [r_p + (1/\Delta\omega C_0)]/(1+\mu)$ the tuned circuit provides an adequately flat response over the whole television band, and therefore no tuning means is required for the antenna circuit for a six-channel receiver.

The circuit diagram of a simple cathode-input radio-frequency amplifier to be used with television receivers is shown in Figure 5. Figure 6 shows such an amplifier mounted in an RCA TRK-120 television receiver. Figure 7 shows the inside of the auxiliary chassis. This amplifier affords an additional amplification of 2 to 4, and a significant improvement of the signal-to-noise ratio. The heterodyne oscillator signal is substantially reduced in the antenna, thereby reducing radio-frequency interference between two receivers. Several of these simple amplifiers were made and attached to receivers in the Princeton area (approximately 45 miles from New York and Philadelphia), and considerable improvements were obtained in every case.

Since the antenna circuit feeding into the cathode is untuned, some thought has been given to the question of cross modulation in the cathode-input radio-frequency amplifier due to two strong carrier

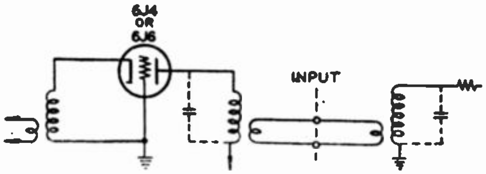

Fig. 5—Television radio-frequency amplifier circuit.

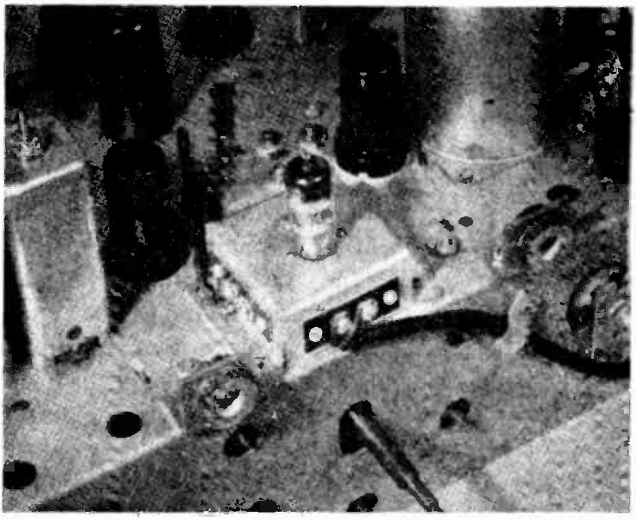

Fig. 6—Radio-frequency amplifier on RCA television-receiver chassis.

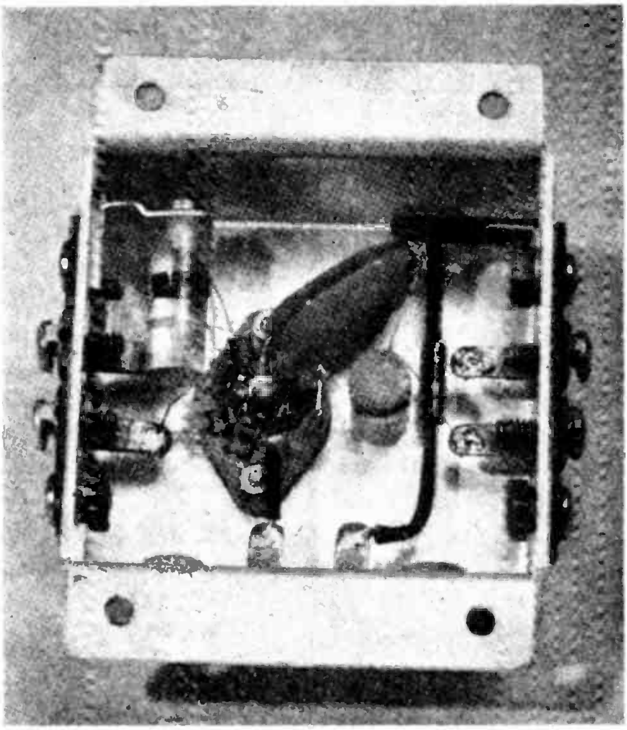

Fig. 7—Internal view of radio-frequency amplifier.

signals. Cross modulation is a function of the strength of the signals and the degree of curvature of the tube characteristic. In this amplifier, the magnitude of signal voltages appearing between grid and cathode is less than those in the antenna, since a 1:1 transformer is used for coupling, and the circuit is highly degenerative. These voltages are less than those appearing at the grid of a converter tube in television sets using a step-up transformer for coupling the antenna to the grid. The amplifier characteristics of a cathode-input amplifier are less curved because of the high degeneration of the cathode circuit. One is therefore led to the preliminary conclusion that cross modulation is less serious in the cathode-input amplifier than in the converter even though the grid of the latter is tuned. The tuned circuit in the converter is too broad to give sufficient adjacent-channel rejection.

In some cases the input loading is far in excess of that required to obtain the desired bandwidth. In such cases a compromise between the grounded-cathode and cathode-input amplifier may be obtained by moving the ground from the grid toward the cathode on the secondary of the input transformer, shown in Figure 8. The limiting factor in this case will be found in the stability since the grid is a less effective shield, but with proper tapping of the coil, stable operation may be obtained. The circuit will behave as if the amplification factor of the tube were increased by the transformer ratio of the cathode part of the coil to the total secondary.

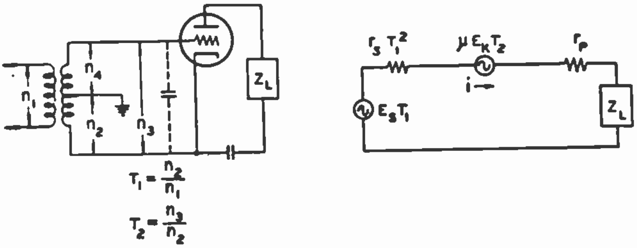

Fig. 8—Tapped cathode-input amplifier and equivalent network.

These circuits are particularly suitable for antenna-plex systems as a consequence of the inherently good noise and wide-band characteristics.

VI. WIDE-BAND CATHODE-COUPLED AMPLIFIERS

As has been shown, the cathode-output circuit provides a comparatively high input-impedance circuit, with the additional advantage that this impedance is not changed materially by external potentials, such as grid bias, plate voltage, etc. It has been proposed to use such a stage in conjunction with a grounded-cathode amplifier,[11] but such a circuit, besides requiring the same number of circuit elements as two stages, uses a pentode tube as the grounded-cathode amplifier. In some cases this dual stage does not provide adequate stability, and also does not provide better noise characteristic than a pentode. By connecting a cathode-output and cathode-input stage together, as shown in Figure 4, we obtain a high-gain wide-band amplifier stage.

The amplification of a wide-band amplifier of this type is $A_{CCTT} = g_m/(2\Delta\omega\sqrt{C_1C_0})$ (See Appendix I (41)) which is favorably comparable to the gain obtained for grounded-cathode amplifiers, particularly since the input capacitance (C_1) is reduced by a factor of $1-(1+2\Delta\omega C_0 r_p)/(4\Delta\omega C_0 r_p)$ (see Appendix I (46)).

The circuit is economical since a coil and a resistor (the coil preferably wound on the resistor) are the only coupling elements required between the two tube units. The resistor and by-pass capacitor customarily required in a screen supply are eliminated. Since the plate currents in the two triode units swing in opposite directions, subsequent similar stages have little influence on each other, due to varying load on the plate supply. By examining the circuit we may notice that the input and output signals are of the same polarity. Hence, when a cathode-coupled amplifier is used for video amplification, no attention need be given to the number of stages in order to obtain the proper polarity.

A twin-triode tube with a common cathode may be manufactured more economically than a pentode of the same transconductance. While this point may be debatable at present, it can be shown that receivers could be designed in which twin triodes were used in nearly all stages, and by reducing the tube types, the cost of the preferred-type tube could be reduced still further. Figure 9 is a block diagram of a 16-tube television receiver in which 12 tubes are of the twin-triode type.

Table I shows the amplification obtainable from conventional high g_m pentodes in grounded-cathode circuits and from twin triodes in coupled-cathode wide-band amplifier circuits. To allow for the capacitances of tube sockets, wiring, etc., 2 micromicrofarads was added to the tube capacitances of each terminal, given in the tube handbook, for miniature tubes. Similarly, for octal metal tubes, 5 micromicrofarads was added. The bandwidth was assumed to be 4 megacycles, and the gain formulas given above were used. The input capacitance

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[11] P. Selgin, "The Cathode Driver as an R-F Coupling Stage," *Radio*, vol. 28, pp. 26-28; March, 1944.

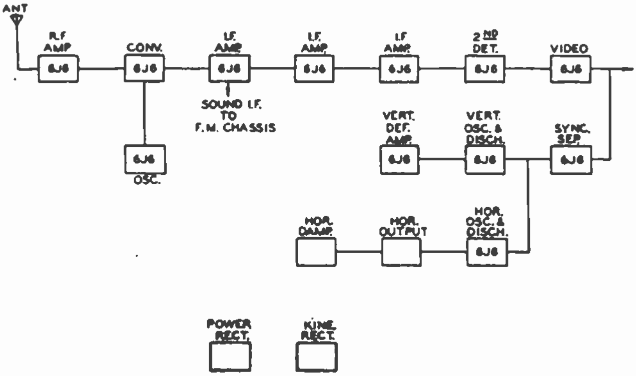

Fig. 9—Block diagram of television receiver using 6J6 tubes.

of the coupled stages was corrected for degeneration. The tube-cost figures were also taken from the RCA Tube Handbook.

VII. SPECIAL CATHODE-COUPLED STAGES

On examination of Figure 4, we may observe that the two grounded electrodes correspond to the input and output electrodes in the reverse direction. By injecting a signal of different frequency through a resonant circuit that appears substantially as a short circuit for the signal applied in the original direction to the grid of the second tube T_2, and taking it off the plate of the first tube T_1 in the same manner, we obtain a bidirectional amplifier as shown in Figure 10.

The terminals of the two signals are completely independent, and the capacitance of each will determine the bandwidth and gain of its own signals. An amplifier of this type may be useful for bidirectional relay stations, reflex circuits, etc. Approximate calculations and experimental results indicate that with simple resonant circuits the two signals must be approximately twice their bandwidths apart. With a

TABLE I

COMPARATIVE WIDE-BAND AMPLIFIER DATA

Tube Types		Circuit	Bandwidth 4 megacycles				Equivalent Root-Mean-Square Grid Noise μV	List Price $
No.	Base		G_m μV	C_{gk} $\mu\mu$f	C_o $\mu\mu$f	Amplification		
6AC7	Octal	Grounded-Cathode	9000	11.0	5.0	14.2	6.8	1.75
6AB7	Octal	Grounded-Cathode	5000	8.0	5.0	8.8	12.6	1.15
6AG5	Miniature	Grounded-Cathode	5000	6.5	1.8	17.6	10.4	2.15
6J6	Miniature	Cathode-Coupled	5300	2.2	1.6	10.4	7.5	1.85

smaller frequency separation, the electrodes, which are supposed to be grounded, do not provide constant potentials, and, due to the regeneration, both pass bands are reduced. Further work on more complex circuits may permit the choice of closer signal frequencies.

The experimental chassis containing a bidirectional cathode-coupled stage is shown in Figure 11. The signals applied were the frequency bands 8.5 to 13 megacycles and 24 to 28.5 megacycles. A gain of approximately 12 was obtained in both directions with a 6J6 tube. Figure 12 shows a simple intermediate-frequency transformer construction for the frequency band 8.5 to 13 megacycles with a 6J6 tube. The advantage of the bidirectional amplifier could be summed up by claiming a total amplification equal to the square of that of the unidirectional stage, or by claiming twice the bandwidth with the same gain.

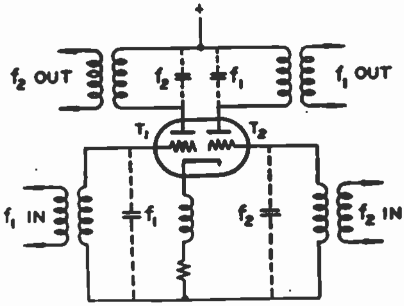

Fig. 10—Bidirectional-amplifier circuit.

The cathode-coupled stage can be used also as a frequency converter as shown in Figure 13. The grid of T_2 is substantially grounded for all frequencies except for the frequency of the tank circuit of the local oscillator. The local oscillator varies the transconductance of the tube, and therefore provides an intermediate-frequency output across the tuned circuit connected to the plate. The second tube T_2 acts as a cathode-output stage for the oscillator signal, and attenuates it by 6 decibels toward the antenna since it works into an impedance like its own. The first tube T_1 further attenuates this signal by providing a divider through its grid-cathode capacitance and the input impedance.

A simple cathode-coupled two-terminal oscillator circuit[12] is shown in Figure 14. This is merely the twin-triode cathode-coupled amplifier described above, in which the output plate is coupled back to the input grid through some coupling impedance. The grid of the cathode-input

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[12] M. G. Crosby, United States Patent No. 2,269,417.

section T_2 is normally returned to ground. However, by properly biasing this grid, it is possible to obtain a frequency variation in excess of plus or minus 75 kilocycles about a mean frequency of 50

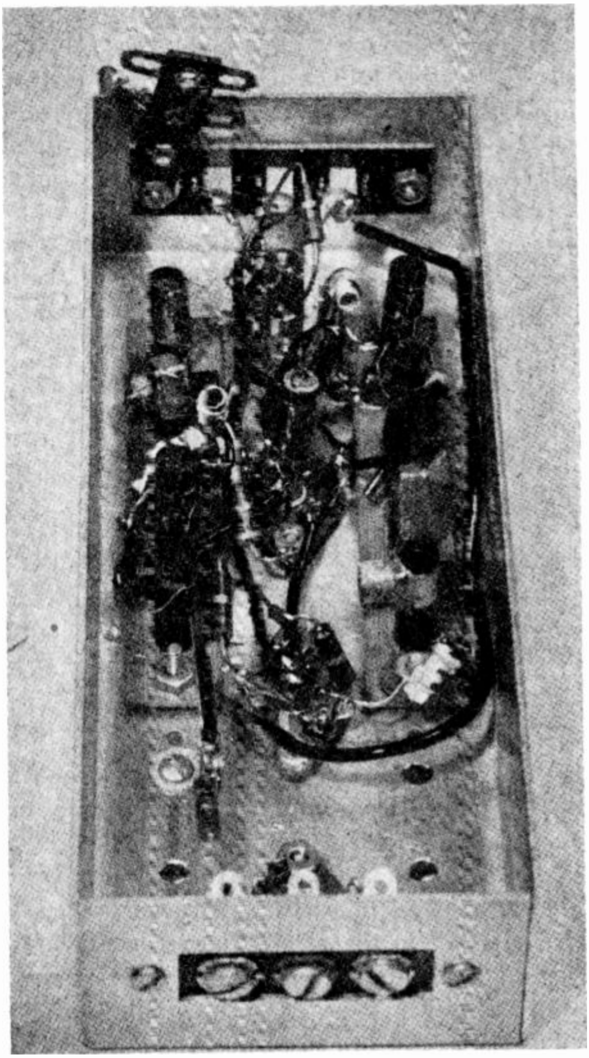

Fig. 11—Experimental bidirectional-amplifier stage.

megacycles with a 6J6 tube, with a bias variation of plus or minus one volt. In a television or frequency-modulation receiver this feature can be used to good advantage to provide vernier tuning or automatic frequency control without adding a reactance tube. In a frequency-modu-

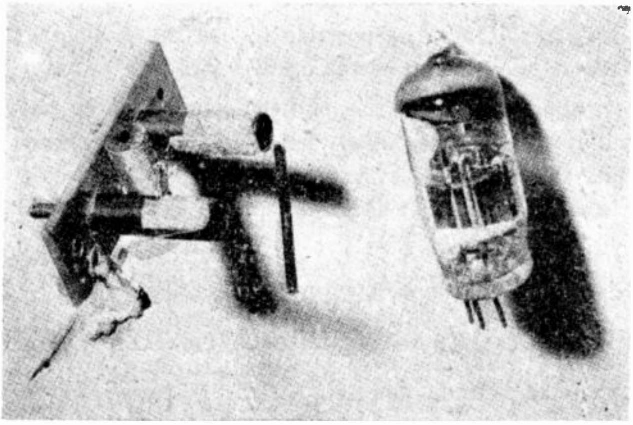

Fig. 12—Miniature intermediate-frequency transformer and tube.

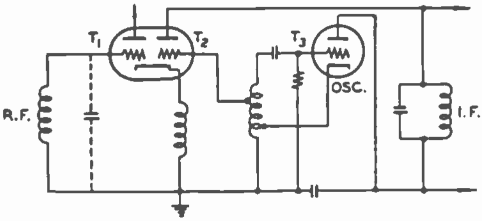

Fig. 13—Cathode-coupled frequency converter.

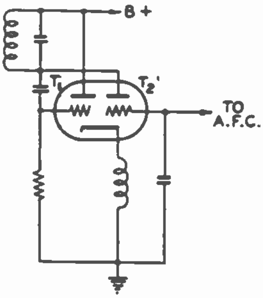

Fig. 14—Two-terminal oscillator.

lation transmitter it may be possible to use this property to obtain direct frequency modulation of the oscillator.

Further applications of the cathode-coupled stage may include lock-in oscillators, reactance tubes, self-oscillating converters, etc. The economy and the standardization possibilities of the circuit may well suit it for a large number of different applications.

APPENDIX I

DERIVATION OF GAIN FORMULAS FOR WIDE-BAND AMPLIFIERS

The grounded-cathode amplifier, with its equivalent network, is shown in Figure 1. If we desire to maintain an amplification at a frequency $f = [\omega_0 \pm 1/(2\Delta\omega)]/2\pi$ that is approximately equal to 71 per cent of the amplification at resonance (see Appendix II)

$$r_s T^2 = 1/(\Delta\omega C_1) \tag{1}$$

provided we have unity coupling in our transformer. C_1 includes the capacitance in the primary divided by the square of the transformation ratio. The source resistance r_s may be a loading resistor, the surge impedance of a transmission line, the radiation resistance of an antenna, etc. The voltage applied to the grid is according to Thevenin's theorem

$$E_{gk} = E_s T. \tag{2}$$

The output of the tube is

$$E_o = E_{gk} [(\mu Z_L)/(r_p + Z_L)] \tag{3}$$

but since Z_L is determined by an external loading resistance which is in shunt with the plate resistance r_p according to the relation

$$(r_p Z_L)/(r_p + Z_L) = 1/(\Delta\omega C_o) \tag{4}$$

or

$$Z_L = r_p/(\Delta\omega C_o r_p - 1) \tag{5}$$

if (1) and (5) are substituted into (3), and both sides are divided by E_s, the amplification A is

$$A_{gc} = E_o/E_s = [1/(\sqrt{\Delta\omega C_1 r_s})] [1/(\Delta\omega C_o r_p)]. \tag{6}$$

If the stage under consideration is preceded by a similar stage, we may set r_s equal to $1/(\Delta\omega C_o)$, in which case we obtain an over-all response of 0.5 at $f_o \pm (\Delta f/2)$, and by replacing μ/r_p by g_m, (6) will take the convenient form of

$$A_{gc} = g_m/(\Delta\omega\sqrt{C_1 C_o}) \tag{7}$$

a formula equally useful for tiodes or pentodes if feedback can be neglected.

The equivalent noise resistance of the grounded-cathode amplifier is given by

$$R_{n\ \mathrm{equ}} = 2.2/g_m \tag{8}$$

while for the pentodes

$$R_{n\ \mathrm{equ}} = [2.2/(g_m(1+\alpha))]\ [1 + q\alpha\ (I_b/g_m)]. \tag{9}$$

$$\alpha = I_{c2}/I_b$$

I_b is the plate current, and I_{c2} is the screen current. The root-mean-square grid-noise input may be calculated then with the aid of the equation

$$\sqrt{\overline{e_{gn}^2}} = 1.3 \times 10^{-10}\ \sqrt{R_{n\ \mathrm{equ}}\ \Delta f}. \tag{10}$$

The cathode-input amplifier, with its equivalent network, is shown in Figure 2. Again for bandwidth considerations we make the assumption that $R_s T_1^2 = 1/(\Delta\omega C_1)$ (see (1)) and

$$[(r_p T_o^2)/(\mu+1) + R_L]/[(r_p T_o^2)/(\mu+1)\ R_L] = \Delta\omega C_o \tag{11}$$

where the input capacitance C_1 is equal to the sum of the reduced grid-cathode capacitance due to degeneration[5] and the incidental capacitance to ground, while R_L is the equivalent parallel resistance of the losses in the output circuit. By rearranging (11),

$$R_L(\mu+1)/T_o^2 = r_p(R_L\Delta\omega C_o - 1). \tag{12}$$

The amplification is

$$A_{co} = T_o\left(\frac{\mu}{\mu+1}\right)\left[\frac{R_L/T_o^2}{\dfrac{r_p}{\mu+1} + \dfrac{R_L}{T_o^2}}\right]$$

$$= T_o \left(\frac{\mu}{\mu+1}\right) \left\{ \frac{\dfrac{R_L(\mu+1)}{T_o^2}}{r_p + \dfrac{R_L(\mu+1)}{T_o^2}} \right\}. \tag{13}$$

If we substitute from (12)

$$A_{co} = T_o \left(\frac{\mu}{\mu+1}\right)\left(\frac{R_L \Delta\omega C_o - 1}{R_L \Delta\omega C_o}\right). \tag{14}$$

If $\mu >> 1$ and we substitute for T_o from (12)

$$\begin{aligned} A_{co} &= \sqrt{\frac{\mu R_L}{(R_L \Delta\omega C_o - 1)\, r_p}} \left(\frac{R_L \Delta\omega C_o - 1}{R_L \Delta\omega C_o}\right) \\ &= \sqrt{\frac{\mu}{r_p}} \left(\frac{1}{\sqrt{\Delta\omega C_o}} - \frac{1}{\sqrt{R_L \Delta\omega C_o}}\right) \end{aligned} \tag{15}$$

if $\sqrt{R_L}$ is high we may replace μ/r_p by g_m, (15) will take the form

$$A_{co} = \sqrt{g_m/(\Delta\omega C_i)}. \tag{16}$$

Comparing (7) and (16), we may notice that the latter is in the order of the square root of the former, thus two cascade stages are required for amplification of the same order of magnitude. The equivalent noise resistance in this case is equal to that of the grounded-cathode amplifier.

The cathode-input amplifier, with its equivalent network, is shown in Figure 3. This circuit may be analyzed in two ways. In one instance, the input impedance of the tube, which is usually very low, is matched to a predominantly resistive input, such as an antenna, a transmission line, etc. In the second case, the transformation ratio is reversed and the input impedance loads a tuned circuit to provide the required bandwidth.

In the first case, for optimum power transfer

$$r_s T^2 = (r_p + Z_L)/(\mu + 1) \approx (r_p + Z_L)/\mu. \tag{17}$$

From the equivalent network, it may be seen that

$$E_k \mu + E_s T = I(r_s T^2 + r_p + Z_L) \tag{18}$$

and

$$E_k = E_s T - I r_s T^2 \tag{19}$$

then

$$I = [(\mu + 1)\, E_s T]/[(\mu + 1)\, r_s T^2 + r_p + Z_L]. \tag{20}$$

If we multiply (20) with the plate load Z_L and divide through with E_s, we obtain

$$A_{ci} = E_o/E_s = [(\mu + 1) T Z_L]/[(\mu + 1) r_s T^2 + r_p + Z_L]. \tag{21}$$

If we substitute from (17) for T, we obtain

$$A_{ci} = [(\mu + 1) Z_L] \quad \left[\left(\frac{\mu + 1}{\mu} + 1\right)\left(\sqrt{\mu r_s (r_p + Z_L)}\right)\right]. \tag{22}$$

If $\mu >> 1$, equation (22), after substitution for Z_L from (5), takes the form

$$A_{ci} = (1/2) \sqrt{\mu/[\Delta\omega C_o r_s (\Delta\omega C_o r_p - 1)]}. \tag{23}$$

For the second case, when the cathode-input amplifier operates from a tap on a tuned circuit fed by a comparatively high impedance, such as another stage of amplifier,

$$T^2 = (\mu + 1)/(r_p \Delta\omega C_0) \approx \mu/(r_p \Delta\omega C_0). \tag{24}$$

If this value is substituted in the gain equation

$$A_{fi} = [(\mu + 1) Z_L]/[(r_p + Z_L) T] \tag{25}$$

if $\mu >> 1$ yields the equation after substitution for Z_L from (5)

$$A_{ci} = \sqrt{g_m/(\Delta\omega C_0)}. \tag{26}$$

The equivalent noise input may be calculated from (8) with the aid of the equation

$$\sqrt{\overline{e_{kN}{}^2}} = \left(\frac{1.3 \times 10^{-10} \sqrt{R_{n\ \mathrm{equ}} \Delta f}}{Z_L + r_p + R_1(\mu + 1)}\right) \mu R_1. \tag{27}$$

A compromise between the grounded-cathode and cathode-input amplifier may be obtained by connecting the ground to a tap on the input transformer, as shown in Figure 8. From the equivalent network we may see that

$$E_sT_1 + \mu E_kT_2 = I(r_sT_1^2 + r_p + Z_L) \tag{28}$$

and

$$E_k = E_sT_1 = Ir_sT_1^2. \tag{29}$$

If we solve for I, we obtain

$$I = [(\mu T_2 + 1)E_sT_1]/[(\mu T_2 + 1)r_sT_1^2 + r_p + Z_L] \tag{30}$$

which is the same as (20) except that in place of μ we have μT_2, and, accordingly, we increased the amplification factor by T_2.

The cathode-coupled twin-triode amplifier is shown in Figure 4, with its equivalent network. From the equivalent network we may observe that

$$\begin{aligned} i_1r_p + (i_1 - i_2)Z_k &= \mu E_{gk} = \mu(E_1 - E_k) \\ &= \mu E_1 - (i_1 - i_2)\mu Z_k \end{aligned} \tag{31}$$

and

$$(i_2 - i_1)Z_k + i_2(r_p + Z_L) = \mu E_k \tag{32}$$

from (31)

$$i_1[r_p + Z_k(\mu + 1)] - i_2[Z_k(\mu + 1)] = \mu E_1 \tag{33}$$

and from (32)

$$-i_1[Z_k(\mu + 1)] + i_2[r_p + Z_L + Z_k(\mu + 1)] = 0. \tag{34}$$

If (33) is divided through with $r_p + Z_k(\mu + 1)$, we have

$$i_1 - i_2[Z_k(\mu + 1)]/[r_p + Z_k(\mu + 1)] = (\mu E_1)/[r_p + Z_k(\mu + 1)] \tag{35}$$

and if (34) is divided through with $r_p + Z_k(\mu + 1)$, we have

$$-i_1 + i_2[r_p + Z_L + Z_k(\mu + 1)]/[Z_k(\mu + 1)] = 0. \tag{36}$$

If we add (35) and (36) we have

$$i_2\left[\frac{r_p + Z_L + Z_k(\mu + 1)}{Z_k(\mu + 1)} - \frac{Z_k(\mu + 1)}{r_p + Z_k(\mu + 1)}\right] = \frac{\mu E_1}{r_p + Z_k(\mu + 1)}. \tag{37}$$

Thus the solution for the plate current of T_2 is

$$i_2 = \frac{\mu E_1 Z_k(\mu+1)}{r_p^2 + Z_L r_p + 2Z_k r_p(\mu+1) + Z_k Z_L(\mu+1)}. \tag{38}$$

If we multiply through with Z_L and divide with E_1, we obtain the amplification from grid number 1 to plate number 2

$$A_{CCTT} = \frac{\mu Z_L Z_k(\mu+1)}{r_p^2 + r_p Z_L + 2Z_k r_p(\mu+1) + Z_k Z_L(\mu+1)}. \tag{39}$$

If Z_k is much larger than $(r_p + Z_L)/(2(\mu+1))$, which is an easy condition to fulfill, (39) will take the form

$$A_{CCTT} = \mu Z_L/(2r_p + Z_L). \tag{40}$$

If we substitute $\Delta\omega C_o$ for $(2r_p + Z_L)/(2r_p Z_L)$ and multiply by the input-circuit transfer $\sqrt{C_o/C_i}$, where C_o is the output capacitance of the preceding stage factor (assumed to be equal to that of the stage under analysis) and C_i is the input capacitance corrected for degeneration, we have

$$A_{CCTT} = g_m/(2\Delta\omega\sqrt{C_i C_o}). \tag{41}$$

The grids of both tubes are at equal gain points with respect to their cathodes (in other words, the same gain is obtained from either grid to the output of the plate circuit, when the signal is applied between the grid and the cathode), and thus both tubes contribute equally to the total noise. The apparent noise generating resistances in either grid is equal to (8), and therefore the equivalent noise resistance between the input grid and cathode is

$$R_{n\ \text{equ}} = 4.4/g_m \tag{42}$$

where g_m is the transconductance of one tube unit. The root-mean-square grid-noise equivalent may be determined with the aid of (10). The equivalent noise resistance of the cathode-coupled twin-triode amplifier is considerably better than that of a pentode amplifier, and therefore a great improvement can be obtained in the noise factor by using cathode-coupled amplifiers in the early stages of amplification. A particularly useful instance is when cathode-coupled intermediate-frequency amplifiers are used after low-gain frequency converters.

To evaluate C_i, we calculate from the equation

$$C_i = C_D + C_{gp} + C_{gk}(1 - A_1) \tag{43}$$

where C_D is the incidental (socket, wiring, coil, etc.) capacitance, C_{gp} is the grid-to-plate capacitance, and C_{gk} is the cathode-grid capacitance. A_1 is the amplification of the first tube only.

$$\begin{aligned} A_1 &= \mu Z_k/[r_p + Z_k(\mu + 1)] \\ &= [\mu/(\mu + 1)]\,[(r_p + Z_p)/(2r_p + Z_p)]. \end{aligned} \tag{44}$$

If $\mu >> 1$ and we substitute

$$Z_p = 2r_p/(2r_p\Delta\omega C_0 - 1) \tag{45}$$

into (44) we have

$$A_1 = (2\Delta\omega C_0 r_p + 1)/(4\Delta\omega C_0 r_p). \tag{46}$$

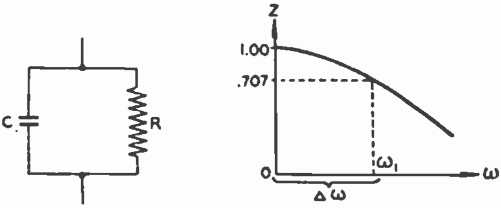

Fig. 15—Impedance characteristic of low-pass filter.

APPENDIX II

The bandwidth in the present paper is considered as the frequency, or the separation between the frequencies, at which the amplification is reduced by a factor of $1/\sqrt{2}$ of the value at the frequency of maximum amplification. The gain is a direct function of the impedance of the output circuit; therefore we may examine the impedance, and particularly its absolute value, directly.

In the case of a simple resistance-capacitance circuit as shown in Figure 15, the absolute value of the admittance at

$$|Y| = \frac{\sqrt{2}}{R} = \left|\frac{1}{R} + j\omega_1 C\right|. \tag{47}$$

If we multiply by R and rationalize

$$\sqrt{2} = \sqrt{1 + \omega_1^2 C^2 R^2} \tag{48}$$

since $\Delta\omega = \omega_1 - 0$,

$$\Delta\omega = 1/(RC). \tag{49}$$

In the case of the band-pass analogy of this circuit, shown in Figure 16, the admittance at the resonant frequency ω_0 is $1/R$, and at the frequencies ω_1 and ω_2 the absolute value of the admittance is

$$|Y| = \frac{\sqrt{2}}{R} = \left| \frac{1}{R} + j\omega_c C - \frac{j}{\omega} \right|. \tag{50}$$

If we multiply by R and rationalize, (50) becomes

$$\sqrt{2} = \sqrt{1 + \omega_c^2 C^2 R^2 [1 - (\omega_0^2/\omega_c^2)]^2} \tag{51}$$

where $\omega_0^2 = 1/LC$ and $\omega_c = \omega_1$, or ω_2.

If (51) is squared, it yields

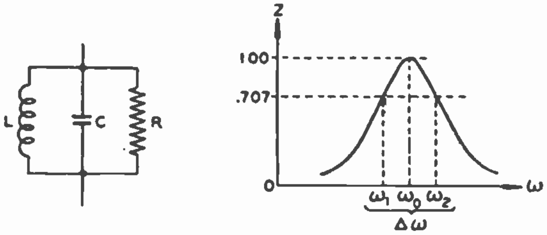

Fig. 16—Impedance characteristic of band-pass filter.

or

$$\left.\begin{array}{ll} \omega_c RC[1 - (\omega_0^2/\omega_c^2)] = 1 & \text{if} \quad \omega_c > \omega_0 \\ \omega_c RC[1 - (\omega_0^2/\omega_c^2)] = -1 & \text{if} \quad \omega_c < \omega_0 \end{array}\right\}. \tag{52}$$

If we rearrange (52), and solve the quadratic

$$\omega_c^2 \pm [1/RC]\,\omega_c - \omega_0^2 = 0 \tag{53}$$

$$\left.\begin{array}{l} \omega_1 = \dfrac{-[1/(RC)] \pm \sqrt{[1/(R^2C^2)] + 4\omega_0^2}}{2} \\ \omega_2 = \dfrac{[1/(RC)] \pm \sqrt{[1/(R^2C^2)] + 4\omega_0^2}}{2} \end{array}\right\}. \tag{54}$$

Since $\Delta\omega = \omega_2 - \omega_1$, $\Delta\omega = 1/RC$.

IMPROVED CATHODE-RAY TUBES WITH METAL-BACKED LUMINESCENT SCREENS*†

By

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Summary—Considerably improved cathode-ray tubes result from the application of a light-reflecting, electron-pervious, thin metallic layer on the beam side of the luminescent screen. Although this has been realized for some time, it is only recently that practical methods for applying such a metallic layer in kinescopes have been developed.

Observations and measurements on such tubes, using aluminum for backing, show that under appropriate conditions such tubes possess many advantages over similar conventional tubes. These are:

1. *Improved efficiency of conversion of electron beam energy into useful light—in other words, more useful light output for a given beam power input.*
2. *Elimination of ion spot—thus making other, generally less direct, means for eliminating the ion spot unnecessary.*
3. *Improved contrast.*
4. *Elimination of secondary emission restrictions—thus permitting the use of high voltages and screen materials with poor secondary emission.*

ONE of the outstanding quests in the cathode-ray tube field has been the search for means of increasing the brightness of the pictures on the face of the tube. Previous methods consisted primarily of efforts to increase beam power—that is, raising the voltage and increasing the current by improvements in electron optics—as well as a search for luminescent materials with greater efficiency in converting beam power into light. The most recent step in increasing light output is the application of a light-reflecting metallic layer on the beam side of the fluorescent screen.

Many practical tubes with metallic layers on the screens were built and used six and seven years ago. However, these tubes were limited to high-voltage operation and the metallic layers did not possess the light-reflecting properties which characterize the new metal films. The advantages of having a thin reflecting layer have long been anticipated and, to a limited extent, observed in the laboratory. It is only recently, however, that methods have been developed which will make such tubes possible and practical.

Before showing how this is accomplished, it is worthwhile to review briefly the pertinent part of the state of affairs at the luminescent

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* Decimal Classification: R583.6 × R138.313.

† Presented at the I.R.E. Winter Technical Meeting, January 24, 1946, in New York, N. Y. Reprinted from *RCA REVIEW*, March, 1946.

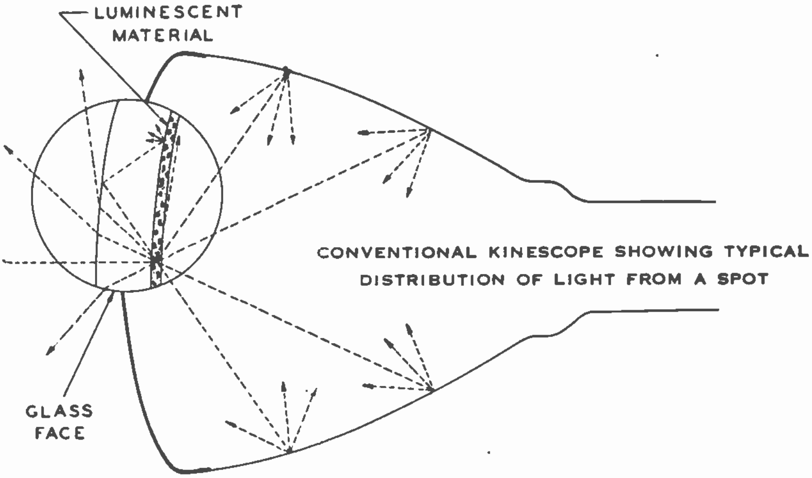
LUMINESCENT MATERIAL
CONVENTIONAL KINESCOPE SHOWING TYPICAL DISTRIBUTION OF LIGHT FROM A SPOT
GLASS FACE

Fig. 1.

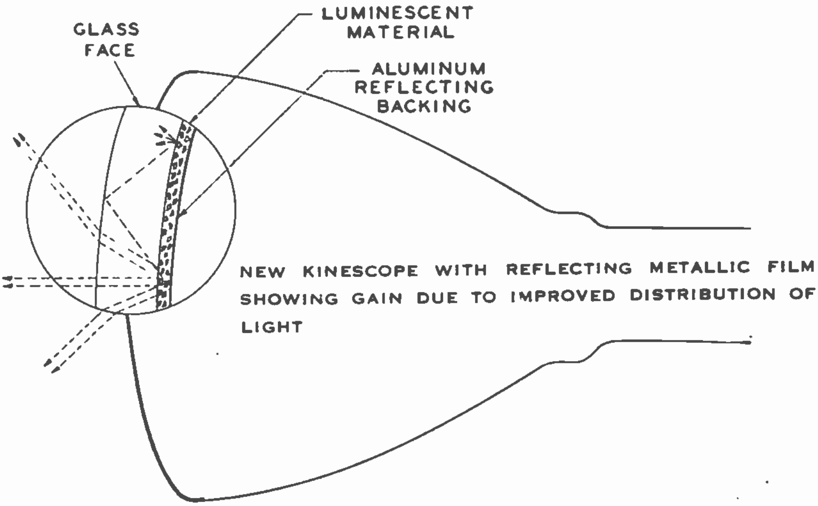
GLASS FACE
LUMINESCENT MATERIAL
ALUMINUM REFLECTING BACKING
NEW KINESCOPE WITH REFLECTING METALLIC FILM SHOWING GAIN DUE TO IMPROVED DISTRIBUTION OF LIGHT

Fig. 2.

screen in a conventional kinescope. This is shown schematically in Figure 1. The region in the circle is a greatly magnified and somewhat distorted diagram of a small section of the tube face of which one element is fluorescing. Generally, at least 50% of the light generated in the screen is emitted towards the electron gun in the tube. Another 15-25% is lost by total internal reflection in the glass of the tube face. Thus only about 25-35% of the total light generated is emitted in the forward direction to constitute the useful light output of the tube. It should also be pointed out here that some of the wasted light is harmful in that it is scattered back onto the screen to set a limit on possible contrast in the picture. There are several mechanisms for this. One is the back-scattering of the light which strikes the inside walls of the tube; although the light is largely absorbed by the blackening on the wall, some of it comes back to the screen. Another is the light from one portion of the screen which can illuminate other regions directly because of the curvature of the face. Some of the totally reflected trapped light in the glass is reflected back onto the screen and is scattered causing what is known as halation.

Figure 2 shows a tube whose screen is covered with an electron-pervious, but light-reflecting, metallic layer. Now it is seen that the light which previously would go towards the rear (electron gun) is reflected forward into the direction of viewing. Thus without an increase of light generated, the efficiency of conversion of electron beam power into useful light has been increased. At the same time, some of the limitations on contrast have been removed, that is, the back-reflected light and the effects due to curvature of the face. Experiments show that the large area contrast is considerably improved by a factor of three to ten times; the detail contrast, being primarily limited by halation, is only slightly improved.

The properties which this metallic layer should possess are: (1) it should be thin enough and of the right kind of metal to cause negligible absorption of the electron beam at the desired operating voltages; (2) it should be opaque, relatively smooth, and highly light-reflecting, so as to act as a mirror; (3) it should have sufficient conductivity to conduct the full beam current; (4) it should be strong enough to withstand the stresses due to effect of the focussed electron beam; (5) it should be durable enough to be able to withstand the necessary subsequent processing of the tube; and (6) it should be of a metal that will not chemically react with the luminescent screen material.

The metal chosen to work with is aluminum, because it combines properties which provide the best compromise in meeting the above conditions. It is easily applied by evaporation and does not affect luminescent screens. Its ability to meet condition (1) is indicated in Figure 3 where are shown a group of calculated curves giving the fraction of electron beam power that is passed by films of various thicknesses, as a function of the initial beam voltage. It will be noticed that a 10,000-volt beam will retain only 15% of its incident energy on passing through an aluminum film 5,000 Å thick. A film 2,000 Å thick will pass fifty-seven per cent and a film 1,000 Å will pass about 77% of the energy. If we assume that the effect of the mirror is to double the apparent brightness, then it is evident that a tube operating at 10,000 volts and with a film about 3,000 Å thick will show no difference from an unaluminized standard tube at the same voltage. It should be noticed, however, that a moderate increase in voltage causes a rapid

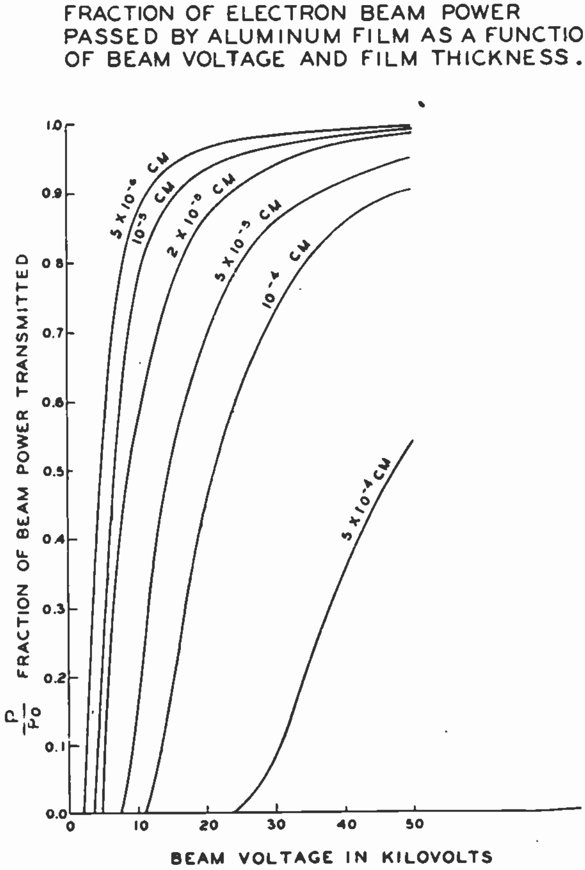

Fig. 3.

decrease in percentage loss of beam energy in the film. Experience has indicated that the most useful range of film thickness is between 500 Å and 5,000 Å.

In order for the film to meet condition (2), "that it be relatively smooth and mirror like", it has been found possible to cover the fluorescent screen with a thin film of organic material stretched over the crystals like a blanket. This provides a smooth surface upon which the aluminum can be evaporated. If such an intermediate film is not present, the aluminum will be broken up on evaporation so that it will not have its reflecting properties nor will it be continuous and conducting in the thicknesses necessary for low voltage operation. In order to obtain conductivity without the organic film, it would be necessary to evaporate five to ten times as much aluminum as is now necessary. This is why previous metallized screen tubes were restricted to high voltage operation.

These earlier tubes were aluminized in order to avoid undesirable effects due to poor secondary emission from the screen. It can readily be shown that if one tries to operate tubes at a voltage such that the secondary emission ratio from the screen is less than one to one, the screen will accumulate sufficient charge to slow up approaching electrons to a velocity at which the secondary emission is unity. This means that the screen is effectively operating at a voltage that may be considerably less than that applied to the tube. This is known as the "sticking" effect and is almost entirely corrected by providing a conducting layer over the screen. The new method of providing an aluminum film makes possible the correction of the effects due to secondary emission difficulties in tubes operating in the voltage range in which kinescopes are now operated. Thus the choice of luminescent materials for the screen is enlarged and improved techniques for applying these screens to the tube face are made available.

This aluminum film also provides a new line of attack on the old television tube problem of ion spot. Ions can be completely stopped by a film of aluminum that will readily pass electrons. Experience with tubes in the laboratory has shown that, with the right set of conditions —such as proper aluminum thickness and reasonably low gas pressure —tubes can be made which will show no ion spot at normal operating voltages.

Among other advantages of the aluminum film are the protection of the phosphor during processing and life and the improvement of the stability of the pattern with regard to displacement due to surface leakages on the face of the tube such as are produced if one touches the face of an operating kinescope.

Figure 4 gives the efficiency in candle power per watt as a function of applied voltage and at a fixed beam current obtained for two laboratory-made 12-inch tubes, identical except that one was aluminized. These curves are typical of measurements made on a number of tubes. It is to be noted that at the lower voltages the unaluminized tube has the higher efficiency whereas above the cross-over voltage the aluminized tube has the higher efficiency. The cross-over voltage which is controlled by the aluminum thickness is primarily dictated by such considerations as operating voltage and ion spot elimination. As seen from Figure 4, the increase in efficiency above the cross-over voltage is quite considerable; for luminescent screens with poor secondary emission characteristics, the gain may be considerably greater than that shown on the figure.

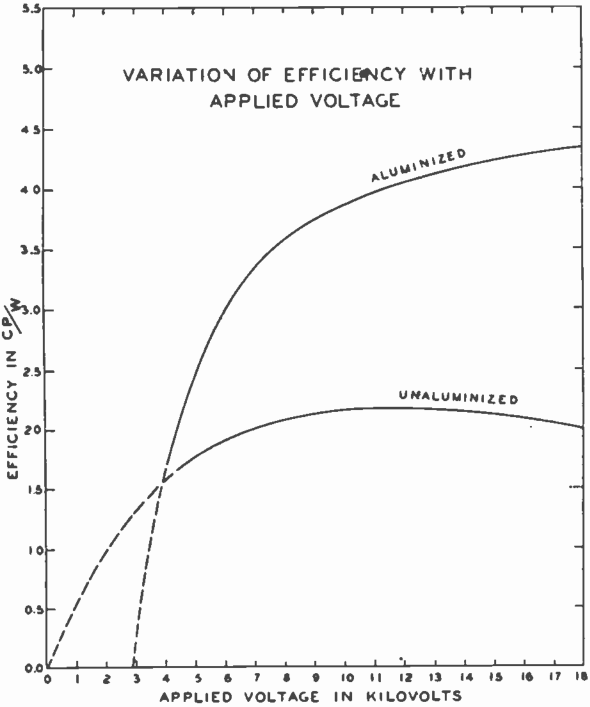

Fig. 4.

LOCAL OSCILLATOR RADIATION AND ITS EFFECT ON TELEVISION PICTURE CONTRAST*†

By

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Summary—The objects of this paper are (1) to investigate the effect on a television receiver of a c-w † interfering signal which lies in the high end of the picture band, (2) to set up a maximum permissible interference level, and (3) to correlate this level with radiation from the local oscillator of superheterodyne receivers.

It was observed that the chief annoying effect of interference at the high end of the video band was a loss in contrast. A strong interference, in fact, caused a complete loss in contrast or even a negative picture. Overall contrast gradation curves were computed theoretically which checked the experimental observations; the observations and computations indicated that a 20 decibels signal-to-interference field strength ratio at the antenna is a minimum satisfactory value. To maintain this ratio in a 500 microvolt-per-meter region of a desired transmitter, nearby receivers must have a radiation below 0.01 microwatts. Pre-war receivers, which used no radio frequency stage, radiated 100,000 times as much as this and were extremely unsatisfactory. A grounded-grid triode radio frequency stage may give a reduction of about 30 decibels or more and a pentode radio frequency stage may be made even better. Other remedies are also discussed but all increase receiver cost somewhat. However, it is made clear that an adequate television service will require suppression of radiation if the frequency assignments are such as to make interference possible.

† Throughout this paper "c-w" indicates "continuous wave".

I. Introduction

THE interference caused by local-oscillator radiation from superheterodynes has long been recognized as an important problem in receiver design. In spite of this, very few published papers indicate quantitatively how much radiation is present from various receiver circuits or how much radiation might be considered tolerable. In the sound broadcast field, even with the commonly used multi-grid mixers and converters which give partial separation of the local oscillator from the antenna, the radiation problem is serious in the short-wave bands.[1, 2, 3] In television reception, it has been common practice

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* Decimal Classification: R583.15.

† Reprinted from *RCA REVIEW*, March, 1946.

[1] R. Moebes, "The Superheterodyne Receiver as a Source of H-F Interference," *Telégr.-Fernspr.-Funk-u. Fernschtech.*, Vol. 29, pp. 199-201, July, 1940.

[2] R. Moebes, "On the Permissible Value of Local Oscillator Voltage at the Antenna of Superheterodyne Receivers," *Telégr.-Fernspr.-Funk-u. Fernschtech.*, Vol. 31, pp. 217-222, August, 1942.

[3] G. S. Wickizer, "Radiation from Superheterodyne All-Wave Receivers," unpublished report of RCA Communications, Inc., April 7, 1937.

to use triode or pentode mixer tubes because of their high signal-to-noise ratio[4] and, when no radio frequency stage is used, the radiation is high. In the New York area, the channel assignments throughout the war were such that, with the usual 12.75-megacycle intermediate frequency, a receiver tuned to channel 1, 50 to 56 megacycles, radiated a local-oscillator frequency of 64 megacycles which lay in the upper video frequencies of channel 2, 60 to 66 megacycles. Post-war frequency assignments and choice of intermediate frequencies will undoubtedly be different but the problem remains and is the reason for the writing of this paper. The work to be described is also applicable to other types of c-w interference (such as sound carriers in the picture channel and harmonic radiation from amateur and other services) and, to a lesser extent, to certain types of noise interference.

To the writer's knowledge, only one study of the interference problem in television has been published to date.[5] This study provided an excellent start, but was made using British television standards, with viewing tubes and picture pick-ups in common use at the time, and was entirely subjective. Furthermore, when c-w interference was studied, the interference was introduced into the video circuit so that an interference pattern might be observed at all light levels. Practically, when the interference comes through a receiver antenna circuit and with light levels such that the picture carrier is of very small amplitude, the interference pattern may not be observable on a kinescope or viewing tube. The U. S. standards, which incorporate negative modulation and vestigial sideband operation, require other special consideration. The present report is intended to treat the problem when U. S. standards are used; the conclusions will be based on objective analysis supported by a subjective study.

In the reception of a television picture, a small interfering signal will give rise to a pattern which can sometimes be observed on the viewing screen at certain light levels. If the interfering frequency is close to the picture carrier, the "beat" interference is of low frequency and gives rise to relatively large vertical or horizontal bars (i.e., large detail patterns). Jarvis and Seaman[5] showed that such a condition is the most annoying to the viewer, particularly when the bars are stationary, i.e., the beat frequency is synchronized with the scanning system. With present U. S. standards, the video channel is about 4

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[4] E. W. Herold, "Superheterodyne Converter System Considerations in Television Receivers," *RCA REVIEW*, Vol. 4, No. 3, pp. 324-337, January, 1940.

[5] R. F. J. Jarvis and E. C. H. Seaman, "The Effect of Noise and Interfering Signals on Television Transmission," *P. O. E. E. Jour.* (Brit.), Vol. 32, pp. 193-199, October, 1939.

megacycles or more wide so that low-frequency beat interference (under 1 megacycle) is not as probable as higher frequency beats (i.e., small detail patterns). Furthermore, it would be wise to choose an intermediate frequency so that receiver local-oscillator radiation will not produce the most annoying interference, namely, large detail patterns. In the present study, therefore, only higher frequency beats will be considered and, since synchronization with the scanning system is unlikely, it will be less important to consider the annoyance of the possible small-detail fluctuating pattern, and more important to consider other effects due to the interference. The chief one of the other effects is a degradation of picture quality due to a loss in contrast.

II. PICTURE CONTRAST WITH SMALL CONTINUOUS WAVE INTERFERENCE

In order to obtain an understanding of how picture contrast is affected by an interfering c-w signal, let us look at Figure 1. At (a) is shown a typical black-white transition in a simulated television modulated signal with conventional negative polarity (i.e., decreasing carrier for increasing light levels). During the black portion, the transmitter sends out nearly maximum output, increasing to a peak only for the synchronizing and blanking interval. During a white picture, the carrier of an ideal transmitter is very low, substantially zero for the brightest light values; only for the synchronizing pulses is peak carrier amplitude attained.* When such a signal is received, the final detector follows the carrier envelope and the direct current restorer system operates so that black level is set at the point shown. Such a signal will produce maximum light output on the viewing tube (kinescope) for the white part, and minimum light for the black part, the ultimate contrast range being set by the picture viewing tube capabilities, room lighting, etc.

In Figure 1 (b) is shown a similar received carrier combined with an interfering unmodulated carrier whose frequency is assumed to be such that the "beat" is in the high video range. Assuming the black level setting of the viewing tube is unchanged from Figure 1 (a), it is seen that the originally black portion of the signal envelope now has small periodic excursions toward white. If the "beat" is high,† the eye will not observe the checkered nature of the pattern so much as the fact that the general black level illumination has been raised, i.e., the blacks now look grey. With the idealized 100% modulated carrier

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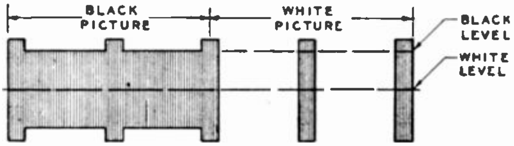
* In practical transmitters, 100% modulation is not always reached so that white level may correspond to a larger carrier than shown on Figure 1. This changes the effects described quantitatively by a small amount but, qualitatively, the idealized 100% modulation herein treated is entirely adequate to explain the behavior.

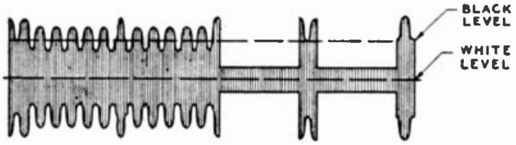
† Or if the viewing distance is sufficiently large.

here assumed, in the white portion of the picture, there is no "beat" since the picture carrier is zero.* The interference, however, causes a spurious carrier to appear at the receiver so that, instead of a completely white output, the brightness is decreased again toward the grey. Thus, both the black and the white portions of the picture are shifted toward each other, i.e., toward a neutral grey. The picture contrast has then suffered.

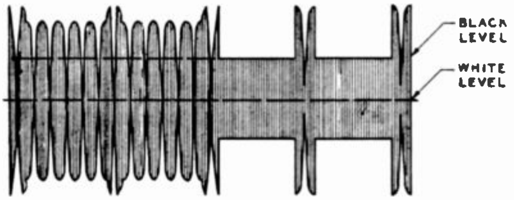
If the receiver controls are readjusted, of course, the over-simplified case of a black to white transition, which we have been discussing, can be corrected while the interference is present. With an actual picture, however, no correction is possible without an almost complete loss of the picture tone values at the extremes of brightness; the loss will be particularly serious in the darker portions and can be interpreted as a loss in gradation contrast or "gamma." If the direct current restoring system of the receiver follows the peak values of the "beat" during synchronizing intervals, there is less tendency for the white part to become



(a) RECEIVED CARRIER WITH NO INTERFERENCE



(b) SMALL INTERFERING C-W ADDED



(c) LARGE INTERFERING C-W ADDED, APPROX. EQUALLING PICTURE CARRIER

Fig. 1—Received television signals showing how an interfering c-w leads to loss in contrast, even to the point of a negative picture (case c).

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* It should be noted that this condition cannot be duplicated by introducing the interference in the video frequency band as was done by Jarvis and Seaman (Reference 2).

grey but the raised brightness of the black becomes worse. Practically, many present television receivers use a direct current restoring system which will follow the average, not the peak, of the synchronizing pulse whenever the "beat" lies above a megacycle or so. This is caused by the relatively poor high-frequency video response of the restorer input, since high-frequency response is not needed in this circuit. Thus, in these receivers, the interference considered here has no effect on the black-level setting and the observed effects will be substantially as indicated on Figure 1, i.e., the black appears grey and the brightness of the white parts is reduced. Similarly, receivers with a limiter ahead of the peak-operated type of direct current restorer, will operate very much like the average-operated type.

III. THE NEGATIVE PICTURE PRODUCED BY STRONG INTERFERENCE

In conducting experiments on the effect of c-w interference, it was observed that a strong interfering signal gave rise to a picture of reversed contrast, i.e., the dark portions of the original became the light portions of the reproduction and vice versa. Although this phenomena had been observed by others, for example when a strong sound carrier was tuned into the picture channel, it had usually been assumed that overloading occurred, or some other unusual behavior was present. However, the writer's experiments showed that the effect existed when there was no overloading and, indeed, was a straightforward extension of the contrast loss phenomenon described above. In fact, the experiments showed that, as the interfering carrier was increased, the picture contrast steadily decreased until, at a definite point, the picture was substantially "washed out." Further increase of interference gave a negative picture of rather poor contrast and, finally, with interference signals far in excess of the black-level picture signal the picture again disappeared. Synchronization was well maintained throughout, with little or no apparent effect due to the interfering c-w. The scanning return lines are, of course, visible in such a negative picture since there is no blanking.

Figure 1 (c) shows how a large interfering c-w can lead to a negative picture. When the interference approximately equals the picture carrier during black transmission, the "beats" produced alternately raise the received signal to double amplitude and reduce it to substantially zero amplitude. Thus, with a black transmitted picture, the viewing tube has excursions extending to full white, leading to an average brightness well up in the grey region. On the other hand, during white transmission, the transmitted carrier is not present and no "beats" occur. The interfering c-w simply replaces the normal picture carrier and makes the picture appear black. To summarize this, the black transmissions now appear grey and the white transmissions appear black, leading to a complete reversal of contrast (i.e., a negative gamma).

It was here assumed that the direct current restorer of the receiver is unaffected by the interference and, as already indicated, this is typical of the many receiver circuits in which either direct current restoring follows the average of the high video-frequency beat or in which limiters are used. A direct current restorer whose input contained all video components and whose output followed peak amplitude would not lead to a complete washout of the picture or a negative picture, although

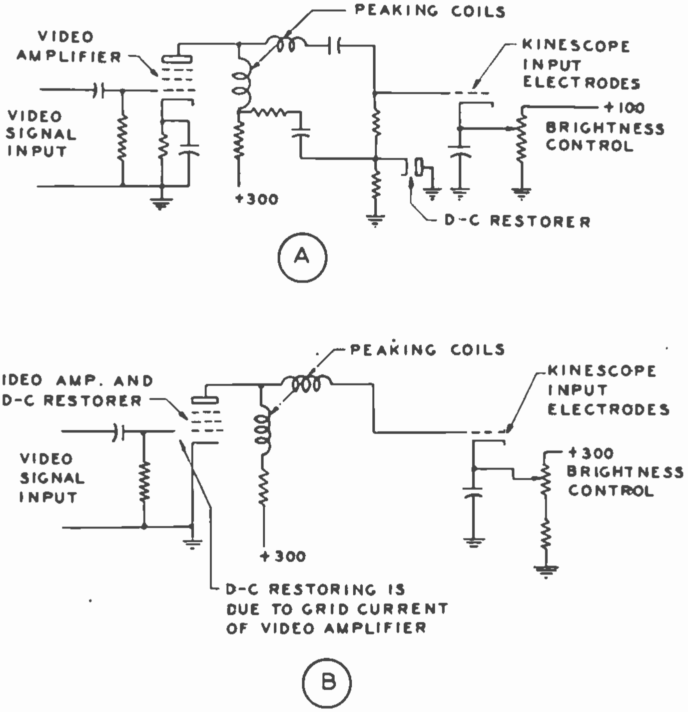

Fig. 2—(A) A direct current restoring circuit which does not follow high video frequencies and so permits strong interference to produce a negative picture; (B) a direct current restoring circuit which operates on the peaks of a synchronizing wave and so does not give a negative picture (unless a limiter precedes the circuit).

the loss in contrast is more serious than with the averaging type of direct current restorer. In this connection, two typical direct current restorer circuits are shown in Figure 2. One of these, Figure 2 (A), is of the averaging type and can give rise to a negative picture while the other, Figure 2 (B), is of the peak type and cannot (i.e., unless preceded by a limiter). In a later section of this paper calculations will be made which show the contrast loss of receivers with each type of restorer as a function of interfering amplitude.

IV. CONTRAST EVALUATION

To evaluate quantitatively the effect of interference on contrast, it is necessary to consider the various ways in which the contrast of a reproduced picture may be expressed. The simplest expression for the contrast is simply the over-all maximum brightness ratio, i.e., the ratio of the light from the brightest portion of the reproduced picture to the light from the darkest portion. In a perfect system, this ratio can be infinite since the darkest portion can be completely black. Maximum contrast ratio has been used to discuss kinescope performance[6] although it is then necessary to distinguish between halation effects and normal large-area contrast ratio. It is often stated[7] that a ratio of 35:1 or more is desirable in a television picture and such ratios are attainable with the best reproducing systems. We shall use degradation in maximum contrast ratio as one criterion for estimating the effect of c-w interference.

In photography, it has long been well known that, even if the maximum contrast ratio is fixed, startling changes in appearance are made possible by difference in the contrast *gradation*, i.e., the way in which various brightness values of an original are interpreted in the reproduction. The same is, of course, true in television. If a curve is drawn of reproduced light values as a function of original light values, complete information on contrast gradation is shown. Furthermore, such a curve may also indicate maximum contrast ratio by the ratio of the maximum to minimum light value at the ends of the curve. The over-all contrast gradation curve is, therefore, an even more significant measure of the degradation caused by an interfering signal; such curves will also be used in this paper.

Practically, whether or not an interfering signal is noticed depends upon the quality of the over-all system when free from interference. There are many grounds for believing that future television pictures will be far superior in their contrast range and low-light tone renditions than those which are presently called high in quality. In considering interference, therefore, it is well to concentrate on the effect which is obtained when the received picture is more nearly ideal, since this will have most value for the future. In this respect, an objective

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[6] R. R. Law, "Contrast in Kinescopes," *Proc. I.R.E.*, Vol. 27, pp. 511-524, August, 1939.

[7] P. C. Goldmark and J. N. Dyer, "Quality in Television Pictures", *Jour. Soc. Mot. Pic. Eng.*, Vol. 35, pp. 234-253, September, 1940.

study is at present more valuable than a subjective one made with less-than-ideal viewing tubes, etc.

V. COMPUTED EFFECTS OF INTERFERENCE

This section is concerned with the computation of the over-all contrast gradation curve when interference is present, and assuming an idealized kinescope. The video wave which results from envelope detection of a television signal which includes c-w interference is derivable as follows. If we call the picture carrier, as it arrives at the second detector, $A \sin \omega t$ and the interfering carrier is $B \sin (\omega + p) t$, then the second detector receives an over-all signal of

$$A \sin \omega t + B \sin (\omega + p)\, t = [\sqrt{A^2 + B^2 + 2AB \cos pt}\,] \sin (\omega t + \beta)$$

where β is a time-variable phase angle which is of no concern here. After detection only the envelope is of interest. It may be written

$$V_c = \sqrt{A^2 + B^2 + 2AB \cos pt} = (A + B) \sqrt{1 - \frac{4AB}{(A + B)^2} \sin^2 \frac{pt}{2}}$$

$$= (A + B) \sqrt{1 - k^2 \sin^2 \phi}\,, \tag{1}$$

where

$$k^2 = \frac{4AB}{(A + B)^2}$$

and

$$\phi = \frac{pt}{2}$$

The picture carrier amplitude, A, has a maximum value, A_{max}, during the synchronizing interval, and a value ¾ A_{max} at the black level.* With maximum brightness of the original picture, A is reduced to zero when the modulation is complete (100%). For intermediate brightness, and a constant transmitter gamma, A follows the relation

$$A = \tfrac{3}{4} A_{max} \left[1 - \left(\frac{L_T}{L_{max}} \right)^{\gamma_T} \right] \tag{2}$$

where L_T is the instantaneous original picture brightness, L_{max} is the maximum brightness and γ_T is the transmitter "gamma," or slope of the modulation characteristic when corrected for negative polarity of modulation and plotted on log-log paper.

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* According to U. S. television standards.

A sufficiently close approximation to an idealized kinescope characteristic is a power law over the range from cut-off to zero bias. We shall assume that the output light is constant for inputs beyond zero bias and zero beyond cut-off. Thus, the kinescope characteristic is represented by Figure 3. Mathematically, the light output is

$$L = K\,(V_{co} + V)^{\gamma_R} \text{ when } O < (V_{co} + V) < V_{co}$$

$$L = O \text{ when } (V_{co} + V) < O \qquad (3)$$

$$L = K\,V_{co}{}^{\gamma_R} \text{ when } (V_{co} + V) > V_{co}$$

where V_{co} is the magnitude of the voltage needed to cut off the tube, V is the applied bias and signal, and γ_R is the exponent of the power law.

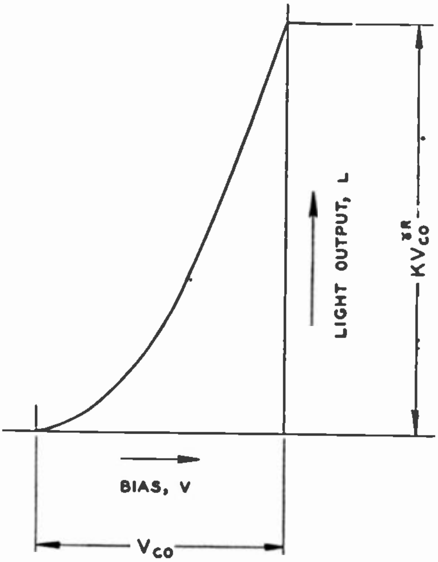

Fig. 3—Characteristic of an assumed power-law kinescope following the equation $L = K\,(V_{co} + V)^{\gamma_R}$

The manner with which the video signal is applied to the kinescope is shown in Figure 4. In Figure 4 (a) is shown the normal, interference-free case. It is seen that the receiver gain control is so adjusted that the range of black to white video signal, which is ¾ A_{max}, equals the assumed kinescope cut-off V_{co}. Furthermore, the direct current restorer, which operates from the synchronizing pulse, together with the kinescope bias control comprise a net bias, $V_{d\text{-}c}$ which sets the black level at cut-off. The difference between A_{max} and V_{co} is shown as ¼ A_{max} on the figure and is adjusted to this value by the bias control (often called the brightness control). The instantaneous video signal is shown as V_c (equation 1).

Figure 4 (b) shows the video signal when a high beat-frequency interference is applied and the direct current restorer is of the type shown in Figure 2 (a), i.e., it operates on the average of the synchronizing video pulse during the fluctuating beat. From equation (1) we find the average to be

$$\overline{V_c} = (A_{max} + \mathrm{B}) \frac{2}{\pi} \int_0^{\pi/2} \sqrt{1 - k^2 \sin^2 \phi}\, d\phi \qquad (4)$$

which can be evaluated for various values of B by the usual tables for the complete elliptic integral.[8]

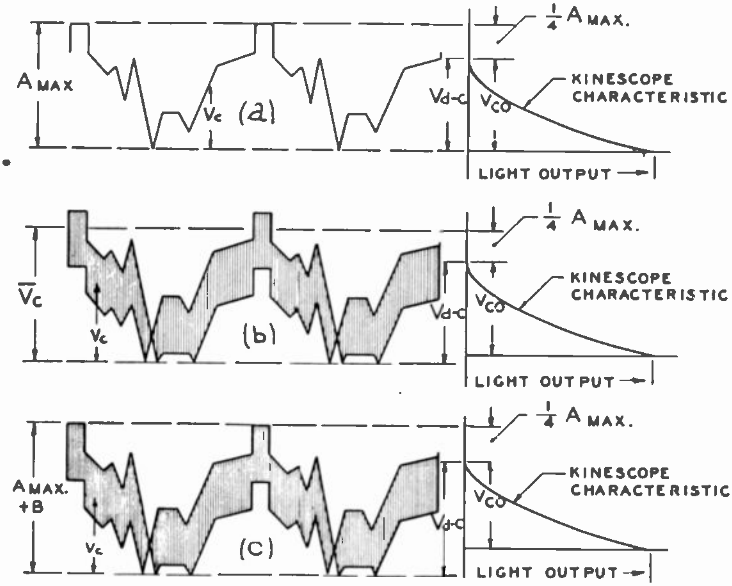

Fig. 4—This figure shows the video signal placement on the kinescope characteristic: (a) with no interference; (b) with high video beat interference and an average-operated direct current restorer; (c) with high video beat interference and a peak-operated direct current restorer.

Figure 4 (c) shows the case when a peak-operated d-c restorer is used without a limiter, such as the one of Figure 2 (b). In this figure, the direct current restorer operates from the peak signal during synchronizing, $A_{max} + B$. When a limiter is used, conditions will be sub-

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[8] B. O. Peirce, A SHORT TABLE OF INTEGRALS, p. 121, Third Edition, Ginn and Co.

stantially the same as in Figure 4 (b) since the increased peak values are clipped by the limiter. It is, of course, clear that a manual change of the kinescope brightness control can change the effect of one of the direct current restorers to that of the other. In this analysis, the controls will be assumed to remain at their normal setting when no interference is present.

Putting equation (1) in equation (3) and including the effect of d-c restorer and kinescope bias, $V_{d\text{-}c}$ (Figure 4) we see

$$\begin{aligned} L &= K\,(V_{d\text{-}c} - V_c)^{\gamma_R} \\ &= K\left[V_{d\text{-}c} - (A+B)\sqrt{1-k^2\sin^2\phi}\right]^{\gamma_R} \qquad (5) \end{aligned}$$

where we must remember the limitations imposed on the equation by the cut-off and zero bias points (see equation 3). These limitations are that

$$\left[V_{d\text{-}c} - (A+B)\sqrt{1-k^2\sin^2\phi}\right] \geq 0$$

and

$$\left[V_{d\text{-}c} - (A+B)\sqrt{1-k^2\sin^2\phi}\right] \leq \frac{3}{4}A_{max} \quad .$$

Each of these limiting conditions may be solved for a value of ϕ which will be needed as an integration limit when finding the average light output. Calling these ϕ_1 and ϕ_2 respectively, we find

$$\phi_1 = \sin^{-1}\sqrt{\frac{1}{k^2} - \frac{(V_{d\text{-}c})^2}{4AB}} \qquad (6)$$

$$\phi_2 = \sin^{-1}\sqrt{\frac{1}{k^2} - \frac{(V_{d\text{-}c} - \tfrac{3}{4}A_{max})^2}{4AB}} \qquad (7)$$

These angles have limiting values of 0 and $\pi/2$ respectively and these limits are used when the arguments of (6) and (7) are greater than unity or imaginary.

The average light output over the fluctuating beats is then

$$L = \left(1 - \frac{2}{\pi}\phi_2\right) K \left(\frac{3}{4}A_{max}\right)^{\gamma_R} + \frac{2}{\pi}K \int_{\phi_1}^{\phi_2} \left[V_{d\text{-}c} - (A+B)\sqrt{1-k^2\sin^2\phi}\right]^{\gamma_R} d\phi \qquad (8)$$

where the first term gives the light output when the instantaneous bias on the kinescope exceeds zero, and the integral gives the total light output averaged over the normal kinescope range. The integration is straightforward for $\gamma_R = 1$ and $\gamma_R = 2$ although the result involves the incomplete elliptic integral. Since tables for these are available[9], a numerical answer may be obtained, although the calculations are very laborious.

The square law relation, $\gamma_R = 2$, is a far better approximation to an actual kinescope than the linear one. The writer has carried through the calculation of equation (8) to find the over-all contrast gradation curves for different interference levels, using $\gamma_R = 2$ and assuming, in turn, each of the two types of direct current restorer which give

$$V_{d\text{-}c} = \overline{V_c} - \frac{1}{4} A_{max} \text{ (average-operated type)}$$

where $\overline{V_c}$ is found from equation (4), and

$$V_{d\text{-}c} = (A_{max} + B) - \frac{1}{4} A_{max}$$

$$= \frac{3}{4} A_{max} + B \text{ (peak-operated type)}$$

The calculations were made by assuming a complementary gamma at the transmitter of $\gamma_T = ½$ (equation 2). The curves can be corrected for other transmitter gammas by an appropriate compression or expansion of the abscissa scale.

Figure 5 shows the calculated reproduced light as a function of original light at the transmitter, using the average-operated direct current restorer. The curves are largely self-explanatory and show the marked decrease in contrast as the interference level is increased. Because vestigial sideband operation was assumed, it should be remembered that the interfering c-w receives 6 decibels more gain in the receiver than the desired picture carrier. Furthermore, with U. S. television standards, the black-level carrier is 2.5 decibels less than the peak carrier which is used to rate transmitters and field strengths. Thus the curve labeled "interference 8.5 decibels down" means an interfering c-w whose antenna field strength is 8.5 decibels less than the peak of the picture carrier field strength; at the second detector of the receiver, because of the increased amplification for the interference, the inter-

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[9] H. Hancock, ELLIPTIC INTEGRALS, John Wiley and Sons, New York.

ference is just equal to the black-level picture carrier. From the point of view of interference calculation, of course, it is the value at the antenna which matters, so that the curves are significantly labeled.

The negative picture for the stronger interference levels is clearly indicated by the reversed slope or negative gamma. One of the more striking features shown by Figure 5, is the rapidity with which contrast is lost as the interference level reaches a point 15 decibels below the picture carrier. Between the 14.5 decibels curve shown, which still gives a positive picture, and the 8.5 decibels curve, which gives a negative picture, the original contrast is substantially wiped out. In Figure 7 will be shown curves of maximum contrast ratio as a function of interference level which show that this rapid loss of contrast holds for other kinescope gammas as well.

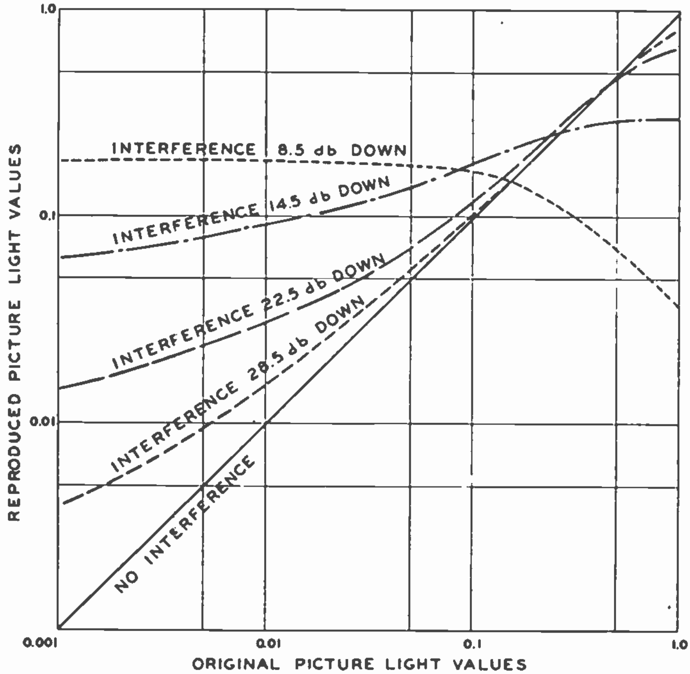

Fig. 5—Effect of c-w interference on television picture contrast using a direct-current restorer which follows the average of the synchronizing pulse. Kinescope gamma = 2, transmitter gamma = ½, and vestigial side-band operation using U. S. standards.

Figure 6 shows a set of contrast gradation curves for the other type of direct current restorer (such as that of Figure 2b). It is here found that no reversal of the picture takes place at any interference level, as had been expected. It should be remembered that the difference between the results of Figure 5 and those of Figure 6 lie only in the kinescope bias provided by the direct current restorer. Thus a manual adjustment of the bias control (brightness control) will change either set of curves into the other.

Since the gamma (γ_R) of existing kinescopes is often in excess of two, it is of interest to examine the contrast degradation for $\gamma_R = 3$ and $\gamma_R = 4$. Over-all contrast gradation curves, such as are given in Figures 5 and 6 for the square-law kinescope, are extremely tedious to compute for the higher-power laws. However, there is a simplification in equation 8 when the interference level is small ($B << A_{max}$). If only the end points are desired, i.e., the output light level for a completely black and a completely white transmission, the small-interference approximation is readily usable. Furthermore, for one case of large interference, namely, when $B = \frac{3}{4} A_{max}$, the end points can also be simply evaluated for higher gammas. If only the light outputs for white and also for black transmission are given, their ratio is the most easily understood evaluation of contrast degradation and, as discussed

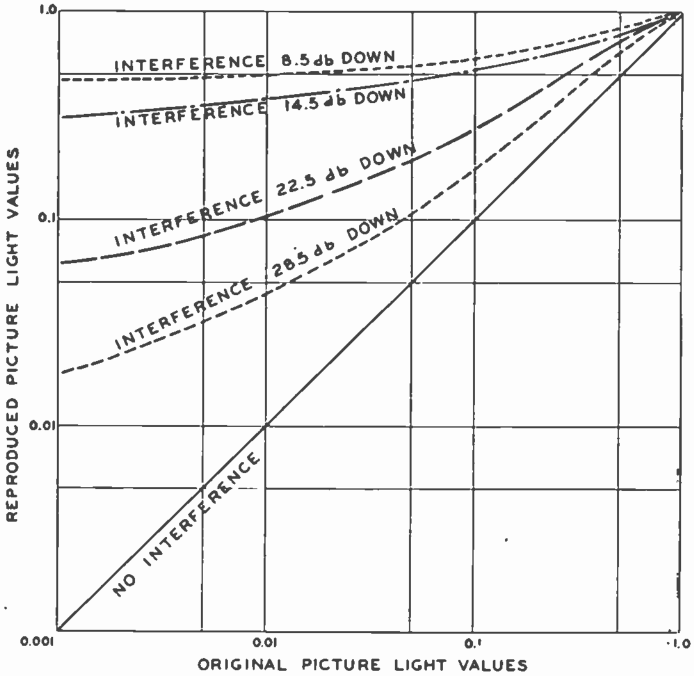

Fig. 6—Effect of c-w interference on television picture contrast using a peak-operated direct current restorer and no limiter. Kinescope gamma = 2, transmitter gamma = ½, vestigial side-band operation using U. S. standards.

in Part IV above, is called maximum contrast ratio. Figure 7 shows curves of the contrast ratio as a function of interference level for kinescope gammas of from 1 to 4, assuming the average-operated type of direct current restorer.

Examining Figure 7, we notice that the interference level at which the picture washes out (contrast ratio of unity) lies between —10 and — 13 decibels for all the kinescope power laws, and that the loss in contrast is quite rapid as this point is approached. If we assume a transmitter gamma, γ_T, which is the reciprocal of the kinescope gamma, γ_R, the interference-free picture for each of the assumed kinescopes will be the same. Figure 7 shows, however, that *small* interference has far less effect on the higher-gamma kinescopes. This illustrates the well-known advantage of gamma compression at the transmitter, with corresponding expansion at the receiver, in improving the signal-to-noise ratio. On the other hand, with *larger* interference, the curves eventually cross and the advantage is no longer present.

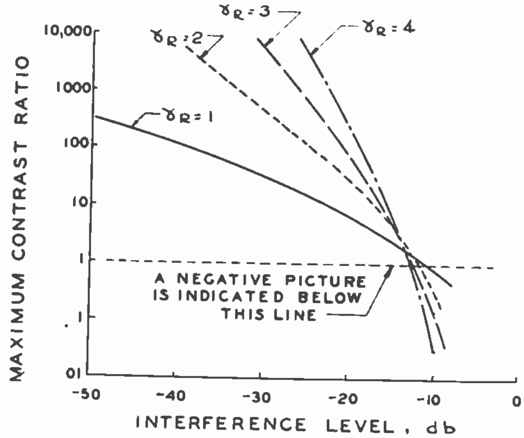

Fig. 7—Effect of c-w interference on ratio of light output during white transmission to light output during black transmission. An average-operated direct current restorer is assumed and the curves show the effect of different kinescope gammas.

VI. EXPERIMENTAL RESULTS AND ESTIMATED TOLERABLE INTERFERENCE

A television receiver in the writer's home was used to obtain experimental subjective data on the effect of interference. This receiver included the type of direct current restorer shown in Figure 2a except for a modification originally made to reduce possible effects of kinescope grid leakage. Normal program material and the test pattern from the National Broadcasting Company's New York Station, WNBT, was used. A Ferris Microvolter was used as the calibrated source of c-w interference and was connected across the receiver antenna transmission line (100 ohms impedance) through two 500-ohm resistors, so as not to interfere with the input alignment or impedance values. With no television signal present, a response curve of the receiver was measured and the diode second detector current calibrated in terms of the signal generator voltage. In this way, when the television signal came on, its relative magnitude with respect to the signal generator readings could be determined by its second detector current. It was necessary, of course, to use a substantially black picture to calibrate the black-level carrier of the received signal. It was then assumed that the peak carrier was 2.5 decibels higher, corresponding to U. S. standards. A check on this relative calibration of the received television signal was made by using a low-frequency "beat" interference and observing the magnitude of the beats on the synchronizing pulses as viewed on an oscillograph across the kinescope grid. The two methods checked very well.

Although many observations were made using various interfering frequencies, giving beats with the picture carrier from some tenths of megacycles to around 4 megacycles, most attention was given to a beat at 3.7 megacycles, well within the video band but at such a high frequency that the predominant effect was loss in contrast, rather than the very fine-grained pattern. In fact the interference pattern could hardly be observed with the kinescope and viewing distances used and might even pass completely unnoticed if attention was not called to it. The contrast changes, on the other hand, were very marked. In such subjective tests, it is not possible to obtain accurate data as to loss of contrast; fortunately, however, the transition point beween the positive and negative picture was quite clearly defined since it led to an almost complete wash-out of contrast values. The data are presented in tabular form in Table I and represent an average over a number of observations. In every case the receiver controls were set as for an interference-free picture and were left untouched for the observation.

Table I

Interference Beat Frequency	Interference Level, decibels	Observed Results
3.7 Mc	— 28	Barely perceptible loss in contrast
3.7 Mc	— 22	Substantial but tolerable loss in contrast
3.7 Mc	— 16	Intolerable loss in contrast
3.7 Mc	— 10	Completely washed-out picture
3.7 Mc	greater than — 5	Negative picture of poor contrast, return lines visible

From the computed data as presented in Figure 7 it is seen that a completely washed-out picture (contrast ratio $= 1$) was predicted with an interference of -10 to -13 decibels, depending on the kinescope gamma. Thus the computation and the observed value of Table I are in close agreement.

A comparison with the results of Jarvis and Seaman,[5] in England, is of some interest, in spite of substantial differences between their technique (which introduced the interference in the video channel instead of at the antenna) and even though the criterion they used was the annoyance value of the beat pattern, rather than contrast loss. When 8.5 decibels is added to their figures, to make them comparable with those used here for U. S. standards, the various results appear as shown in Table II.

Table II

Observed Result	Interference Level Using 3.7 Mc Beat and 525 Line System	Seaman and Jarvis Results With 2.0 Mc Beat and 405 Line System
Just visible change	— 28 db	— 26 db
Just tolerable change	— 22 db	— 19 db

On the basis of the calculations, as supported by the experiments, it is clear that relatively small differences in interference level near the critical point will cause rapid deterioration of the picture (see Figure 7). There can be no denying that the permissible interference limit must be below the point at which a complete wash-out of the picture occurs. If the interference is 10 decibels below this limit, a reasonable safety factor is allowed, although a noticeable deterioration of picture quality is still present. On these grounds we may say that the type of interference here considered, i.e., in the high-video range, should be at least 20 decibels below the picture carrier at the antenna. This number will be used as a criterion in the discussion of local-oscillator radiation below.

VII. QUANTITATIVE LIMITS ON RECEIVER LOCAL OSCILLATOR RADIATION

Since the superheterodyne receiver is here acting as an interfering transmitter, it is logical to measure its radiation in terms of the power which the local oscillator delivers to the antenna. With long transmission lines having appreciable loss, the radiated antenna power may be appreciably less than would be measured on a bench test with a receiver connected directly to a calibrated measuring receiver. Although this loss should be considered in special cases, the more general approach should assume no loss in the connection to the antenna, since negligible loss is readily obtained by use of good lines.

The type of interference considered here will be most serious between nearby antennas; thus complications introduced by propagation phenomena need not be considered. It can be assumed that "free-space" propagation will occur.* Thus a receiver radiating W watts into a half-wave dipole will give a field strength at a distance, d, of

$$E_I = \frac{\sqrt{45\,W}}{d}$$

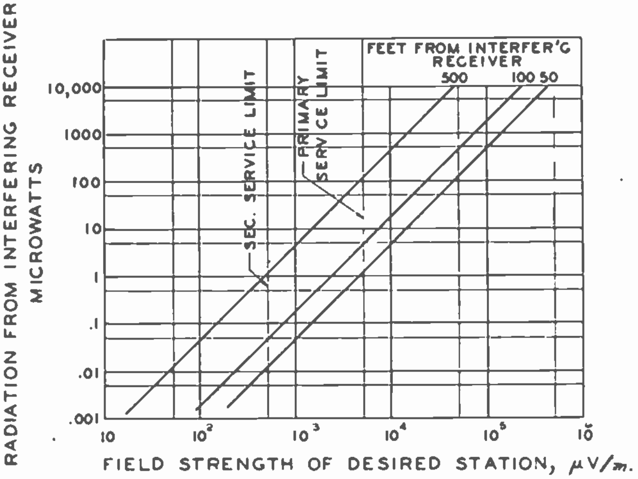

Fig. 8—Permissible radiation from television receiver to give an interference level of –20 decibels. A half-wave dipole is assumed on the receiver causing the interference.

If we consider a nearby receiver, with an antenna so situated as to receive both the interfering radiation and a desired signal from a television transmitter whose field strength is E_s, then, using the 20 decibels criterion,

$$\frac{E_s}{E_I} = \frac{E_s d}{\sqrt{45\,W}} \geqslant 10$$

Thus the radiated receiver power should be $$W \lesssim \frac{E_s^2\, d^2}{4500} \qquad (9)$$

This equation has been plotted in Figure 8 and shows that, to protect

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* The limiting distance for antennas 30 feet high, within which free space propagation may be achieved (on the average) is some 700 feet at 60 megacycles and greater than this at higher frequencies.

the 500-microvolt/meter area, the radiating receiver should radiate less than 0.01 microwatts if receivers are to be as close as 50 feet. This may be contrasted with prewar receivers using a 6AC7 mixer and no radio frequency stage which radiated in the order of 1,000 microwatts; to avoid annoyance to other receivers at a distance of 50 feet, it is necessary to have a signal of over 100,000 microvolts/meter. Fortunately, channel assignments, station hours, and the small number of existing receivers have been such as not to bring this problem into prominence except under special circumstances.* However, this situation cannot continue and measures should be taken to reduce radiation on all future receivers.

VIII. RECEIVER DESIGN CONSIDERATIONS

It is, of course, one thing to suggest 10^{-8} watts as a maximum permissible radiation and another to achieve it. A carefully designed pentode radio frequency stage between the antenna and a pentode mixer, together with a reasonably high intermediate frequency (so as to tune the oscillator far away from the band-pass of the amplifier), can be expected to provide enough attenuation.

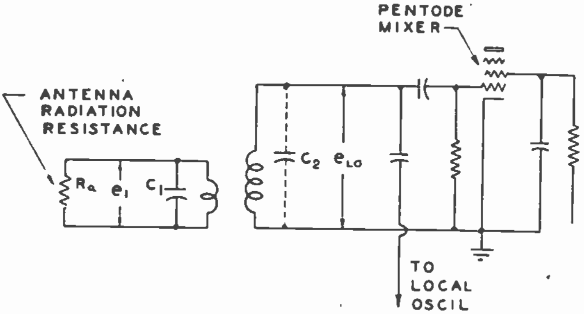

Fig. 9—Typical pentode mixer with double-tuned input circuit. The required local oscillator voltage is indicated as e_{LO}, whereas the resulting antenna voltage is shown as e_1.

To consider further, let us calculate the local oscillator radiation to be expected from the simplest television receiver with good performance, which uses a double-tuned circuit to couple the antenna to a pentode mixer. This type of input was commonly used in prewar receivers and is shown in simplified form in Figure 9. The local oscillator injection must be sufficient to give good results with the mixer, so that the local oscillator voltage across the input circuit, e_{LO} in the figure is fixed. Thus it is clear that the selectivity of the sec-

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* In the New York area, as already mentioned, prewar receivers tuned to WNBT (50—56 megacycles) radiated at 64 megacycles in the WCBW band (60—66 megacycles).

ondary of the double-tuned circuit is not effective in reducing the radiation. It can be inferred, on the same grounds, that a single-tuned circuit would have no selective effect at all on the local-oscillator radiation. In either case, however, the secondary capacitance does affect the radiation since its value determines the antenna-to-grid step-up for a given band width. With the double-tuned circuit, it can be shown that, to a fair approximation, the primary voltage, e_1 of Figure 9, induced by the local oscillator, is

$$e_1^2 \approx \frac{e_{LO}^2}{2} \frac{C_2}{C_1} \frac{1}{1 + 2\left(\frac{\omega_{i\text{-}f}}{\Delta\omega}\right)^2} \tag{10}$$

where $\omega_{i\text{-}f}$ is the angular, mid-band intermediate frequency, and $\Delta\omega$ is the 3 decibels down, angular, input circuit band width. This approximation assumes adjustment for a flat-top response curve and a high secondary Q (i.e., the damping is provided entirely by the antenna). Under these conditions the primary Q determines the band width according to the relation

$$Q_1 = \omega C_1 R_a = \frac{1}{\sqrt{2}} \frac{\omega}{\Delta\omega}$$

Solving this for C_1 and putting it into the previous expression, it is seen that the radiated watts are

$$W = \frac{e_1^2}{R_a} \approx \frac{e_{LO}^2}{\sqrt{2}} \frac{\Delta\omega C_2}{1 + 2\left(\frac{\omega_{i\text{-}f}}{\Delta\omega}\right)^2} \tag{11}$$

which is conveniently independent of antenna radiation resistance.

Assuming an 8-megacycle circuit band width, which is suitable for the 6-megacycle channel of present U. S. transmissions, and an oscillator excitation of 2 volts, which is satisfactory for the high-transconductance pentodes, Table III was calculated for two currently available tubes.

Table III

Intermediate Frequency	Microwatts Radiated	
	6AC7	6AK5
10 Mc	680	340
20 Mc	210	105
30 Mc	96	48
100 Mc	9	5

Even with an intermediate frequency of 100 megacycles, the local-oscillator radiation is far from the 0.01 microwatts desired.

A grounded-grid triode radio frequency stage, such as the one recently described[10] using a 6J4 or 6J6 tube, requires a slightly different approach. A double-tuned circuit between the triode and the pentode mixer will give slightly less radiation than a single-tuned circuit. With the double-tuned circuit, the local oscillator voltage at the triode plate (a loading resistor for primary damping is assumed) is given by equation (10) above. This voltage reacts on the input antenna circuit only through the plate-cathode capacitance and through the plate resistance, if we may assume good grounding of the grid. Assuming an antenna directly connected to the cathode-grid input circuit, we find an antenna radiation around 30 decibels less than the figures in Table III. Although this is a substantial improvement, it may be insufficient unless a high intermediate frequency is chosen. A pentode radio frequency stage, on the other hand, will have much less output-to-input coupling and may have an additional pair of selective circuits in its input. This should allow such a radio frequency stage to reduce oscillator radiation sufficiently to give satisfactory performance even for lower intermediate frequencies. The use of a mixer with less inherent radiation, such as the cathode-coupled, double-triode mixer (see Figure 13 of reference 10), will decrease the isolation requirements on the radio frequency stage.

Still other arrangements which are possible make use of balanced triode or pentode mixers with the local oscillator driving the tubes in parallel, while the antenna signal drives them in push pull. These circuits must be carefully balanced to give effective reduction of radiation. If combined with a grounded-grid triode radio frequency stage, however, it may be possible to attain the desired 50 decibels or so of radiation reduction. Neutralization of the radiation from a single mixer tube is possible but again, to be effective, is achieved by a rather critical adjustment. Particular methods of operating balanced mixers will give conversion using an oscillator of half of normal frequency. This places the oscillator so far away from the normal signal channel that, when combined with the neutralization due to a balanced oscillator feed and unbalanced signal feed, adequate reduction of radiation may be achieved.

The one solution, which unfortunately cannot be proposed with those tubes which are commercially available on the open market, is the separation of oscillator and signal circuits of the mixer by the

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[10] G. C. Sziklai and A. C. Schroeder, "Cathode-Coupled Wide-Band Amplifier," *Proc. I. R. E.*, Vol. 33, pp. 701-709, October, 1945.

methods used in such low-frequency tubes as the 6L7.[11] The signal-to-noise ratio of this illustrative type of mixer has not been adequate for television service. Thus, if a radio frequency stage cannot be used, the only practicable remedy is the use of additional selective circuits between antenna and mixer, possibly with a rejection band at local oscillator frequency.

In all cases, local oscillator shielding should be employed to prevent direct chassis radiation.

IX. CONCLUSIONS

The tolerable amount of local oscillator interference, or other c-w interference, in a television picture is greater when the interfering frequency is at the high end of the picture band. In this case, the chief annoyance is loss of picture contrast which, for strong interference, can be very bad, even to the point of a negative picture. The transition between a slight loss in contrast and a completely washed-out picture is sufficiently sharp to make a choice of minimum interference level not too difficult. A value of signal-to-interference field strength ratio of 20 decibels may be considered a satisfactory minimum when the interference is near the upper end of the picture band.

On the basis of an interference 20 decibels below a desired carrier, and assuming channel assignments and choice of intermediate frequency so that an interference does take place with another channel, it is found necessary to reduce receiver radiation to 0.01 microwatts to protect the 500 microvolt per meter field strength contour with receivers 50 feet apart. Prewar receivers radiated 10^5 times as much as this and so were extremely unsatisfactory. A grounded-grid triode radio frequency stage is not a sufficient safeguard, though a carefully designed pentode radio frequency stage may be. Other alternatives lie in the use of balanced or radiation-neutralized mixers, or additional selectivity with an oscillator rejection circuit between antenna and mixer.

Although none of the suggested remedies lend themselves to a minimum-cost receiver design, an adequate television service will require substantial suppression of local oscillator radiation if frequency allocations are such as to make interference possible.

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[11] C. F. Nesslage, E. W. Herold and W. A. Harris, "A New Tube for Use in Superheterodyne Frequency Conversion Systems," *Proc. I. R. E.*, Vol. 24, pp. 207-218, February, 1936.

DEVELOPMENT OF AN ULTRA LOW LOSS TRANSMISSION LINE FOR TELEVISION*†

BY

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Summary—The development of a low loss 300-ohm parallel wire polyethylene dielectric transmission line is described. Loss curves, as well as a photograph of a production run sample of the line, are included.

INTRODUCTION

TRANSMISSION lines for use on home television receiver installations have been very unsatisfactory to date primarily because of their very high losses. In addition to the high signal attenuation these lines have had many other undesirable characteristics.

Twisted Pair Lines

Transmission line losses in twisted pair lines have been so great that installation men have recommended installing television receivers on top floors in buildings so as to shorten the transmission line and thereby improve the signal intensity by eliminating as much of the transmission line loss as possible. It was not uncommon to find a receiver installation within a mile of the transmitting antenna that did not have sufficient signal to override the local noise level. Two such installations recently investigated had transmission lines 600 and 400 feet in length and attenuations of 400 to 1 and 275 to 1 respectively. Each of these installations was less than a mile from the transmitting antenna and there was not enough signal to operate the receivers satisfactorily due to the long length of high loss transmission line used. Typical line losses varied from 5 to 12 decibels per 100 feet at 50 megacycles.

The manufacture of transmission lines has involved many different operations with a resultant high cost. Typical construction is as follows:

(a) Small gauge copper wire is tinned so as to prevent corrosion from the various compounds in the wire covering.

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* Decimal Classification: R320.41 × R583.

† Reprinted from *RCA REVIEW*, June, 1946.

(b) The small tinned copper conductors are twisted forming a flexible stranded conductor.

(c) The stranded conductor is rubber covered.

(d) Two of the stranded rubber covered conductors are twisted together.

(e) The twisted pair is rubber covered. (On the less expensive lines this operation is omitted.)

(f) A cotton braid is placed over the line to prevent the rubber from deteriorating in the sunlight.

(g) The line is given an impregnating dip in an asphaltic compound for weatherproofing.

Coaxial Lines

Coaxial cables and twin coaxial cables having medium loss characteristics have been available for some time. However, the cost of these lines has been so high as to prohibit their general use on home television receiver installations. One example is a twin coaxial cable having a loss of 1.4 decibels per 100 feet at 50 megacycles and selling for $1.25 per foot. The average home receiver installation requires 70 feet of transmission line, thus making a total cost of $87.50 for the transmission line alone. Therefore, practically all of the home receiver installations have been made with one of the twisted pair lines.

DEVELOPMENTAL SPECIFICATIONS OF LINE

The developmental problem was to produce a line that did not have the undesirable characteristics previously enumerated. The following developmental specifications were established.

(a) *Low Loss*—Loss should be less than any twisted pair of coaxial transmission line available in the pre-war period for the receiver installation. 1 decibel per hundred feet at 50 megacycles was set as the goal.

(b) *Low Cost*—The manufacturing cost of the line should be very low to permit the installation of both the receiver and antenna in the most desirable locations. A goal of six cents a foot list price was established.

(c) *Weather Resistance*—The average life of transmission lines used in the pre-war period was very low. In a few months the cotton braid failed and the sunlight hardened the rubber covering which cracked and permitted the absorption of moisture with a resultant increase in line losses. A minimum life of five years was desired for the new line.

(d) *Deterioration Due to Heat*—The asphaltic impregnating compounds used on the cotton braided lines softened and came off on the hands, clothing, woodwork, furniture and rugs in hot weather. The elimination of this undesirable characteristic was of primary importance.

(e) *Flexibility at Low Temperatures*—The asphaltic impregnating compounds used on the cotton braided lines hardened in cold weather and movement on the line during installation, or by the wind after installation, caused the line to crack and break the cotton fibers in the braid. This soon resulted in failure of the line insulation. The new line should be flexible under all temperature conditions.

DEVELOPMENT OF THE LINE

Work was begun to develop a transmission line meeting the specifications outlined above. A thorough check was made on all available transmission lines to determine what might be done to reduce their electrical loss.

With a parallel wire type of transmission line the loss varies inversely with impedance (See Figure 1). If such a transmission line has an impedance of 72 ohms and a loss of 6 decibels per hundred feet at 50 megacycles and the wires are separated far enough to produce an impedance of 144 ohms, the loss will decrease to 3 decibels per hundred feet at 50 megacycles or one half of its original value. It was thought that a high impedance transmission line of the conventional twisted pair type could be produced which would meet the specifications. However, it was found that the maximum improvement that could be obtained by increase in impedance was about 2:1.

(1) *Dielectric*

Along with the work on conventional transmission lines, one manufacturer developed a parallel wire transmission line with a spun glass woven web. This line had excellent low loss characteristics which met the loss specifications. However, the line was very hygroscopic and required impregnation. Eventually several good weather-proofing compounds were found which did not increase the line loss appreciably. The cost of this line however did not meet the tentative specifications.

Prior to the war, research work was done by this company on some relatively high impedance parallel wire transmission lines insulated with polystyrene. These lines had desirable electrical characteristics, but were unsatisfactory mechanically due to the brittleness of the polystyrene.

Polyethylene, suitable for transmission line insulation, was developed during the war for use at ultra-high-frequencies. This dielectric, while expensive, has very excellent electrical and mechanical properties. A full description of this material is beyond the scope of this paper and the reader is referred to one of the excellent papers on polyethylene.[1, 2, 3] Polyethylene has a power factor of approximately .0003 at frequencies as high as 1000 megacycles and a dielectric constant of about 2.29. It is a very strong, tough and flexible material that is not affected by acids, alkalis, ozone, sunlight and water. These properties

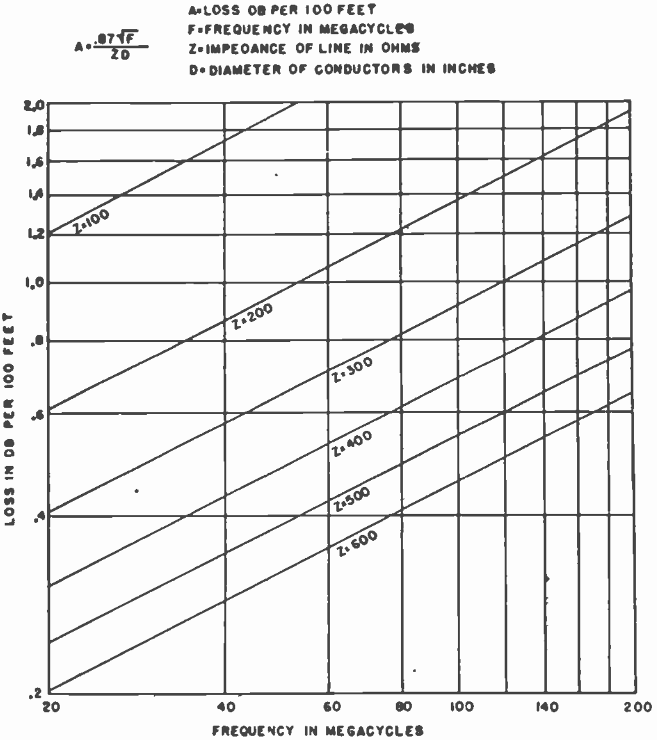

Fig. 1—Computed loss of open wire transmission lines using No. 20, A.W.G. wires.

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[1] M. C. Crafton, Jr. and N. B. Slade, "A New Dielectric For Cables" *Modern Plastics*, Vol. 21, No. 11, pp. 90-93, pp. 168-170, July, 1944.

[2] "Polyethylene Plastic—It Floats", *Modern Industry*, Vol. 9, No. 1, pp. 45 and 137-140, January 15 1945.

[3] "War Time Trends in Insulated Wire and Cable", Publication C-56, Anaconda Wire & Cable Company, 1944.

make it an outstanding dielectric for use at high- and ultra-high-frequencies. Several experimental lines were made of both twisted pair and parallel wire types. Measurements and field tests on these samples indicated a line could be constructed using polyethylene as the dielectric that would meet the specifications.

(2) *Line Conductors*

The conductor should be stranded to give the required flexibility and prevent breakage during use. The choice of conductor size is a compromise between mechanical strength, line loss and cost. Past experience on transmission lines proved that seven strands of No. 28 AWG would meet the structural requirements. A conductor of this size was also found to be satisfactory from the standpoint of line loss and cost. In the past transmission lines have had the copper conductors tinned because the bare copper wire was attacked by the various compounds in the insulating materials. Polyethylene is very inert and contains nothing that will react on copper, therefore bare copper conductors can be used. This is a fortunate condition because a lower loss line is obtained at reduced cost. At 100 megacycles the radio frequency currents are all on the outer surface of the conductors. As a matter of fact, the skin depth at this frequency is only .00067 inches. This means that with tinned wire most of the current is flowing in the tin surface layer, and since tin is a poorer conductor than copper the line loss is increased. The reduction in cost is an important item since the cost of the tin is saved as well as the expense of the tinning operation.

(3) *Line Impedance*

The development of the transmission line departs radically from past practices in connection with its surge impedance. Most of the transmission lines used in conjunction with television receivers have had surge impedances of between 70 and 125 ohms. A resonant dipole in free space has an impedance of approximately 72 ohms, therefore a 72-ohm line gives the best transfer of power when it is desired to receive signals on but one frequency. The problem of receiving television signals on a number of television bands is an entirely different problem. If a one-half wave dipole is designed to be resonant at 50 megacycles and used as an antenna, it will have an impedance of approximately 72 ohms at 50 megacycles and an impedance of about 2000 ohms at 100 megacycles. If a reflector is used in connection with this dipole the antenna's impedance can be as low as 200 ohms. Therefore the antenna's impedance may vary from some 20 ohms to 2000 ohms.

If a fixed antenna is to be used to cover a two-to-one or greater frequency range, it is desirable to use a line having an impedance such as to provide a maximum amount of energy over the desired frequency band. The line impedance would then be something less than one half the difference between the lowest and highest value of impedance, probably near 600 ohms. The actual value of optimum impedance is a complex affair and dependent upon many things such as the frequency response of the antenna, type of antenna load, impedance of antenna load, loss characteristics of the trasmission line, and other factors. It is sufficient to say here that the value of impedance would in all cases be many times higher than 72 ohms or the impedance of transmission lines used in the past. From an electrical viewpoint the line impedance should be high and probably between 300 to 600 ohms with the higher value of impedance favored, because the line loss is inversely proportional to its impedance for any given set of conditions.

There are, however, other considerations which have a bearing on line impedance. The line should be of such a size that standard hardware equipment can be used for the installation of the line. Standard bakelite screw eyes have a 9/16″ hole. This type of screw eye has been produced by various manufacturers for years and represents a standard transmission line support. The outside dimensions of the line should not be greater than 9/16″ if standard hardware equipment is to be used. With seven strands No. 28 AWG conductors this will limit the line impedance to a maximum value of about 400 ohms.

The amount of polyethylene used in a web line construction is approximately proportional to the square of the conductor spacing. If the web spacing is doubled the thickness of the web must also be doubled to maintain good mechanical design and have the web of sufficient thickness that the line cannot be crushed by the hands during installation and use. The use of a minimum amount of polyethylene favors lower line impedance.

The line impedance therefore should be approximately 300 to 400 ohms, with the cost favoring the lower value and the line losses favoring the higher value. A folded dipole antenna has an impedance of 288 ohms and was a deciding factor in choosing a line impedance of 300 ohms as the best value to give an ultra low loss transmission line for a minimum cost. The folded dipole is useful in receiving signals in a relatively narrow frequency band and provides a higher signal level than is obtained from simple wide band antennas.

The 300-ohm line used in connection with a half wave dipole gives a broad frequency response and permits multi-channel reception without cutting the antenna elements or changing their spacing as has been

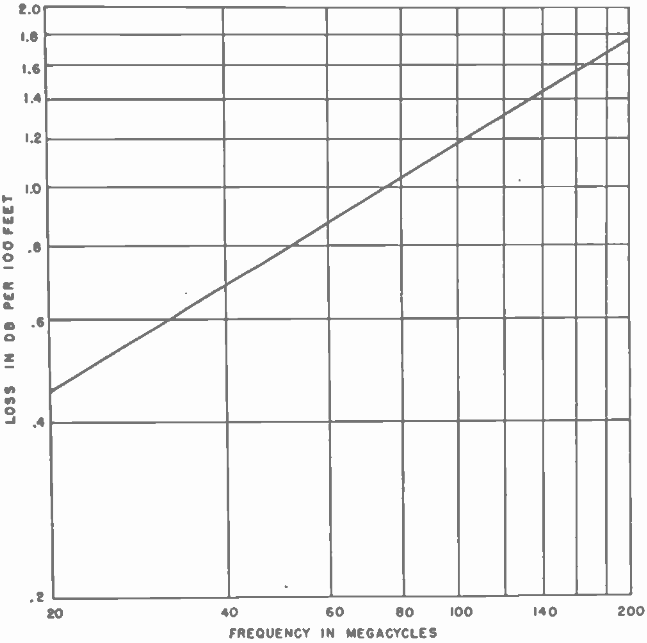

Fig. 2—Loss characteristics of the new television transmission line.

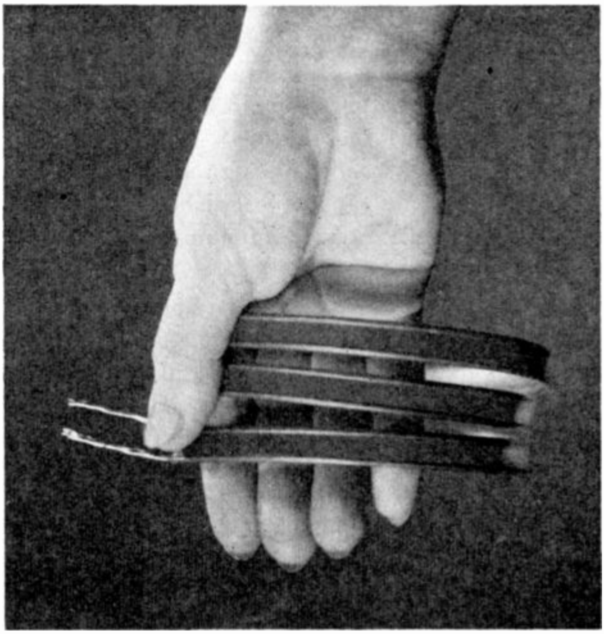

Fig. 3—Sample of production run transmission line.

required in the past. Figure 2 shows the loss characteristics; Figure 3 is a photograph of a sample of the production run transmission line. Figure 4 (Curve A) shows the relative response with frequency of a 44-megacycle half-wave dipole and reflector in conjunction with the 300-ohm line. Curves C1 to C6 give the relative response frequency characteristics of folded dipoles adjusted for each of the first six television channels and used with the 300-ohm line.

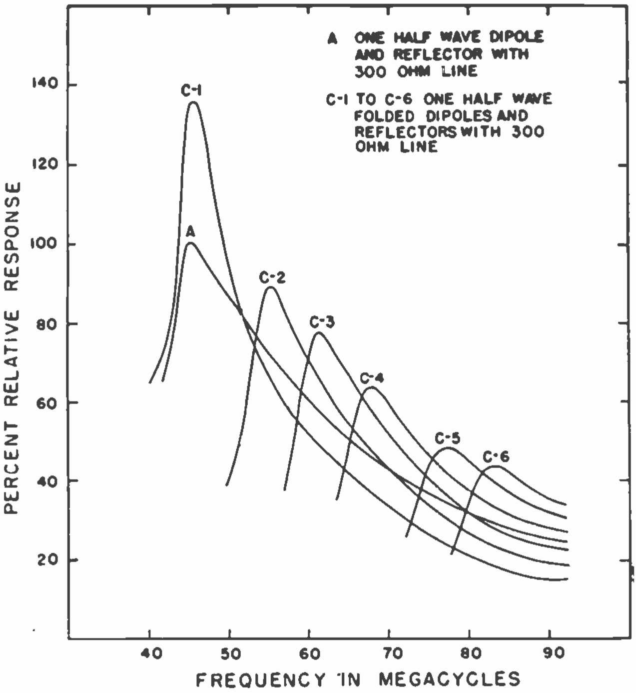

Fig. 4—Television antenna characteristics using the new 300 ohm transmission line.

Field Tests

A quantity of the 300-ohm transmission line was made on a developmental basis and installed, for field test purposes, in forty-seven test locations in the New York and Philadelphia areas prior to April, 1945. Some of the lines have been in service for over two years. Loss measurements on these lines show they have not changed by any measurable amount.

CONCLUSIONS

This developmental project resulted in an ultra low loss transmission line of unusual characteristics. It more than meets the specification requirements set forth. The line has a loss of less than 0.8 decibel per hundred feet at 50 megacycles. Polyethylene is a very strong, tough flexible material which is not affected by acids, alkalis, ozone, sunlight or water. This produces a line which does not crack during cold weather or soften during hot weather, and which give long trouble-free service. The line can be used with folded dipoles or dipole antennas giving high gain single channel or medium gain multi-channel reception respectively.

AN EXPERIMENTAL COLOR TELEVISION SYSTEM*†

BY

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Summary—*A description is given of color television apparatus using an image orthicon in the color camera for direct pickup of studio scenes. A sequential three-color semi-mechanical system is used. Provision is made for demonstration of color pictures in three dimensions. The associated sound channel is transmitted on the edge of the picture during a portion of the horizontal blanking period.*

INTRODUCTION

LATE in 1945 and early in 1946, a series of demonstrations were given of television in color and of television images in three dimensions, also in color. These demonstrations were conducted to show the status and to point out the problems remaining to be solved before color television could be considered ready for development as a service. The remaining problems are such as to require much additional research and development. However, this article is not concerned with these aspects of the situation but rather with a description of the system and apparatus used during the demonstrations.

Work on broadcast television was interrupted by the war, but advances in electronic and radio techniques during the war period did have a direct influence on television, particularly monochrome television. In order to resume the studies of color television and to evaluate these advances as they applied to television in color, laboratory facilities for research on the various problems involved in the generation, transmission and reception of television images in color together with new studio facilities, new circuits and apparatus were developed and put in operation.

CAMERA STUDIO SETTING

A small studio set was constructed in the laboratory to make possible small scale productions of colorful program material. The set is shown in Figure 1. Illumination is obtained from an overhead bank of 36 100-watt fluorescent lamps which provide an incident light of about 200 foot-candles. Two auxiliary banks of 24 100-watt lamps

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* Decimal Classification: R583

† Reprinted from *RCA REVIEW*, June, 1946.

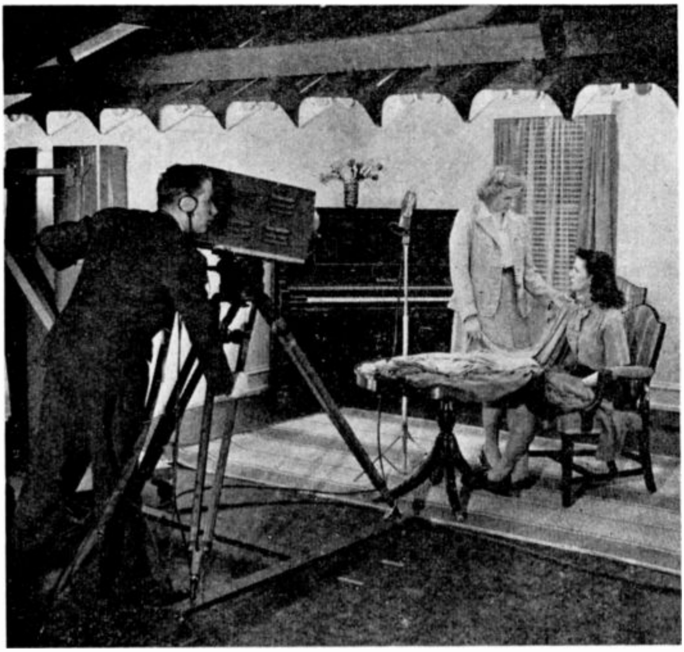

Fig. 1—Camera Studio Set.

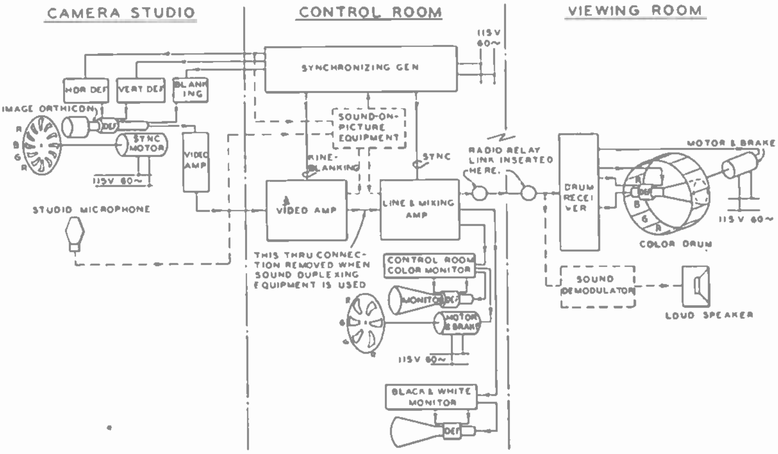

Fig. 2—Block Diagram of Color Television System.

each can be moved about as desired in front of the scene to bring the combined illumination up to more than 400 foot-candles. In order to obtain a more uniform light spectrum, half of the lamps are of the white type and half are of the daylight type. By distributing the lamps uniformly on the 3-phase 60-cycle power supply, no difficulty is experienced due to the lights operating on alternating current power.

CIRCUITS AND APPARATUS

A complete set of experimental equipment was constructed as shown in the block diagram of Figure 2. The system employs the latest and most suitable devices and circuits which have resulted from many years of extensive research in the field of electronic monochrome television. Added to these electronic components are two mechanically rotated tri-color filters so arranged that when the observer is viewing the picture on the kinescope through a red section of the filter in the receiver, for example, the pickup tube is being exposed to the televised scene through a red section of the filter in the camera. Similarly, when the blue and green filter sections, in turn, are in front of the kinescope, the blue and green sections are correspondingly in front of the pickup tube. The red, blue, and green images are repeated frequently enough so that the three are superimposed by the "persistence of vision" of the observer, to create the illusion of a single picture in multiple colors.[1,2,3,4,5]

The operating standards used are: 120 fields per second, 60 frames 2 to 1 interlaced, 525 lines, 40 single-color fields or 20 interlaced full color pictures per second. The color sequence is red, blue, green. With these operating standards, the resolution obtained with the overall system is about 250 lines.

The apparatus is designed so that by slight modification the transmission and reception of color pictures in three dimensions can be demonstrated. For this operating condition, polarizing light filters are incorporated with the rotating color filters at the camera and the kinescope. Special polaroid spectacles are provided for the observers to enable them to separate the right and left images.

During public demonstrations of the color equipment, when it was necessary to transmit the signal to a point several miles away by means of a microwave relay link, it was found convenient to transmit the asso-

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[1] J. H. Hammond, U. S. Patent No. 1,725,710.

[2] R. D. Kell, U. S. Patent No. 1,748,883.

[3] J. L. Baird, British Patent No. 473,323.

[4] P. C. Goldmark, J. N. Dyer, E. R. Piore and J. M. Hollywood, "Color Television, Part I", *Proc. I.R.E.*, Vol. 30, No. 4, pp. 162-182, April, 1942.

[5] CBS Engineers, "Color Television on Ultra High Frequencies", *Electronics*, Vol. 19, No. 4, pp. 109-115, April, 1946.

ciated sound on the same radio carrier as the picture by means of a time division duplexing circuit.[6]

CAMERA

One of the most outstanding new components incorporated in the color system is a special form of the image orthicon.[7] This camera tube is found to have sufficient sensitivity and a sufficiently uniform spectral response to make possible direct pickup of studio and outdoor scenes having an illumination level of from 150 to 300 foot-candles.

Equalization of the sensitivity of the orthicon to the three colors is accomplished by appropriately masking down the aperture on the filter disc for those colors to which the tube is most sensitive, thus reducing the time devoted to storing charges during those particular color fields. Computation of the degree of masking required is based upon measurements of photo-cathode current flowing when the rotating color disc is

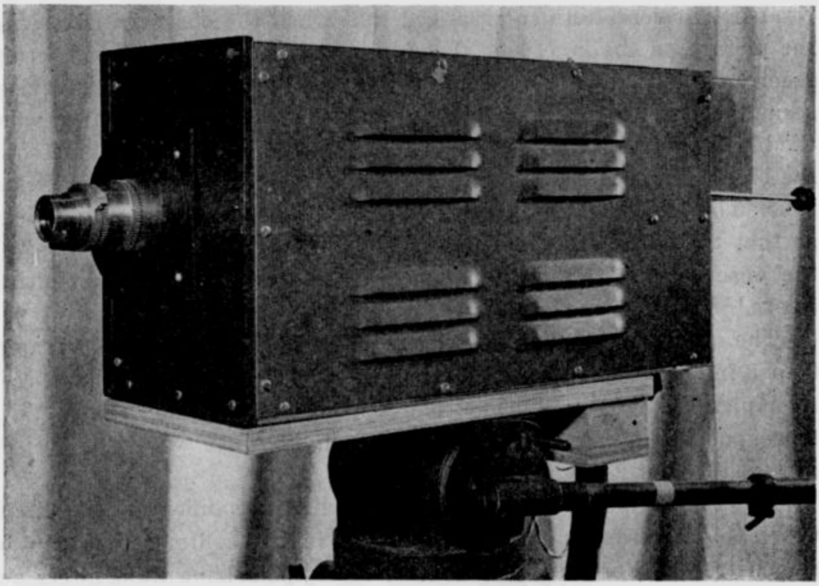

Fig. 3—Experimental Color Camera, Left Side.

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[6] G. L. Fredendall, Kurt Schlesinger, and A. C. Schroeder, "Transmission of Television Sound on the Picture Carrier", *Proc. I.R.E.*, Vol. 34, No. 2, pp. 49-61, Feb., 1946.

[7] Albert Rose and P. K. Weimer, "The Image Orthicon, a Sensitive Television Pickup Tube", presented at the I.R.E. Winter Technical Meeting on January 24, 1946 in New York, N. Y.

temporarily replaced by individual color filters. The subject for this test should be a white surface illuminated by the studio lights. After this first color balance has been obtained with a given camera tube, it is possible to operate the camera in much the same way as a conventional photographic camera. For televising scenes under illumination of different color temperature, the correction is made by the addition of a correcting filter over the lens.

The camera is used with either a 90 millimeter, f:3.5, or a 50 millimeter, f:1.9, Eastman Ektar lens. Both are color corrected. A lens aperture of about f:4.5 is required for the illumination present in the studio.

Side views of the camera are shown in Figures 3 and 4.

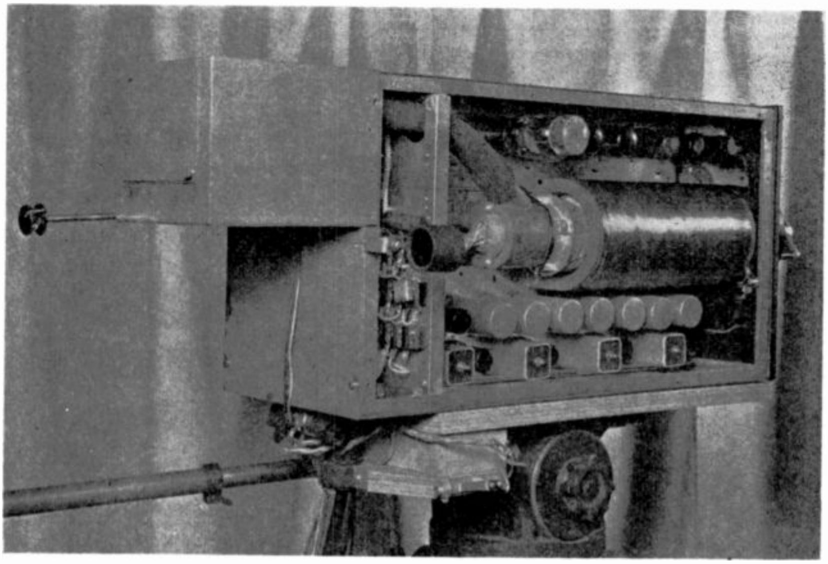

Fig. 4—Experimental Color Camera, Right Side.

The rotary color disc for the camera, Figure 5, is 6⅞ inches in diameter. It has twelve filter sectors clamped with the color balance mask between two plates of glass. The filters used are the conventional Wratten tri-color photographic filters, numbers 25 (red), 47 (blue), and 58 (green). The disc is rotated at 600 revolutions per minute by gearing from a small synchronous motor powered directly from the 60-cycle mains. The phase position of the motor with respect to vertical scanning is adjusted by manually rotating the motor frame. The disc is placed as close as possible to the face of the image orthicon in order to get it near the focal plane of the lens, thus minimizing any optical distortion that may be introduced. Since the image orthicon must operate in a uniform magnetic field of from 60 to 90 gauss, preferably extending beyond the image section of the tube, the focusing coil is made in two sections, with the smaller forward section being mounted in front of the color disc.

A multi-conductor cable connects the camera with the control room equipment which is mounted on racks of the conventional type. All electrical controls are on panels in the control room. The video amplifier and the deflection circuits are located inside the camera. All high voltage and plate supply units are located in the control room.

To overcome the difficulties in the camera and terminal equipment due to 60-cycle hum and crosstalk, special power supplies are required. The heaters of the various tubes in the camera are operated from the laboratory direct current power supply. The heaters of the tubes in the control racks are operated on 120-cycle power obtained by using selenium rectifiers across the output of special 60-cycle filament transformers. All plate voltage supplies are regulated and are operated from a 400-cycle source.

Fig. 5—Camera Color Filter Disc.

The video pre-amplifier used in the camera consists of five stages employing a combination of series and shunt peaking to obtain adequate bandwidth with sufficient gain to raise the signal level to approximately 1 volt peak-to-peak at the sending end of the camera cable.

CONTROL ROOM EQUIPMENT

A view of the control room equipment is shown in Figure 6. Here is located the synchronizing signal generator, the main video amplifier, the sync-mixing and output line amplifiers, the sound-on-picture terminal equipment, a black-and-white picture monitor, and the direct current power supplies with all camera controls. Control of the video signal amplitude is accomplished by a gain control circuit in the main video amplifier.

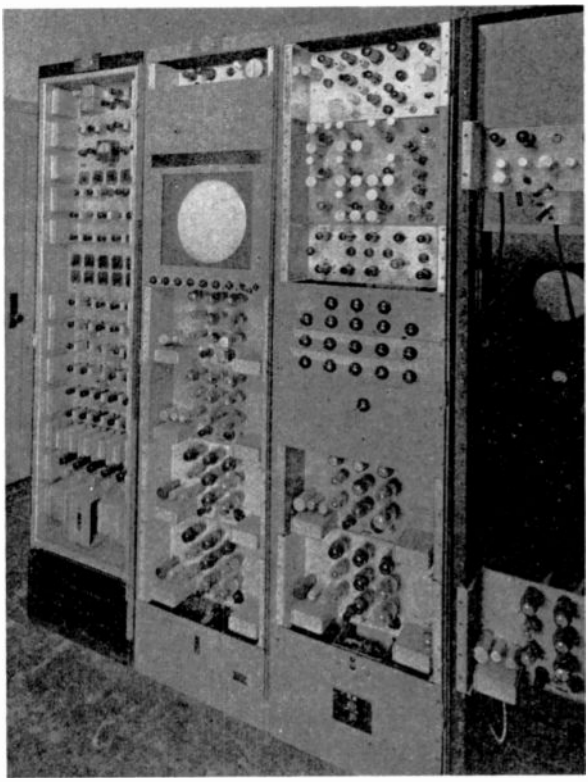

Fig. 6—View of Control Room Racks.

Both the synchronizing signal generator and the video amplifiers are standard black-and-white picture equipment modified to operate on color standards. All of the amplifiers used in the system are equalized to beyond 9 megacycles. This would make possible the same degree of horizontal resolution (at twice the scanning rate) ordinarily obtained with black-and-white standards using a 4.5 megacycle channel, provided the other limitations imposed by the color system did not exist.

CONTROL ROOM COLOR MONITOR

In addition to a standard 12-inch black-and-white monitor which is useful for checking camera focus, color phasing, scanning, etc., there is provided a color monitor using a 9-inch kinescope having an aluminized screen[8] and operating at a second anode voltage of 15 kilovolts. Front and rear views of this unit are shown in Figures 7 and 8. A 21¾″ diameter color disc is used in this receiver carrying six filter sections and rotating at a speed of 1200 revolutions per minute. Power to turn the disc is supplied from a ⅛ horsepower 60-cycle induction motor through a belt and pulley drive. This permits the motor with its disturbing magnetic fields to be located at a considerable distance from the cathode ray tube. Synchronization of the color disc speed with the color field repetition rate is accomplished by a magnetic brake. Proper color frame phasing is obtained by momentarily releasing the brake by manually switching off its controlling current.

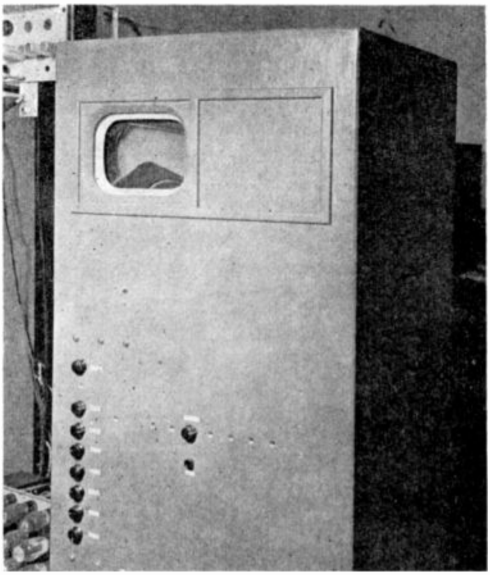

Fig. 7—Color Monitor, Front View.

DRUM COLOR RECEIVER

For best viewing by a large number of observers a demonstration

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[8] D. W. Epstein and L. Pensak, "Improved Cathode-Ray Tubes with Metal-Backed Luminescent Screens", *RCA REVIEW*, Vol. VII, No. 1, pp. 5-10, March, 1946.

receiver was built using a 12-inch, short persistence, aluminized-screen kinescope operating at a second anode potential of 17 kilovolts. Photographs of this receiver are shown in Figures 9 and 10.

The rotary color filter is in the form of a large drum, one end of which is open to allow the kinescope to be supported inside the drum and at right angles to the axis of the drum by a stationary bracket. The other end of the drum is closed to provide attachment to the drive shaft. The periphery of the drum consists of 12 rectangular red, blue, and green color filter sections clamped in a suitable framework. The drum is rotated at 600 revolutions per minute, in the direction of vertical scanning and in synchronism with the picture field repetition rate. The mechanical drive for the drum is similar to that used for the disc of the control room color monitor.

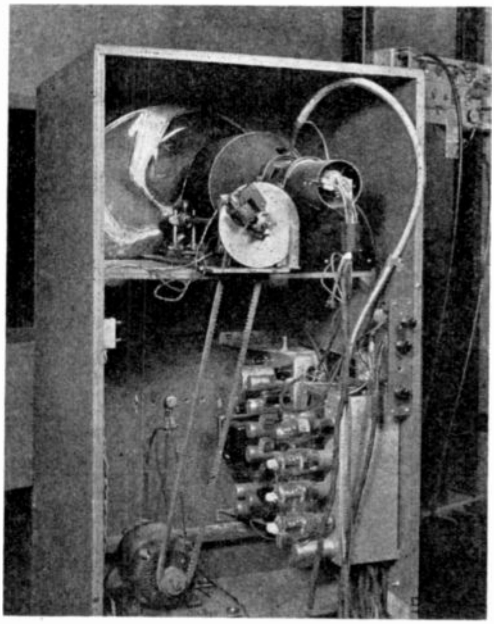

Fig. 8—Color Monitor, Rear View.

To overcome the difficulties due to 60-cycle interference in various components of the receiver, several precautions are taken. The kinescope is placed in a large mu-metal shield to protect it from the magnetic fields of the motor and power transformers. The power supply for the tube plates is obtained from a regulated source. All heater power is obtained from a full-wave selenium rectifier.

High voltage for the kinescope accelerating electrodes is obtained from a pulse power supply in which the high voltage pulses developed in an auxiliary winding on the horizontal deflection transformer during the "fly back" time are used in a voltage-quadrupling rectifier to obtain 17 kilovolts at a current of several hundred microamperes. With this high anode voltage and the advantage of the aluminized kinescope screen, a screen brightness of 4.5 foot lamberts is obtained in the highlights of the pictures. Since a light loss of approximately 90 per cent is introduced by the color filters, the actual kinescope brightness is 45 foot lamberts. At present a minimum of 10 foot lamberts is considered satisfactory for monochrome television.

Fig. 9—Drum Receiver, Front View.

TELEVISION PICTURES IN THREE DIMENSIONS

In natural stereoscopic vision the distance to any object (and hence the sense of depth) in the scene is determined by three different properties of the views seen by the eyes. The first is the difference between the two images resulting from the different points of view of the two eyes; the second is the focusing of the individual eyes; and the third is the amount of convergence or toe-in of the two eyes to see a given object in the scene.

In the stereoscopic television system described here, the images intended for the right and left eyes, respectively, are reproduced on the kinescope screen in time sequence. The two images are separated by polarizing the light from them in planes at right angles to one another by means of sheets of polaroid filter material associated with the color filters on the rotating drum of the receiver as described previously. The observer wears a pair of special polaroid glasses, in which the plane of polarization of each lens is set to agree with the plane of polarization of the picture intended for the corresponding eye.

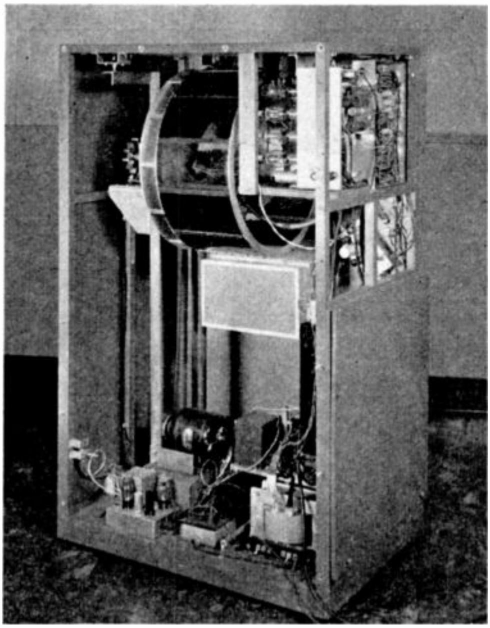

Fig. 10—Drum Receiver, Rear View.

At the camera, a light splitter is mounted in front of the lens. This attachment, shown in Figures 11 and 12, consists of a system of mirrors set at 45 degrees behind each of two windows which are spaced horizontally on centers 3½ inches apart. This spacing is a function of the normal interpupillary distance and the overall magnification of the system both optical and electrical. The factors are related by Rule[9] in the equation— $T = wed/sf$

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[9] John J. Rule, "The Geometry of Stereoscopic Projection", *Jour. Opt. Soc. Amer.*, Vol. 31, No. 4, pp. 325-334, April, 1941.

where w = width of image on photo-cathode of camera; s = width of image on viewing screen; T = lens separation of the camera; e = human interocular distance; f = focal length of camera; d = distance of camera lens to plane of object which is intended to appear coincident with the plane of the viewing screen. Sequential separation of the two images is achieved by means of polaroid filters which are placed over each of the windows so that the light coming from the scene as viewed through the left "eye" is horizontally polarized, while that through the right "eye" is vertically polarized. Selection of the particular image to be transmitted during a given field is made possible by means of additional polaroid filters mounted on the rotating color disc with their planes of transmission arranged in quadrature and in alternate fashion. Thus the vertically polarized image is transmitted during

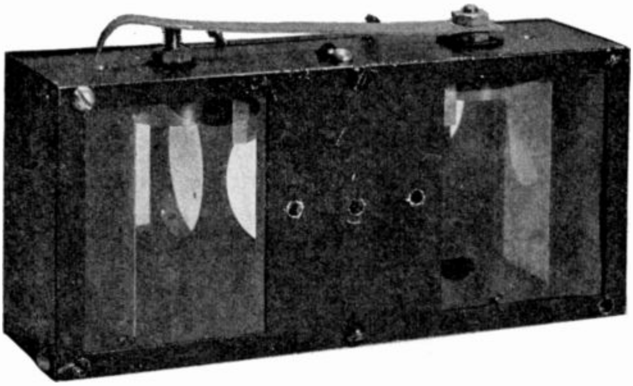

Fig. 11—Stereo Attachment for the Camera (Light Splitter).

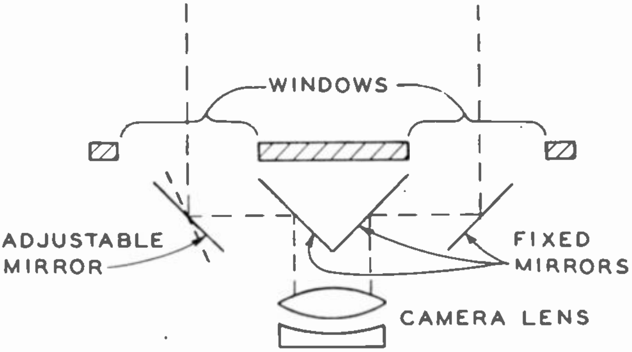

Fig. 12—Light Splitter.

one field and the horizontally polarized image during the next, the unwanted image being suppressed by crossed polarization.

The angular setting of one of the mirrors in the light splitter is adjustable so that the convergence of the camera "eyes" can be set to bring into register some object near the center of the useful depth of field. Observation of this point on the screen of the kinescope corresponds to the situation illustrated in Figure 13(A). In this case the object appears to be in the plane of the viewing screen since the angle of toe-in of the eyes is commensurate with the focal distance. The image of objects farther away from the camera will not fall at the same place on the screen, but will be separated horizontally a small amount, depending upon their position. Thus the horizontally polarized image intended for the left eye is displaced to the left as indicated in Figure 13(B). This condition gives the impression that the object is behind the screen as a result of the angle that the eyes must take in order to obtain fusion of the two images. Conversely, objects nearer to the camera are displaced horizontally in the opposite direction on the screen as shown in Figure 13(C) and thus appear to the observer to be located in front of the screen.

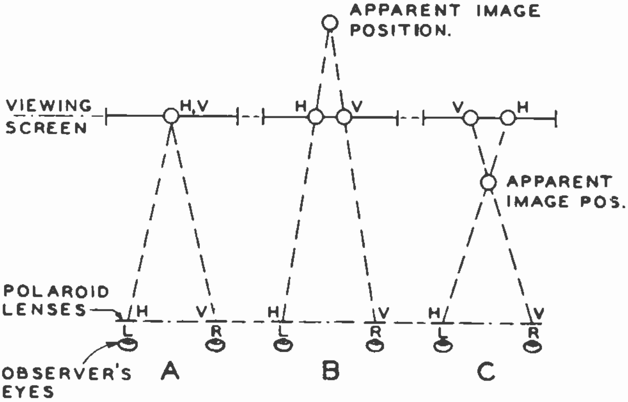

Fig. 13—The Geometry of Stereo Viewing.

The optical adjustments of the system can be made to give orthostereoscopic pictures only for one viewing distance and screen size. When more than one receiver is to be operated from a given camera, and if different amounts of magnification are to be used, some correction for the depth and perspective distortion can be obtained by keying the horizontal positioning circuit field by field thus changing the horizontal displacement of a given object on the screen, and hence its apparent position relative to the observer.

SOUND-ON-PICTURE EQUIPMENT

During certain public demonstrations of the color television apparatus the receiving equipment was in another building several miles from the studio. A microwave relay link operating on a frequency of 10,000 megacycles was employed on these occasions for transmission. The associated sound signal was transmitted on the picture carrier by means of a time division duplexing system.[6]

The essential elements of the sound-on-picture equipment are shown in dotted lines on the diagram of Figure 2. The sound modulator is inserted between the video amplifier and the output line and mixing amplifier and keys a rectangular pulse into the "back porch" of the horizontal blanking. This pulse is of constant amplitude extending down to white level and is adjustable in width in accordance with the audio modulation amplitude. In fact, the variable width modulation system is similar in many respects to the variable area sound track as commonly used for sound motion pictures.

Demodulation of the duplexed sound channel is accomplished at the receiver by means of a synchronized electronic switch which serves to exclude all but the sound carrier pulses. An amplitude limiter removes amplitude noise. This sound system has the basic theoretical limitation that the maximum audio frequency that can be transmitted is one-half the horizontal line scanning rate, or 15,750 cycles for the color system employed. A further minor reduction in the available bandwidth is due to the cutoff characteristic of the low-pass filter which is required to exclude frequencies above 15,750 cycles.

SIMULTANEOUS ALL-ELECTRONIC COLOR TELEVISION*†

A Progress Report

By

RCA Laboratories Division, Princeton, N. J.

***Summary**—This paper presents the latest progress report on **Color Television** and the first on the new simultaneous all-electronic system. Basic design and operating characteristics are reviewed. The apparatus for scanning color slides and color motion picture film together with the color television receivers are described.*

IN OCTOBER and November, 1946, the Radio Corporation of America gave several demonstrations of color television to press, industry and Government groups. These demonstrations constituted a progress report on the work done in color television, which follows the program announced at the time of earlier demonstrations in December, 1945.[1] In the current demonstration, important advances in color television were shown. The new system is *all-electronic,* having the potential flexibility inherent in electronic arrangements, and *simultaneous,* all three color images being transmitted continuously. This system has many operating and performance advantages and is compatible with the present black-and-white television. Since each of the three color channels employs the same standards as those now in use for black-and-white transmission, the green channel is suitable for monochrome presentation. Color television of this type can be introduced at any time it is made ready and can be operated interchangeably with black-and-white television; undesirable obsolescence is not created.

The recent demonstrations included television pictures in natural color scanned from kodachrome slides and from 16-millimeter color motion picture film. In order to demonstrate interchangeability, pictures in monochrome using signals of present black-and-white standards were shown on the color receivers; pictures in monochrome, using signals of the simultaneous color transmission, were then demonstrated on a current model black-and-white receiver.

Research work is under way and progress is being made in the radio transmission and reception of simultaneous all-electronic color

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* Decimal Classification: R583.

† Reprinted from *RCA REVIEW*, December, 1946.

[1] R. D. Kell, G. L. Fredendall, A. C. Schroeder, and R. C. Webb, "An Experimental Color Television System", *RCA REVIEW*, Vol. VII, No. 2, pp. 141-154, June, 1946.

television and in the building of television cameras for studio and outdoor pickup of this system. This work, together with propagation tests and field surveys, is a part of the over-all schedule yet to be fully worked out, but already well along.

Since simultaneous all-electronic color television is of far-reaching importance, the experimental equipment used during the recent demonstrations is described herein. This includes the apparatus for scanning color slides and color motion picture film together with the television receivers for color. Some of the basic design and operating characteristics are also reviewed. Figure 1 is a block diagram of the system.

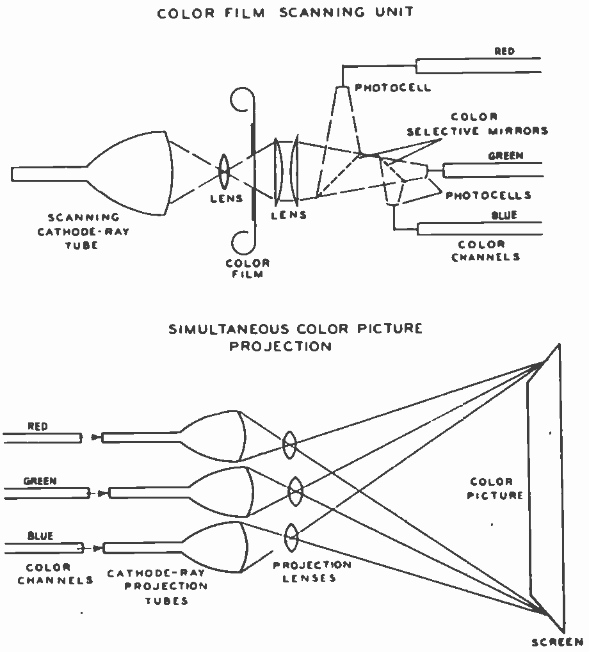

Fig. 1—Block diagram of the simultaneous all-electronic color television system. (Each of the color channels, shown at the right of the upper figure, have the same operating standards as the current black-and-white system. At the receiving end, shown at the left of the lower figure, each color channel is associated with its separate cathode-ray projection tube.)

Stationary Picture Signal Generator

One of the primary needs for the development of a simultaneous color television system is a standard source of tricolor video signals on which one may rely for good resolution, good registration, high signal-to-noise ratio, freedom from spurious signals, and good color fidelity. A special slide scanner utilizing a cathode-ray tube as a flying-spot scanner, a beam splitter, and three photoelectric tubes, were developed for this purpose.

A photograph of this apparatus with a superimposed phantom view of the kinescope and the light paths is shown in Figure 2. The raster formed on the screen of the kinescope is imaged on the slide by means of a lens. The light rays transmitted by the slide are condensed and then divided by dichroic mirrors which pass one color of light and reflect the other colors. The use of dichroic mirrors for a light splitter instead of half-silvered mirrors and color filters, reduces light losses and therefore provides a signal with higher signal-to-noise ratio. The divided light beams are further filtered by color absorption filters, then

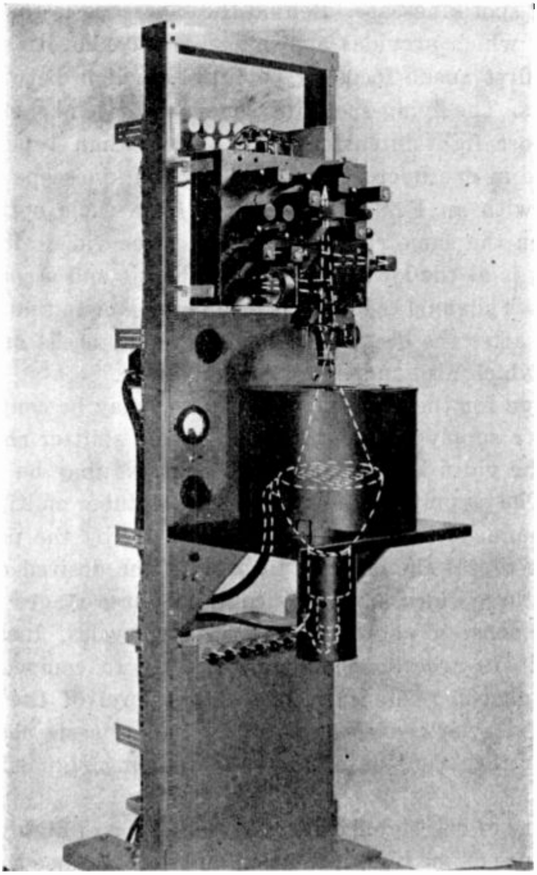

Fig. 2—Stationary picture signal generator.

collected by multiplier type phototubes which convert the varying light intensity of the spot as transmitted by the slide into video signals corresponding to the three primary colors of the slide. The use of multiplier phototubes provides a high video input to the amplifier. The amplifiers are equalized to correct for the decay characteristic of the phosphor used in the flying-spot kinescope.

On the bottom of the rack in the photograph (Figure 2) is the chassis containing the synchronizing, blanking, and deflection circuits for the flying-spot kinescope. Behind the kinescope is the high-voltage power supply which provides approximately 30 kilovolts for the kinescope and a first anode focusing potential variable between four and seven kilovolts. The flying-spot kinescope has a special short persistence phosphor whose light intensity drops to less than 1 per cent of its original value in one microsecond. Facing the kinescope screen is the slide holder with an F:2 objective lens. The whole optical assembly is mounted on the same chassis with the three video amplifiers. The beam splitter is at the lower end of this chassis and a condensing lens system for each channel reduces the beam diameter to that of the photocell aperture after the beam is divided. The photocells are enclosed in shield cans, which also support the color filters.

The voltage for the phototube multipliers may be controlled by the variable power supply directly under the beam splitter chassis, and by this means the video levels of all three channels may be varied simultaneously. The supply voltage of the phototube multipliers of the individual channels may be varied individually by the potentiometers visible at the top of the chassis, to provide the desired color balance. Each of the three video amplifiers contains three stages having a flat frequency response to approximately 5.5 megacycles. Included in each of the amplifiers are the equalizing circuits to compensate for the various phosphor persistences. The output level of the amplifiers is approximately 1 volt peak-to-peak. The small chassis above the beam splitter is for the insertion of the synchronizing signals in the green video signals.

The quality of the signal from this generator is highly satisfactory not only because of the high resolution, but also because the blacks have the unusual characteristic of being practically free from noise. Noise in the picture therefore has the general appearance of the equivalent effect found in motion pictures from photographic grain and dirt. The registration of the three signals is inherently correct.

MOTION PICTURE FILM SCANNER

The motion picture film scanner was built with no attempt at refinement or optimum design, in order to hasten preliminary tests of reception with moving subjects. Its general scheme is the same as that of the slide scanner, with the film gate replacing the slide holder. A photograph of the apparatus is shown in Figure 3. A standard 16-millimeter home sound film projector was modified by substituting a synchronous motor drive so that the film speed was changed to 30 frames per second (instead of 24). Each frame was then scanned twice to give 60 fields per second. The pull-down mechanism, which was unchanged, is so slow that it was necessary to blank approximately 30 per cent of the field time to avoid showing the distorted picture produced during the film pull-down time. The picture therefore actually contained only about 370 lines, although the nominal number of lines was retained at 525.

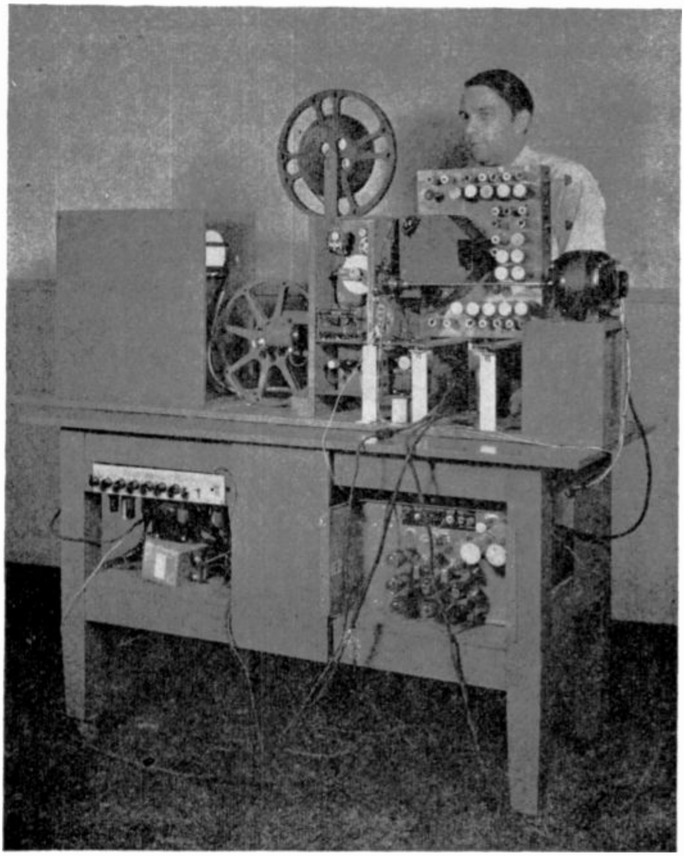

Fig. 3—Motion picture film scanner.

The picture quality was judged to be good, particularly when allowance is made for the fact that part of the picture area was missing, due to the compromise in design of the film projector. The sound was usable, but was not very satisfactory due to the improper film speed.

REPRODUCING EQUIPMENT

The picture reproducer, as shown in Figure 4, 5, 6, contains three three-inch kinescopes arranged side-by-side in an equilateral-triangular group, each having an associated projection lens and deflection yoke. The kinescopes are identical except that phosphors selected for producing red, green, and blue light, respectively, are used. The kinescopes and lenses are mounted in an assembly frame which also holds the yokes in such a manner that each yoke may be adjusted in rotation and height without disturbing the kinescope mounting position. Each

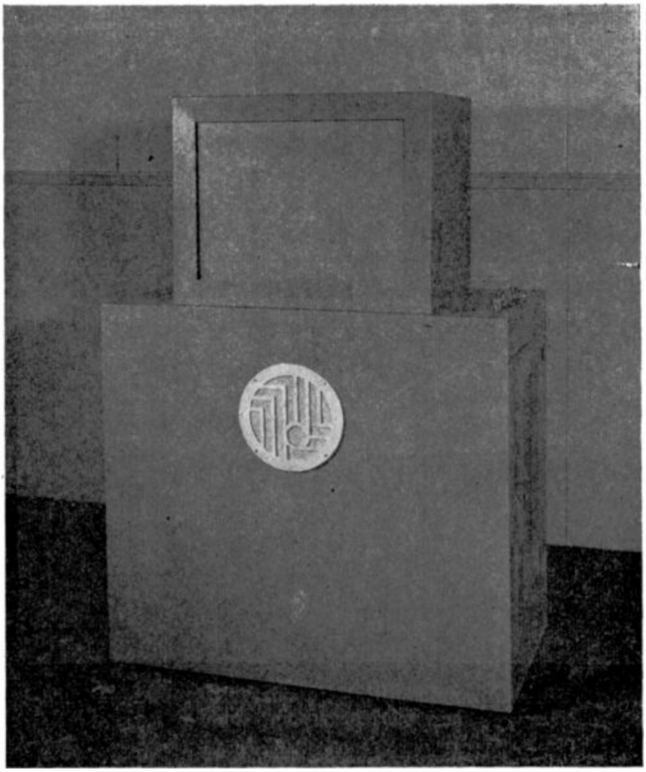

Fig. 4—Laboratory model simultaneous all-electronic color television receiver.

kinescope is provided with the video signal corresponding to its particular primary color, and has a scanning raster which produces light for its primary color image in the completed picture. The lenses project these three pictures simultaneously to the translucent viewing screen by way of a 45-degree mirror, as shown by dotted lines drawn on the photograph (Figure 6). The kinescopes are operated at a second anode potential of 25 kilovolts.

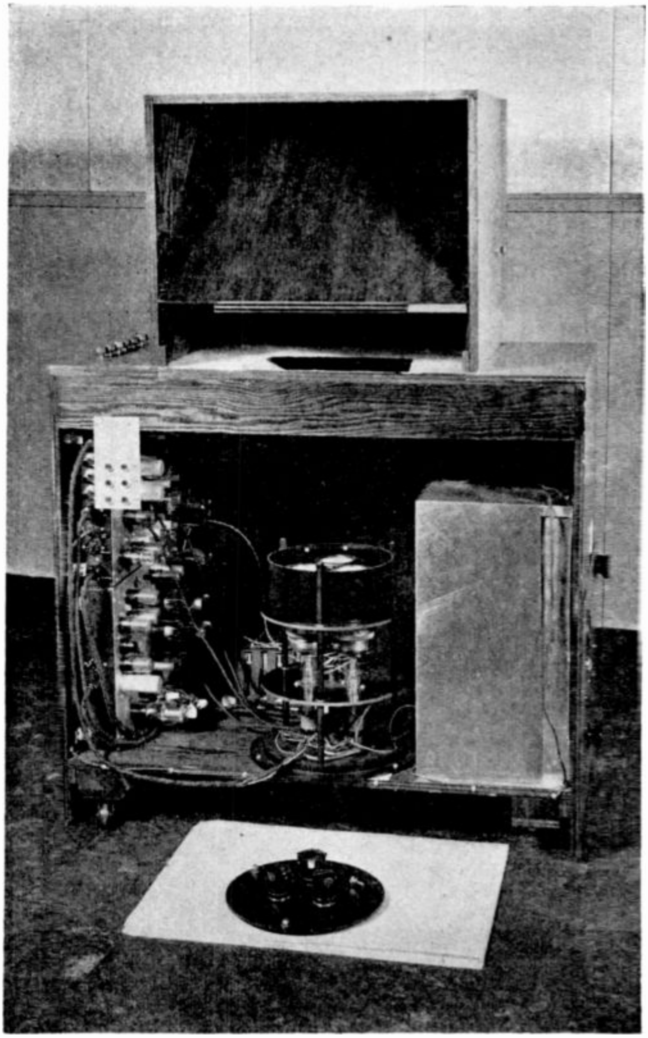

Fig. 5—Receiver with three cathode-ray projection tubes, lenses removed.

The optical system serves to focus and combine the three pictures on the translucent screen. In so doing, the images must not differ from one another in geometric distortion or location. Prevention of such difference is accomplished by placing the three kinescope faces in the same plane and mounting the three lenses above this plane, with their axes perpendicular to it. The axis of each lens is offset from the center of the kinescope face toward the center of the assembly by an amount sufficient to bring the three pictures into approximate register on the

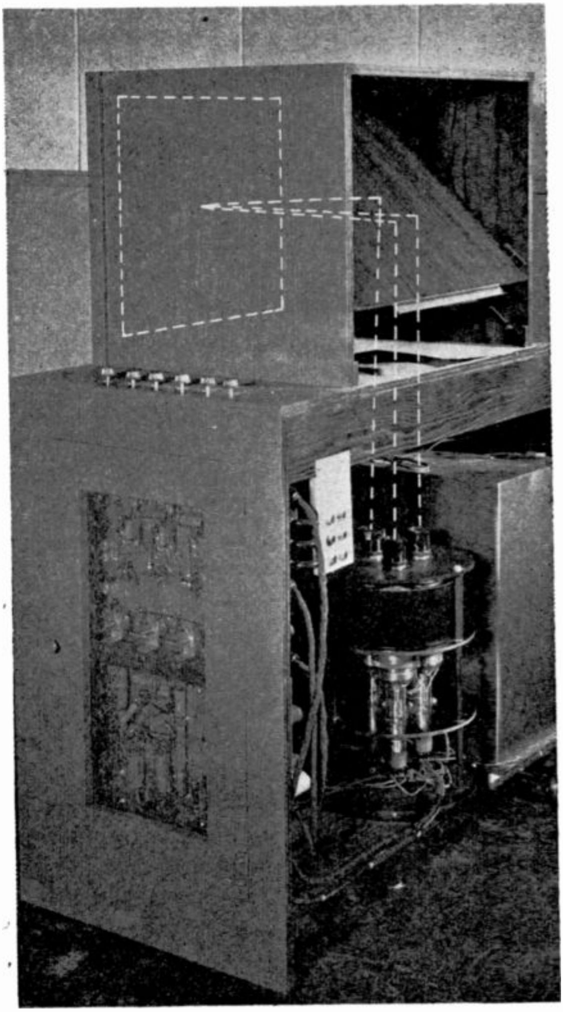

Fig. 6—Receiver with lenses in position, showing projection paths.

screen. Exact registry is then obtained by moving the raster on the kinescope face electrically. This offset, which is similar to a rising front on a photographic camera, causes no distortion, but requires extra covering power in the lens. The lenses used are F:2 projection lenses. They are threaded into the lens plate at the calculated positions and the threads serve as the focus adjustment.

The registration requirements are similar to those existing in color printing and color photography. Ideally, the three rasters should be identical and properly positioned within a fraction of the width of a scanning line. Practically, a considerable amount of misregistration may be present without being objectionable.

The scanning rasters are made substantially identical by using three similar yokes, and supplying them with power from the same deflecting circuit. The three yokes are connected in parallel rather than in series. This permits a simple individual centering or positioning arrangement and also insures more nearly identical deflection fields. In a series arrangement, one yoke would operate at a higher alternating-current potential with respect to ground, and would thus be shunted to a greater extent by the stray capacitances.

The positioning or centering arrangements are the usual television centering circuits, the only requirements being that the centering supply voltage must be stable, and that adjustment of the potentiometers must not alter the current waveshape through one yoke with respect to that through the others. To insure the latter, the horizontal centering potentiometers are by-passed very lightly and the vertical ones are unby-passed. Enormous capacities would be required to by-pass the vertical centering potentiometer properly. However, since the vertical circuit is essentially resistive, the addition of more resistance would simply change the amplitude, which is easily corrected.

The procedure used in registering the kinescope assembly is as follows. The kinescopes are adjusted for the proper height and clamped, then the lenses are focused optically. The yokes are then rotated until the edges of the three rasters are parallel. One of the three rasters is then considered standard for horizontal size, and one of the others is adjusted to it by moving the deflecting yoke up or down. Although this, of course, varies the raster size both horizontally and vertically, only the horizontal size is considered during this adjustment. This adjustment is then repeated on the remaining raster until the three horizontal sizes are alike. The vertical sizes are then adjusted by varying the value of small resistors in series with the vertical deflecting coils. The positioning or centering controls are then set to register properly the three rasters which are now of the same size.

Successful registry of the three rasters has been greatly facilitated by the use of aluminized kinescope tubes.[2] The aluminum film insures that the phosphor screen and the glass wall are at second-anode potential and hence do not collect charges that will divert the electron beam erratically.

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[2] D. W. Epstein and L. Pensak, "Improved Cathode-Ray Tubes With Metal-Backed Luminescent Screens", *RCA REVIEW*, Vol. VII, No. 1, pp. 5-9, March, 1946.

MILITARY TELEVISION*†

BY

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Summary—A general discussion and short history of television for military uses are presented, followed by a brief description of military television developments of World War II. Operational and other uses of these developments are discussed. The future of television for military and other purposes is considered in the concluding portion of the paper.

A NUMBER of excellent military developments completed during World War II saw only limited combat service or were never introduced into action at all. The reasons underlying such limited employment furnish ample material for an exhaustive treatise, but it is sufficient here to state that, whatever the reasons, television for military use was one such outstanding development.

Why, then, spend time reviewing and studying the subject? There are several valid reasons for so doing. First, in one sense, military television did see extensive war service in the forms of radar, radar countermeasures, loran, altimeters, shoran, advanced communication aids, and other devices whose war records need no discussion. Practical television resulted from two major developments which were of extreme importance because they not only made television possible but also laid the foundations for other important war services—such as radar. Because of the greater immediate importance of radar, it received more intensive development and much wider use in World War II than did its parent, television. The two major developments which made possible television (and radar, shoran, loran, etc.) were: (1) improved cathode-ray tubes for converting light waves to electrical signals and for reconverting these signals into visible light; and (2) electrical circuits and components associated with cathode-ray tubes capable of controlling, timing and utilizing these tubes in an extremely precise manner. The new tubes and circuits provided revolutionary means of effecting time control and measurement with a previously-unknown accuracy. In furnishing the basic ground work for such tubes

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* Decimal Classification: R583 × R560.

† Detailed technical papers on military television equipment are included herein following this introductory paper.

‡ Lieutenant Commander, United States Navy (Retired).

and circuits, television saw war service in many forms and consequently provides material which merits close study.

A second valid reason for examining military television is found in the latest commercial television equipment and techniques. Television's war service is now being repaid; war experience has given added impetus to commercial television and this, in turn, is responsible for the much-improved cameras, receivers, relay-links and tubes which today are being incorporated into a growing domestic television service. A background knowledge of war results and a detailed study of military developments are important for understanding and ability to work with advanced commercial television equipment.

There is another reason for reviewing wartime television developments. To the difficulties attending the introduction of *any* new military development, another may be added in the case of television for military uses. Pre-war commercial television equipment required complete and radical redesign, and operation of television equipment in planes and missiles raised some of the most troublesome engineering problems of the war.[1] A study of these problems and a thorough understanding of their implications is a prerequisite for advanced television development.

Finally, where military television did see action, it proved beyond question that here was still another example of the effectiveness of applied modern electronics. The future role of military television is now clearer and more easily understood; heretofore unthought-of military applications have become apparent with passing time and coincidentally with the development of other new military devices. Concurrently, it has become possible to perceive new and startling peacetime uses for television as a direct result of the experience gained in war.

It is well, then, to review very briefly the history of military television[2] and mention the operational and other uses to which it was assigned.

In 1934, Dr. V. K. Zworykin first advanced a concrete proposal for using television for military purposes to substitute for human eyes where it was not advisable for men to go. This first system is described

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[1] These problems are fully discussed in other papers:

Charles J. Marshall and Leonard Katz, "Television Equipment for Guided Missiles", *Proc. I.R.E.*, Vol. 34, No. 6, pp. 375-401, June, 1946.

M. A. Trainer and W. J. Poch, "Television Equipment for Aircraft", *RCA REVIEW*, Vol. VII, No. 4, pp. 469-502, December, 1946.

[2] A more detailed history of airborne television is included in another paper: Henry E. Rhea, "Airborne Television", *Broadcast News*, No. 43, pp. 24-41, June, 1946.

in detail in a memorandum by Dr. Zworykin.[3] In 1935, work began on a lightweight airborne reconnaissance television system which was successfully demonstrated in 1937. Two years later intensive development work was commenced on still smaller and lighter equipment. This new equipment was flight tested in 1941. The next three years were devoted to overcoming a multitude of technical difficulties attending the installation of the new television equipment in aircraft and guided missiles,[1] and in the design of the missiles themselves. Finally, in 1944 military television came into actual combat use.

Three basic military television systems were developed during World War II. They are known by the code names BLOCK, MIMO, and RING. These systems are described in detail in other papers[4]; only a general discussion of each is included herewith.

The BLOCK system consists of several designs built to operate on different frequencies. This equipment, manufactured in quantity, was used by both the Army and Navy—by the Army in the GB-4 radio-controlled glide bomb and in old "war-weary" B-17's, and by the Navy in TDR-1 drones and in the GLOMB radio-controlled glider bomb. BLOCK equipment, employing the iconoscope or image orthicon, is lightweight and compact—the camera unit weighing but 33 pounds and the transmitter unit 26 pounds. The entire weight of all television equipment in the drone or bomb is 100 pounds. It was designed to operate unattended and is, in the main, expendable. This equipment is also used, however, for certain other non-expendable applications.

The MIMO system is similar in most respects to BLOCK. It is, however, lighter and more compact, employing a new developmental MIMO-miniature image orthicon. The entire system, mounted in the Army ROC high-angle radio-controlled bomb, for which it was specifically designed, weighs but 50 pounds.

The RING equipment provides a more elaborate, high-resolution, airborne television system for reconnaissance. It was designed for attended operation with two or more cameras and is not considered to be expendable.

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[3] V. K. Zworykin, "Flying Torpedo with an Electric Eye", *RCA REVIEW*, Vol. VII, No. 3, pp. 293-302, September, 1946.

[4] M. A. Trainer and W. J. Poch, "Television Equipment for Aircraft", *RCA REVIEW*, Vol. VII, No. 4, pp. 469-502, December, 1946.

R. D. Kell and G. C. Sziklai, "Miniature Airborne Television Equipment", *RCA REVIEW*, Vol. VII, No. 3, pp. 338-357, September, 1946.

Paul K. Weimer, Harold B. Law and Stanley V. Forgue, "MIMO-Miniature Image Orthicon", *RCA REVIEW*, Vol. VII, No. 3, pp. 358-366, September, 1946.

R. E. Shelby, F. J. Somers, and L. R. Moffett, "Naval Airborne Television Reconnaissance System", *RCA REVIEW*, Vol. VII, No. 3, pp. 303-337, Sept., 1946.

BLOCK and MIMO transmitted television pictures to the controlling planes, which pictures gave the necessary control information. RING transmitted the television reconnaissance pictures to receivers at the base, in ships, or in other planes. All three systems provided information which fulfilled their design requirements. In the case of the systems used in drones or bombs, the detail in the picture improved as the need for more accurate control increased—i.e. as the drone or bomb neared the target—and was adequate for control purposes. In the case of the reconnaissance system, the detail, of course, varied with altitude and visibility conditions but was generally sufficient to distinguish all important details of the terrain under observation and to differentiate, for example, between the types of various motor vehicles and parked aircraft. The distance a RING-equipped plane could transmit clearly to its base varied between 100 and 200 miles, with values of over 200 miles having been recorded.

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In August, 1944, the Navy used television-equipped TDR-1 drones against Japanese shipping in the Northern Solomons. A few months later, a radar-equipped lighthouse at Rabaul Harbor was destroyed by the same means after having successfully withstood repeated bombing attacks of the ordinary type. Previous to these events, in the summer of 1943, the Army began using old "war-weary" B-17's, television-equipped and radio-controlled, to destroy submarine pens in Helgoland. Later, GB-4 glide bombs equipped with television were sent against the V-1 and V-2 launching sites on the French coast.

BLOCK equipment was also installed in reconnaissance planes after the capture of the Philippines and used for patrol work and battle damage survey.

Within this country, military television equipment played its part in the Manhattan project and in other wartime manufacturing processes which required constant surveillance of operations too dangerous to be approached by men themselves. Military television also participated in Operations Crossroad, where modified BLOCK equipment was used for telemetering and also for viewing the blasts from towers on Bikini Atoll and from drones flying through and around the atomic clouds—providing a close-range picture of these significant events to observers who otherwise would have had to be contented with film records and long-range views.

* * * * *

All people hope that wartime developments have seen their last combat action. It does no harm, however, to consider how this electronic weapon can help in our nation's defense should ever there again be cause to do so.

Military television will find continued usefulness in guided missiles. In fact, it is probable that most radio-controlled weapons of the future will include television equipment to make it unnecessary for the controlling organization to have either the missile or the target in direct view. This is obviously a definite requirement in the case of very-high altitude bombs or stratosphere rockets, or for ground-control of any type of guided missile.

The range of daylight reconnaissance television equipment will be extended materially to provide detailed information on distant points as readily as it now does on fairly close-range targets. The ultimate extension of this system is interesting and has already been the subject of some conjecture. The central war rooms of the future will not have ordinary walls; they will have huge television screens. The personnel within will be able to see their actual surroundings in all directions, with various range segments brought on the screens at will, and with closeup studies available on any section as desired.

Command posts behind battle lines will see their entire sector as one continuous panorama, the television signals coming from the "walkie-lookie" transmitters in the foremost positions, and from reconnaissance planes overhead. Officers controlling tank units will have similar panoramic television information from leading tanks and planes. Artillery observers will no longer receive "spots"; they will see the shots landing with relation to the target and can make their own corrections—an extension of the current high-resolution radars.

The flag bridge of ships will be smaller models of the larger land-based war rooms, with the entire fleet spread out before the Admirals' eyes. Enemy forces can be viewed as a whole or in part, either separate from or dubbed into their true location with reference to land positions, as in the case of contested amphibious operations. Another use for television aboard ships will be in maintaining a display of CIC and other data boards in gunnery plot, flag plot, on the bridge, and in other important stations.

Night will not necessarily stop military television's activities of the future. Tubes, sensitive to infrared (as are certain image orthicons today) will make it possible to view scenes so illuminated almost as if by daylight.

* * * * *

The principles of military television and the experience gained in tests and combat operations are destined to play a large part in the development of systems for air navigation, such as Teleran. This same experience and these same principles will also greatly expedite the use of television in connection with personal communication, air coverage of news events, control of laboratory and production processes, air supervision at scenes of disasters, etc., use by police in traffic and crime control, and the development of "walkie-lookie" television. As previously pointed out, many wartime television developments have *already* been incorporated into postwar commercial equipment.

Eventually, means will probably be devised which will overcome, through the use of radar techniques, the restrictions that are currently imposed on television by weather and visibility conditions or the short-range of infrared illumination. Until then, and after this occurs, military television will be a part of all operations—an extension of human vision. Even in this *enlightened* age, humans, with the limitations of their eyes, will have to fight the wars, if wars ***must*** be fought, and this new electronic weapon will assist them.

INTRODUCTION

to

TECHNICAL PAPERS ON AIRBORNE TELEVISION*†

THIS issue of RCA REVIEW contains the first of a series of technical articles on **airborne television** — a system of sight transmission having momentous military and civilian applications. Prepared and written by scientists and engineers of Radio Corporation of America, they are presented to readers of *RCA REVIEW* as an historic record of pioneering and scientific progress.

The idea behind airborne television and its development originated in RCA more than twelve years ago. It was in the spring of 1934 that Dr. V. K. Zworykin formulated plans and submitted to me a memorandum suggesting the creation of such a system to serve as "electronic eyes" in guiding radio-controlled aerial torpedoes. At that early date, Dr. Zworykin foresaw the threat of Japan's "Kamikaze" or Suicide Corps, and sought to achieve by technological means what the Japanese hoped to attain by psychological training. I was so impressed that, accompanied by Dr. Zworykin, I went to Washington and presented his plans to the War and Navy Departments. Some time elapsed before the armed services became actively interested in airborne television, but our scientists, meanwhile, continued to experiment and pioneer with this revolutionary method of extending human sight. First Ray D. Kell and Waldemar Poch developed light-weight cameras and associated equipment. Then Henry Kozanouski joined in and produced research equipment which was field tested in an airplane. When the war emergency arrived, the entire organization was ready to meet the challenge.

Three airborne television systems — designated "Block", "Ring" and "Mimo" projects as security pseudonyms — evolved for secret wartime purposes. Television pick-up and transmitting equipment that once might have filled a large room was redesigned, modified and built to "suitcase" compactness for military uses in the Block system, which was employed effectively in the war by both the Army and Navy. The heavier, longer-range Ring system was developed during the final stages of the conflict by engineers of the National Broadcasting Company, Inc., in conjunction with the U. S. Navy. The Mimo equipment was the midget of the three systems, being even smaller than the Block

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* Decimal Classification: R583 × R560.

† Reprinted from *RCA REVIEW*, September, 1946.

apparatus. It was developed primarily for use in guided missiles where space was insufficient to accommodate the Block equipment.

The articles in *RCA REVIEW* will trace the development of the three airborne television systems and relate in technical terms how the special requirements were met for equipment that would operate satisfactorily under the unusual handicaps of aerial warfare. These reports will tell of the design of small antennas practicable for airplanes, the use of the airplane's power supply, and the overcoming of the problems of noise and vibration. Special emphasis will be given to the development of the now celebrated image orthicon tube. Accounts likewise will be printed of technical aspects in the development of other vital electronic tubes and equipment. The full text of Dr. Zworykin's original memorandum of 1934 is published in this issue.

Great praise is due the scientists and engineers whose research and pioneering, technical knowledge and ingenuity made possible airborne television as a successful weapon of war and opened the way for monumental progress in widening television's scope of service in peace.

David Sarnoff

DAVID SARNOFF, *President*
Radio Corporation of America

FLYING TORPEDO WITH AN ELECTRIC EYE*†

By

V. K. Zworykin

Director of Electronic Research Laboratory, RCA Laboratories Division
Princeton, N. J.

A **memorandum** sent to DAVID SARNOFF, *President, Radio Corporation of America*, on **April 25, 1934.**

Summary—This paper, written in April, 1934, presents a detailed suggestion for the control of guided missiles using information obtained by television. Shortcomings of previous systems of guided missile control are briefly mentioned and a general description is given of the television apparatus for use in the new method of control. Approximate weight composition of such a television-controlled aerial bomb or torpedo is included. The suggestion envisions that the torpedo or bomb (or standard airplane) should be equipped both with automatic pilot control and remote radio control with the instrument and target data supplied by an iconoscope camera and transmitter in the piloted weapon.

Television information furnished would be of two kinds, and would be given simultaneously: (1) an actual view of the target which could be sighted upon by means of crosshairs; (2) accurate information on the readings of instruments in the piloted weapon, given by the position of bright spots on the edges of the picture and read on scales attached to the receiving tube in the control ship. This latter feature is designed to facilitate the checking of instruments in the torpedo prior to release and also while in flight when actual target view is obscured and the automatic pilot is in control.

The particular significance of this paper lies in the large time interval which has elapsed since its preparation. This time element gives adequate proof of the author's foresight and ingenuity, particularly when the details of the system outlined are compared with those of systems in use today, 12 years later.

THERE have been quite a number of attempts to devise an efficient flying weapon. The aerial bomb is the simplest form, and the recent improvements in aerial ballistics make these bombs a most formidable modern weapon. The use of such a bomb usually requires a close approach of the bombing airplane to the target, thereby subjecting the plane to the barrage of the anti-aircraft batteries. It follows that, simultaneous with the development of aerial bombing, there has been improvement in anti-aircraft artillery which has considerably lessened the effectiveness of the aerial bomb.

Considerable work has been done also on the development of radio-controlled and automatic program-controlled airplanes having in mind

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their use as flying torpedoes. The possibilities of such airplanes were demonstrated repeatedly in various countries during the past few years. Both these methods, however, have the same fundamental difficulty, viz., that they can be used efficiently only by trained personnel at a comparatively close range, thereby being subjected to anti-aircraft gun-fire. Both radio and automatic-controlled planes lose their efficiency as soon as they are beyond visual contact with the directing base. The solution of the problem evidently was found by the Japanese, who, according to newspaper reports, organized a Suicide Corps to control surface and aerial torpedoes. The efficiency of this method, of course, is yet to be proven but if such a psychological training of personnel is possible, this weapon will be of the most dangerous nature. We hardly can expect to introduce such methods in this country, and therefore have to rely on our technical superiority to meet the problem.

GENERAL DESCRIPTION

One possible means of obtaining practically the same results as the suicide pilot is to provide a radio-controlled torpedo with an electric eye. This torpedo will be in the form of a small steep angle glider, without an engine, and equipped with radio controls and an iconoscope camera. One or several such torpedoes can be carried on an airplane to the proximity of where they are to be used and there released. After it has been released the torpedo can be guided to its target by short-wave radio control, the operator being able to see the target through the "eye" of the torpedo as it approaches.

The carrier airplane receives the picture viewed by the torpedo while remaining at an altitude beyond artillery range. It is not even necessary to have direct visibility of the target from this plane, as the information is supplied by the torpedo from a much closer range. The distance between the plane and torpedo will always be short; therefore the power of the short-wave radio transmitter on the torpedo can be very low. A transmitter of 5 or a maximum of 10 watts, operating between 3 and 10 meters, will be sufficient for this purpose. Since the image of the target increases in size as the torpedo approaches, it is not necessary to provide an electric eye with great resolution or with highly efficient optics. Therefore, an iconoscope camera operating with a 90-line picture and with a wide-angle lens will be sufficient for this purpose. When the torpedo is first launched from the plane it may not immediately supply any useful information to the control due to the excess height or intervening clouds, but when it begins to approach the ground the visibility will gradually increase and the accuracy of the aiming will be improved. At close range the target will be sufficiently large to provide good visibility even at 90 lines and the accuracy of the aiming will be the greatest just before the moment of contact of the torpedo with the target. This introduces an entirely new principle in ballistics, since in all existing methods the operator has no way of controlling a projectile once it has been released.

The radio receiving equipment of the torpedo can also be simplified by using a directional and more powerful radio transmitter on the mother plane and also by a decrease in the width of the communication channel, due to the necessity of transmitting from the controlling plane only three or four sets of signals. This can be accomplished by a short-wave carrier modulated with widely separated frequencies. The necessary electrical supply for the torpedo is easily obtainable from a propeller-driven generator with several commutators supplying all necessary direct-current potentials.

The radio control of the torpedo can be accomplished by using one of the already-developed schemes, but can be considerably simplified by using, for control purposes, a circuit and tubes which were developed during the past couple of years in connection with radio communication.

It is very difficult to specify at present the probable weight of the total electrical equipment without actually building a model. A preliminary estimate shows that the total weight of the equipment with automatic pilot, including wind-driven generator, will be below 150 pounds, or less than the weight of one pilot. This weight is composed of the following items:

(1)	Iconoscope camera for 90 lines with deflection and tilting arrangement, and short-wave radio transmitter with modulation up to 100,000 cycles and 10 watt power	40 pounds
(2)	Wind-driven generator for 1000, 300 and 6 volts—125 watts, and three control drums	45 pounds
(3)	Short-wave radio receiver for 3 audio tuned channels with relays	15 pounds
(4)	Automatic pilot with controls and accessories	40 pounds
	Total	140 pounds

The weight of the torpedo can be composed, for instance, of: Control equipment — 140 pounds; fuselage — 120 pounds; explosives — 300 pounds, making the total weight 560 pounds per unit.

Due to the fact that the torpedo has no landing speed, the load per square foot of the wing area can be increased several times in comparison with that of an airplane, therefore the whole torpedo can be made very compact. Four such torpedoes can be packed under the wings of a normal sized bomber. If necessary, the amount of explosives indicated above can be increased by increasing the size of the torpedo.

An approximate idea of the appearance of such a torpedo is shown in Figure 1, which, of course, is probably very far from the actual shape that such a torpedo will have after its final development.

ICONOSCOPE CAMERA AND TRANSMITTER

The iconoscope camera has already been developed by us for television purposes. However, due to the decrease of the required number of lines from the present 340 to 90 lines, the whole apparatus will be substantially simpler and smaller. The associated circuits for deflection of the electron beam and the amplifier will contain only a fraction of the number of tubes used in the present system.

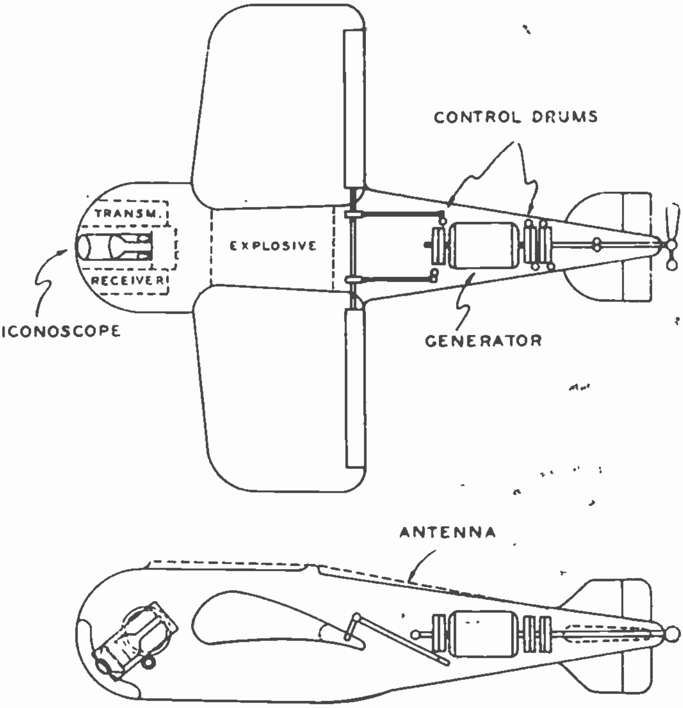

Fig. 1—The Flying Torpedo.

The camera mounting is provided with a tilting arrangement, which points it always in such a manner that the center of the received picture coincides with the point to which the torpedo is heading. The tilting is controlled by the same device which controls the level flight of the torpedo and which will be described later. The optical lens of the camera is provided with a sighting cross-wire which, on the reproduced picture, gives the point on which to sight the torpedo.

It appears that it is desirable to watch from the controlling plane not only the picture viewed by the torpedo, but also the condition of the controls of the torpedo, the acceptance of the controlling signals, altitude, etc. This is particularly important if the torpedo is not launched directly at the visible target, but has to pass first through intervening clouds. Such an arrangement can easily be achieved practically without introducing any additional complication in the camera circuits. All the necessary information can be transmitted on the edges of the picture by projecting the small light spots on the sensitive mosaic of the iconoscope. These spots are reflected by mirrors attached either to the controlling mechanism directly, or through the medium of small sized electric motors. This arrangement is shown in Figure 2.

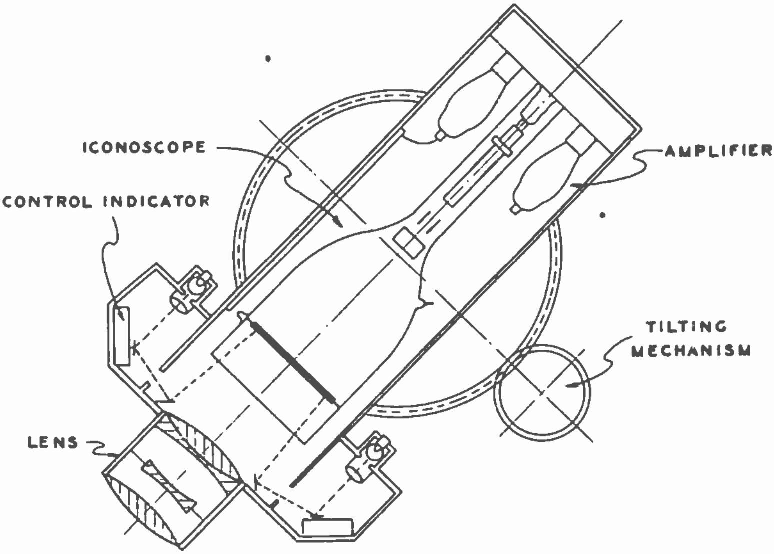

Fig. 2—Iconoscope camera for the Flying Torpedo.

In this way, the information will be given by the position of bright spots on the edges of the picture and can be read accurately on scales attached to the receiving tube in the controlling ship. Since all instruments can be set in operation while the torpedo is still attached to the controlling ship, the function of these instruments, and therefore the preparedness of the torpedo for action, can be checked all the time and particularly just before its launching. It is easy to provide the adjustment connection which would enable the operator on the controlling ship to reset the instruments in the torpedo according to the reading of the accurate instruments of the controlling ship. Due to the fact that the actual free flight of the torpedo will take from one to a maximum of 10 minutes, this initial setting will be kept by the instruments of the torpedo, and the readings on the scale attached to the receiving tube will be very accurate. The appearance of the picture on the receiver with the control indicating spots is shown in Figure 3.

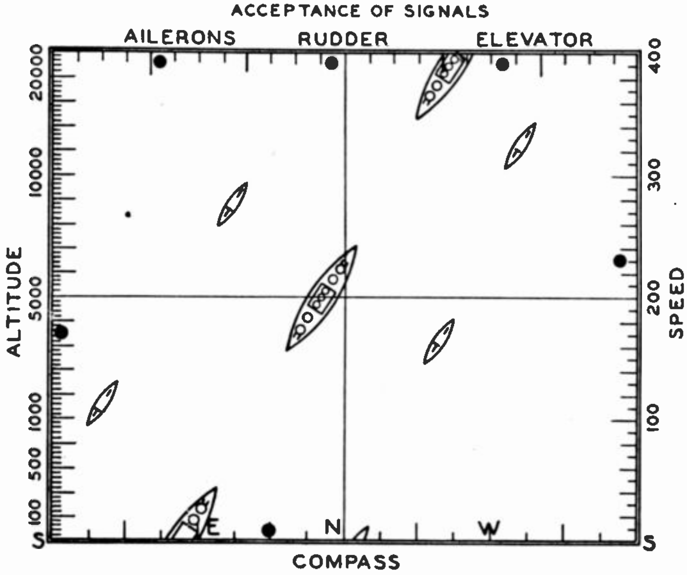

Fig. 3—Sketch of a typical scene reproduced by the Flying Torpedo.

The radio transmitter for the 90-line picture with 16 frames per second requires a modulation band of approximately 100,000 cycles, or only one-tenth of what we are using in our present television system. Such a transmitter for 10 watts output is very simple and requires a small number of tubes. The antenna will be located on top of the fuselage and combined with the receiving antenna.

POWER GENERATOR

The voltages necessary to operate the picture transmitter as well as the radio receiver and most of the controls can be confined to three potentials of 1000, 300 and 6 volts. The total power requirement will be about 125 watts. A conventional wind-driven generator with three commutators and automatically adjustable propeller will answer the purpose. The shaft carrying the generator also has three drums, which supply the power for the controlling mechanisms.

CONTROLS

To guide the torpedo from the mother plane, the controlling signals are supplied by the second radio channel, also on ultra-short waves, from the transmitter located on the mother plane. The sensitivity of the radio receiver on the torpedo can be very low, on the order of 10 millivolts. Signals operating the different controls can be separated from one another by sharply tuned filters in the output of the radio receiver.

In order that the torpedo will respond quickly to the controlling signals, the power necessary to move the rudder, elevators and ailerons is taken from drums rotated by the same propeller-driven shaft which carries the generator. Each drum has two friction bands which can be energized by the output of the amplifying tube. This energizing is accomplished either by a relay which tightens the grip of the band around the drum, or directly by an electrostatic or electromagnetic field supplied by the output of the control tube to the band. These two friction bands are connected to a corresponding controlling device, for instance, the rudder, moving it in either of two opposite directions. The shaft operating the rudder carries two potentiometers varying the biases of the corresponding controlling tubes in such a way that the tube which is energized, and therefore moves the rudder in one direction, is biased gradually negatively and the opposing tube at the same time is biased positively. This biasing serves two purposes: First, it allows the control to operate fastest near the neutral position and slow down according to the prescribed formula with the increase of the angle of rotation. It also checks the maximum permissible controlling angle. When the controlling impulse ceases then both tubes, the one which just functioned and the opposing one, will be in an unbalanced state and immediately start to move the controlling element toward the neutral position.

The banking of the torpedo will not require a separate controlling signal because it is possible to arrange the banking to follow automatically the controlling directional signal. This is accomplished by coupling the tube of the rudder so that the movement of the rudder will be automatically followed by the movement of the ailerons according to the prescribed relation. If necessary, this movement can also be made a function of the speed of the torpedo by operating the bias of the aileron control tube from the altimeter or the rate-of-descent indicator. In this way, the banking can always be made exactly right regardless of the speed at which the torpedo turns.

AUTOMATIC PILOT

Where the torpedo is to be used against an objective that is not visible from the launching point, it will be necessary to equip it with an automatic pilot as well as remote control. The additional apparatus necessary to accomplish this does not add greatly to its weight or complication.

Stabilization of the torpedo when it is not under remote control is accomplished by two gyros. The construction of these gyros can be the same as the gyro employed in the "artificial horizon" or in the "directional gyro." They are operated either by air from the Venturi tube, or by an electric motor supplied by the main generator. The gyros ought to be fully stabilized, but can be made much smaller than the gyros used in the above mentioned instruments. The control of these gyros is accomplished by means of a mirror, attached to the gyros, which reflects a beam of light into the photocells. The photocells are arranged in pairs serving as two arms of a Wheatstone bridge. The cells, or rather the openings through which the light falls on the cells, are of a wedge shape, as shown in Figure 4.

By this arrangement, the reflected light at zero position of the gyro produces two equal impulses in both cells and therefore balances the bridge. When the body of the torpedo turns with respect to the gyro, the line of light begins to turn with respect to the neutral axis and increases the impulse in one of the cells, decreasing it in the other. At the limiting angle, the impulse from one cell will be zero and maximum in the second, giving a maximum unbalance. Of course, the shapes of the openings can be made according to any prescribed condition so that the change in impulses can follow a desired mathematical relation between the turn of the bomb and the condition of the electrical circuit. By using an alternating-current amplifier tuned to the frequency of the rotation of the gyro, multiplied by the number of mirrors attached to it, it is possible to operate the amplifier, not only from the actual displacement of the torpedo, but also from its first or second derivative, thereby increasing the sensitivity of the controlling circuit. In order to set the control to a desired condition, it is necessary only to turn the housing with the photocells and gyro to a certain angle in respect to the axis of rotation of the stabilizing gyro. This will upset the equilibrium condition of the circuit and will energize the controlling tubes operating the controlling bands, as mentioned in the description of the remote control of the system. When the torpedo attains the new prescribed flying condition the circuit is again balanced and the course of the torpedo will coincide with the neutral position of the controlling gyro. In this way, any outside influence such as a gust of wind, air pockets, etc., which affect the initial course of the torpedo, will be corrected immediately by the controlling gyro.

In order to keep the torpedo under control when the controlling signals cease, the automatic control should be adjusted to a new set of conditions prescribed by the controlling signals. For this purpose, when the controlling signal changes the initial course of the torpedo, the position of the photocells with respect to the controlling gyro will automatically reset itself to balance the bridge according to the new set of conditions, and therefore will be automatically established for this new course in the controlling gyro circuit. The motion for this adjustment of photocells can be made either by friction from the same drums, or can be provided by separate small electric motors.

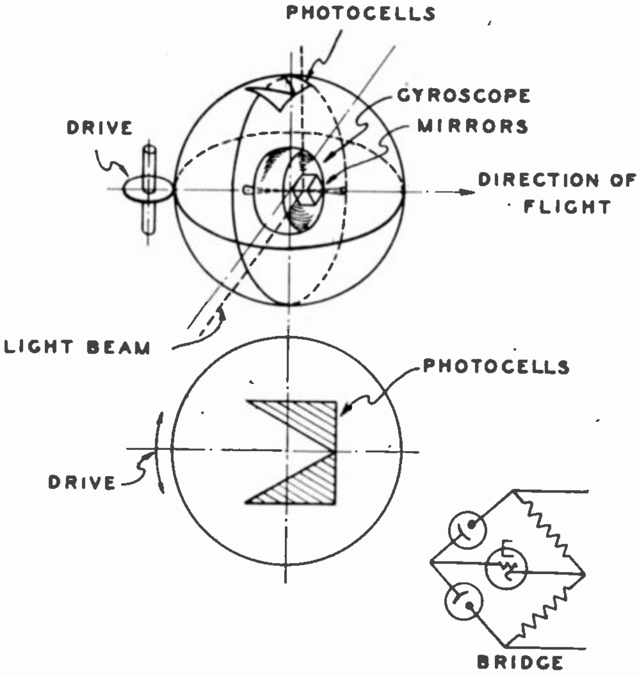

Fig. 4—Control mechanism of the Flying Torpedo.

METHOD OF OPERATION

When the torpedo is attached to the mother plane, provision is made to manually adjust all the necessary instruments according to the more accurate instruments located in the mother plane. Also, all the apparatus and controls of the torpedo can be checked by starting the camera and observing the position of control indicators in the receiver.

If the torpedo is to be launched directly at the visible target, the control is very simple and all that is necessary is to keep the center of the picture or the cross-wire sight on the target all the time. For this kind of operation, the automatic pilot is unnecessary; therefore, if only this type of operation is expected, the equipment for the automatic pilot can be omitted. If, while the torpedo is approaching the target, it becomes clear that due to the faulty steering, cross winds, etc., it may miss the target, it is entirely possible to steer the torpedo through a loop, gain altitude and repeat the attempt two or more times before its speed is lost, preventing the further repetition of this maneuver.

It may be desirable to launch the torpedo while the mother plane is at a very high altitude, or screened from the target by intervening clouds, with only approximate information of the position of the target. In this case, the torpedo is launched down in a spiral glide and kept under control of the automatic pilot and also under manual control by observing the controlling marks on the picture. When the torpedo descends low enough to make the target visible, then the target can be brought into coincidence with the cross-wire sight and the torpedo started to glide directly to the target.

A more elaborate form of this torpedo is a regular airplane equipped with an engine and the same controls as described above. This plane can be launched either from a small surface craft or from the shore at a very distant target, and then controlled according to the picture received through the iconoscope camera. This makes this new weapon very versatile since it can be used both on the sea and land.

NAVAL AIRBORNE TELEVISION RECONNAISSANCE SYSTEM*†

By

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Summary—A high fidelity long range television reconnaissance system developed during World War II for the Navy Department is described. The Project Ring equipment was designed for multi-camera attended operation at 20 frames/second, 40 fields/second, 567 lines/frame interlaced and utilizes a 5 megacycle video bandwidth. A high power (1400 watt peak) airborne television transmitter is employed. The maximum plane-to-ground transmission range attained during tests was over 200 miles. Very consistent operation with satisfactory signal-to-noise ratio has been obtained with this equipment at ranges of 100 miles or more with the aircraft flying at altitudes of 7000 to 10,000 feet. The equipment differs from the light weight simplified television gear designed during World War II for unattended operation in guided missiles.

INTRODUCTION

DURING World War II two general types of airborne television equipments were developed for the U. S. Armed Forces. One type, known by the code designation "Block" was a simplified, light-weight system designed for unattended operation in drone aircraft and guided missiles. A second type, described herein, was developed for long-rang, high-altitude reconnaissance operations. This equipment was designed for attended operation, with weight and complexity considerations secondary to the production of high definition television pictures suitable for airborne military reconnaissance. The project under which the development was carried out was known by the code designation "Ring".

Work on Project Ring was initiated in November 1942 when standard broadcast type transportable television pickup equipment utilizing the type 1840 orthicon camera tube was demonstrated to representatives of the Navy Department and Marine Corps. As a result of this demonstration, in which the pickup equipment was installed on the 85th floor of the Empire State Building to simulate an aircraft at 1000 foot altitude, it was concluded that the information presented by the television screen probably was sufficiently detailed to have value for airborne military reconnaissance. In order to check this conclusion, it

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was necessary to conduct a series of actual flight tests using television pickup equipment installed in aircraft. Accordingly, arrangements were made by the Navy to carry out such tests at the Naval Air Station, Banana River, Florida using a PBY-4 "Catalina" flying boat.

In making these preliminary flight tests, the use of readily available equipment, even though designed for a different service, was dictated by two factors—the need for a quick answer on the possibilities of television reconnaissance, and the lack of adequate previous experience or data on which the design of specialized high-fidelity equipment might be based. In order to make a start, therefore, standard commercial transportable television pickup gear utilizing the type 1840 orthicon tube, was mounted in the PBY-4 aircraft. The equipment consisted of two cameras, one in the bow position and the other in the waist gunner's position arranged to view from the starboard machine gun blister with the latter open. The cameras were connected via standard multi-conductor cables to the control position, which was located amidships in the space normally used by the navigator. At the control position were the camera control units with their self-contained video monitors, the master switching unit with its monitor showing the outgoing picture, the synchronizing generator and the electronically-regulated power supply rectifiers. A photograph of the control position installation is shown in Figure 1.

In order to obtain sufficient power at 115 volts 60 cycles alternating current for operation of the television equipment, a special auxiliary power unit had to be installed in the aircraft. Here again, it was necessary to use equipment which was readily obtainable. An available 5-kilovolt-ampere, 115-volt, 60-cycle, gas-engine-driven alternator was mounted in the compartment aft of the control position. This unit weighed several hundred pounds and had to be dismantled and then reassembled inside the aircraft in order to place it in position.

A low power video transmitter (60 watts peak), developed for another application, was adapted for the experiments to expedite the work. The receiving equipment on the ground consisted of Navy type Block 1 television receivers feeding viewing monitors equipped with 12-inch kinescopes. A projection type viewing monitor, utilizing refractive optics and producing a picture 18" x 24" in size, was also installed at the ground station.

The ground station receivers, having been designed for airborne use, were operated from 28 volts direct current. The viewing monitors were modified commercial television receivers and were operated from the local 60-cycle mains. The transmission standards used were the 525 line, 30 frame, 60 field interlaced commercial broadcast standards for which the equipment (except the Navy Block I receivers) was designed. The transmitter operated on a carrier frequency of 90 megacycles with an effective bandwidth of 4.5 megacycles on each side of the carrier. Negative transmission (maximum carrier amplitude corresponding to tips of synchronizing pulses) was used. The antenna polarization was vertical. This polarization was chosen principally because it is simpler to obtain uniform azimuthal radiation from an aircraft with a vertical antenna.

A comprehensive series of flight and ground tests using this hastily assembled airborne television system was conducted during the spring and summer of 1943 at the Naval Air Station, Banana River, Florida, culminating in a demonstration for Navy, Marine Corps and NDRC personnel in September of that year.

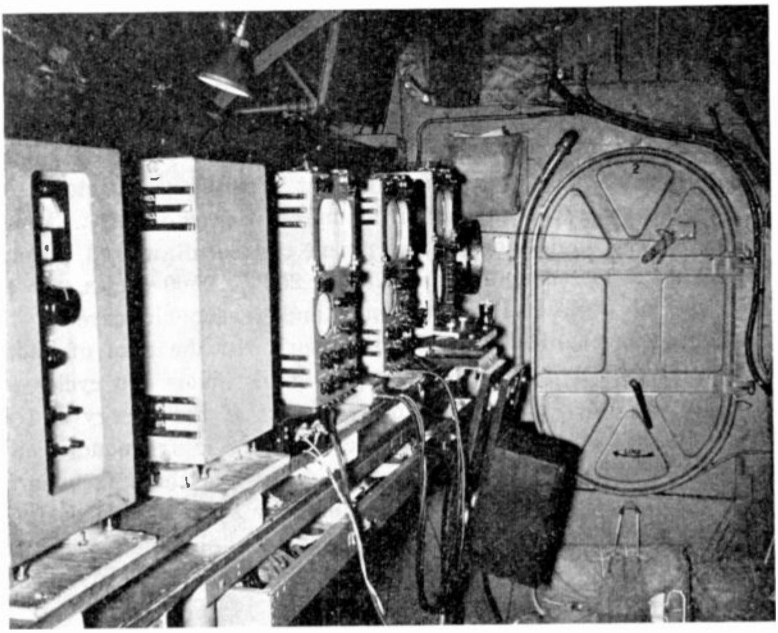

Fig. 1—Control position in PBY-4 aircraft installation using commercial television equipment.

Results of Initial Airborne Tests

These initial tests using available commercial television equipment in an airborne system provided information and experience on which to base the design of specialized airborne television equipment.

This experience and information may be classified under the follow-

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† NDRC—National Defense Research Committee.

ing headings:

(a) General electrical and mechanical requirements for airborne television equipment.

(b) Desirable operating and maintenance features in airborne television equipment.

(c) Purely technical features, such as transmitter power output, types of antennas, practical video bandwidths, etc.

A brief summary of the conclusions drawn from the results of these tests is given in the following paragraphs.

In regard to the general electrical and mechanical design requirements for airborne television equipment, it may be said that they are similar to the requirements for satisfactory airborne radar or radio communication gear. Ruggedness, compactness and light weight, as well as the ability to operate satisfactorily over wide ranges of altitude, temperature and humidity are required of airborne television gear. In addition, special precautions must be taken to eliminate microphonic disturbances, as television equipment is more susceptible to this type of trouble than either radar or communication equipment. Microphonic effects in communication equipment can be minimized through the use of a restricted audio bandwidth, (200 to 2500 cycles, for example), special high-level microphones and reasonable care in the application of mechanical vibration isolators. In the case of radar, most systems do not deal with low-level signals below 500 cycles per second. On the other hand, an important part of the energy and information in a television signal is included in the frequency range from 20 or 30 cycles to 1000 cycles per second and these frequencies must be dealt with in low-level video amplifier stages. The designer of airborne television equipment must, therefore, observe special precautions to minimize the effects of vibration and acoustical noise if microphonic effects are to be minimized.

The initial flight tests using commercial transportable television equipment also served to emphasize the practical operating requirements for airborne gear. The equipment should be simple to operate. All electrical adjustments should be made at the control position, leaving the camera operators free to direct the cameras, change lenses and adjust lens iris stops as required. All electrical controls which may have to be changed in flight should be equipped with adjustment knobs rather than being left as "screwdriver adjustments" as the latter are not practical when operating in bumpy air. Also, complete equipment bench test facilities should be provided at the ground station for servicing the airborne equipment, as only a limited amount of servicing is possible when the aircraft is in flight. Ground test facilities should also include an auxiliary power source for pre-flight ground checks of the operation of the equipment prior to take-off.

Definite conclusions on the technical requirements for a satisfactory airborne television system were drawn from the initial tests. Some of these are:

1. The television pickup equipment preferably should include two cameras, one with a short focal length lens for wide angle views and the other with a long focal length lens for telephoto views.
2. All electrical adjustments should be made by an operator at a master control position. The master control position should provide facilities for rapid switching of the output of either camera to the video transmitter. The master control operator should be provided with an oscilloscope, for checking video signal levels, and two video monitors. Switching facilities should be arranged so that either of the monitors or the oscilloscope can be independently switched to camera No. 1 video output, camera No. 2 video output, the input to the modulator of the video transmitter, or to the output of a radio frequency detector coupled to the transmitting antenna.
3. Means should be provided for rapid checking of the depth of modulation of the video carrier during the operation of the transmitter and pickup equipment.
4. Interlaced scanning should be used to gain the maximum picture resolution, within the limits of tolerable image flicker for a given video bandwidth.
5. The video transmitter should have adequate power output to provide reliable picture transmission up to 100 miles.
6. Special precautions should be observed to eliminate the effects on interlaced scanning of power supply frequency differences between the airborne power supply and the ground station power supply.
7. The type 1840 orthicon was found to be unsatisfactory for airborne operation under illumination conditions combining both high peak scene brightness and high scene contrast. The iconoscope type pickup tube, though satisfactory in this respect, is lower in light sensitivity than is desirable for general airborne use. It was therefore concluded that newer developmental types of tubes, which gave promise of improved results, should be considered in the design of specialized equipment for airborne use.

As a result of these tests, a study of general military television requirements, observations and reports on other military television equipments, and a study of commercial television systems, recommendations and specifications for a set of airborne television reconnaissance equipment were drawn up and submitted to the Bureau of Ships of the Navy Department. It was considered that a practical airborne television reconnaissance system should provide reliable picture transmission over a distance of at least 100 miles with the aircraft at an altitude of 7000 feet. Having settled on the range requirements, the other features of the proposed system—keeping in mind that maximum image resolution was required—were largely determined by the practical limitations imposed by such factors as the maximum practical video transmission bandwidth, the minimum frame repetition rate within the limits of tolerable image flicker and blurring (due to the relative motion between the aircraft and objects on the ground), attainable signal-to-noise ratios, and the state of the television art at the time. Determination of the system standards is discussed in the next section.

SYSTEM DESIGN

The radio propagation characteristics between an aircraft at various altitudes and distances and a receiving antenna 30 feet high is shown for vertical polarization over sea water by the theoretical curves[1, 2] of Figure 2. Figure 3, which gives calculated[1, 2] values for horizontal polarization over land, is considered to be reasonably accurate also for vertical polarization. Supplementing the above propagation studies, it was determined by measurements on practical television receivers and the results of the initial flight tests at Banana River, that, in this application, the minimum useable signal at the receiver input terminals would be on the order of 50 microvolts, (5-megacycle video bandwidth). From the above data, it was estimated that for a 100 mile range (aircraft at 7000 ft.) with a vertical dipole receiving antenna 30 feet above the ground, the average carrier power output of the transmitter should be in excess of 200 watts (800 watts peak). Vertical polarization of the antennas was chosen because of its non-directional characteristic in the horizontal plane; also, cancellation of the direct path and multi-path signals is not as severe, especially at relatively short distances with the aircraft at high altitudes, as when horizonal polarization is used.

Consideration of various limiting factors, such as weight, size, the

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[1] K. A. Norton, "The Effect of Frequency on the Signal Range of an Ultra-High Frequency Radio Station with Particular Reference to a Television Broadcast Service", Statement made before the Federal Communications Commission, Television Hearing, Report No. 48466, March 20, 1941.

[2] K. A. Norton, "The Calculation of Ground-Wave Field Intensity Over a Finitely Conducting Spherical Earth", *Proc. I.R.E.*, Vol. 29, No. 12, pp. 623-639, December, 1941.

required power output, available carrier frequencies and the requirement for maximum video bandwidth led to the design of an amplitude modulated video transmitter producing an average carrier power output of 350 watts (1400 watts peak) with a video bandwidth of 5 megacycles (i.e. 5 megacycles above and below the carrier frequency), operable at carrier frequencies of either 90 or 102 megacycles.

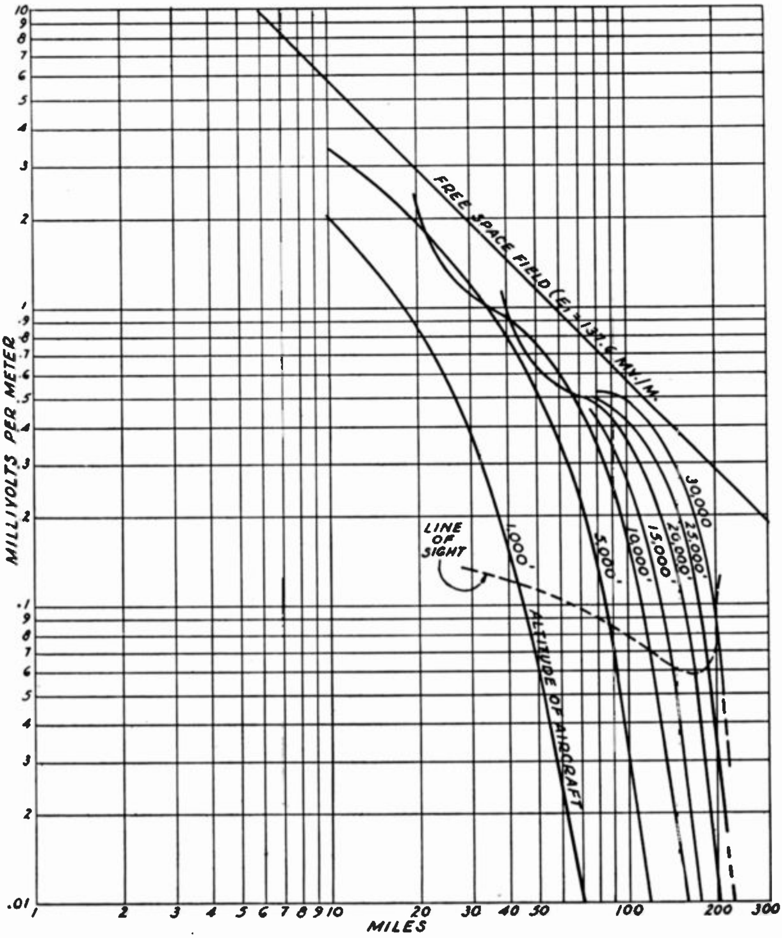

Fig. 2—Calculated field strength intensity versus distance for the indicated altitudes of the aircraft. (Receiving antenna 30 feet high, 200 watts radiated power at 100 megacycles, for vertical polarization over sea water with a ground conductivity of 4.3 mhos/meter and a dielectric constant of 81.)

Figure 4, which is a plot of actual flight test data obtained with a developmental model of this transmitter, shows that this power output is adequate for the required transmitting range and shows good correlation with the calculated curves of Figures 2 and 3. While a video bandwidth greater than 5 megacycles could have been obtained with this design at a correspondingly lower power output, the 5 megacycles was chosen as the best compromise value.

Having determined the video bandwidth (5 megacycles), the next step in the design was to select the system standards to attain the greatest practical image resolution. To secure the maximum amount of information in the television image, it is desirable to operate at the

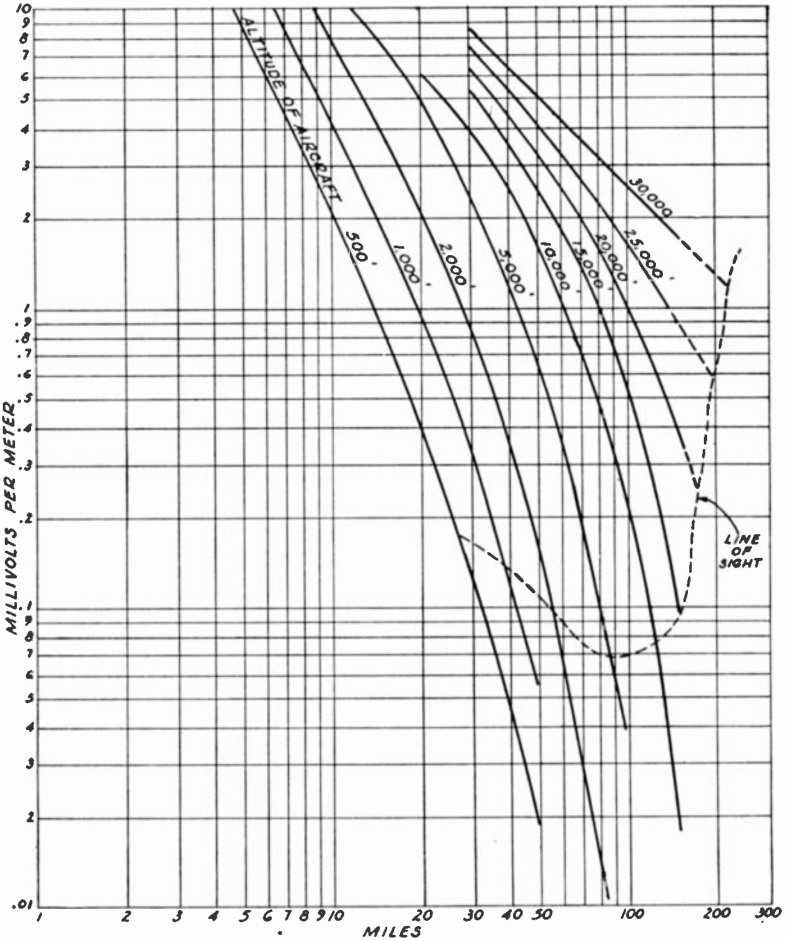

Fig. 3—Calculated field strength intensity versus distance for the indicated altitudes of the aircraft. (Receiving antenna 30 feet high, 200 watts radiated power at 105 megacycles, for horizontal polarization over land with a ground conductivity of 5×10^{-14} electromagnetic units and a dielectric constant of 15.

lowest frame repetition rate practicable. However, the frame rate cannot be reduced below the point where blurring occurs due to the relative movement between the aircraft and objects on the ground or where image flicker becomes troublesome. In addition, interlaced scanning is dictated if maximum picture resolution with minimum flicker is to be realized. As a result of a number of laboratory tests to determine minimum frame frequency, it was found that a frame repetition rate of 20 per second with a field frequency of 40 per second (2:1 interlaced) was the lowest practical value that could be used for this appli-

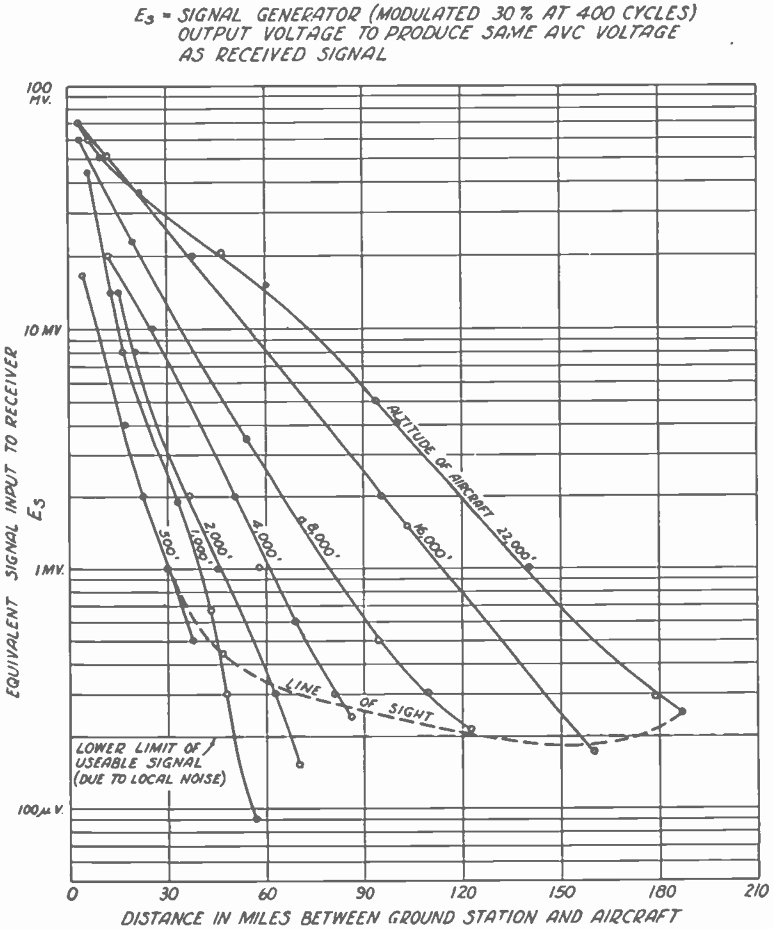

Fig. 4—Measured values of relative signal input to the receiver versus distance, the aircraft flying at the altitudes indicated, receiving antenna height 30 feet, and 350 watts radiated power. (Multiply E_s by 0.2 to obtain the approximate field strength at the receiving antenna.)

cation. The amount of flicker produced by this low frame rate is greater than would be acceptable for entertainment television but is tolerable for television reconnaissance. The aspect ratio of the image was chosen to be 4:3, the same as commercial television standards, after due consideration of military requirements. The Project Ring standards and commercial television standards are compared in Table I. The theoretical total numbers of picture elements per frame for the two systems are included as a matter of interest. The theoretical resolution, however, generally cannot be fully realized in practice for various reasons, such as the relative movement between the object being televised and the camera, aperture effects of the camera pickup tubes and the receiver kinescopes, the effects of shot and thermal noise, residual phase distortion in the overall system and other practical limitations.

Table 1

	Project RING	FCC Standards
Video bandwidth—megacycles	5	4.25
Fields per second (2:1 interlaced) .	40	60
Frames per second	20	30
Aspect ratio	1.33	1.33
Vertical blanking—per cent of field period	8 per cent	7 to 8 per cent
Horizontal blanking—per cent of line period	18 per cent	16 to 18 per cent
Lines per frame	567	525
Line frequency	11,340	15,750
Vertical resolution (lines)	522	483
Horizontal resolution (lines)	500	311
Horizontal resolution × Aspect ratio	667	415
Total picture elements/frame	348,174	200,445

In the above choice of standards, the horizontal resolution of the system has been made approximately equal to the vertical resolution. The horizontal resolution (lines) is:

$$N_h = \frac{2f\,(1 - T_v)\;(1 - T_h)}{N_v\quad(a)\,(r)}$$

where f is the video bandwidth in cycles, T_v is the fraction of the vertical period devoted to blanking, T_h is the fraction of the horizontal period devoted to blanking, N_v is the number of lines per frame, a is the aspect ratio and r is the number of frames per second.

To simplify the design of the synchronizing generator, as well as to reduce its weight and size, a somewhat simplified type of synchronizing signal was chosen for interlaced scanning. This type of synchronizing signal, developed by RCA Laboratories Division,[3,4] is shown in Figure 5. This waveform is not as flexible as to the number of different ways it may be utilized in a television receiver as is the synchronizing signal prescribed by the Federal Communications Commission for commercial television broadcasting. The integration time required for vertical synchronization, to secure a well-interlaced picture, is several times longer than that required by the standard commercial synchronizing waveform. However, this type of synchronizing signal has proven satisfactory for this application and resulted in the saving of at least ten tubes and associated components in the synchronizing generator.

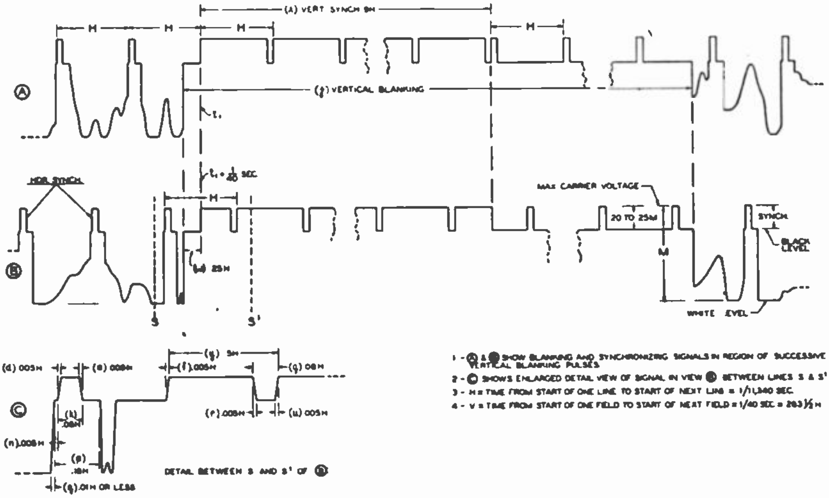

Fig. 5—Synchronizing signal waveform used for Project Ring.

In addition to the above standards which have to do with scanning and video bandwidth, it was necessary to standardize on several other items. It was decided to use amplitude modulation and transmit both side bands. The choice of amplitude rather than frequency modulation was based upon the following considerations:—

1. A great deal more experience was available in the use of amplitude modulation for video transmission than in the use of frequency modulation.

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[3] A. V. Bedford, "Synchronizing in Television", RCA Report prepared for use by Panel 8 of the National Television Systems Committee, October, 1940.

[4] D. G. Fink, TELEVISION STANDARDS AND PRACTICE, McGraw-Hill Book Company, New York, N. Y., 1943.

2. Pre-war field tests employing FM† with vestigial side band transmission and without limiting in the receivers gave results that were unsatisfactory, since this system was considerably more vulnerable to multipath transmission effects than was the standard AM* system.
3. Theoretical studies and laboratory tests had been reported showing that a FM† signal would be more vulnerable to multipath effects than an AM* signal.[5]
4. Under an NDRC contract covering work on the Block project, comparative tests were made between FM† and AM* using airborne transmitters operating near 300 megacycles. This work was not carried to conclusion but the preliminary results obtained indicated again that FM† was more vulnerable to multipath effects than was AM.*

Offsetting the above unfavorable experience with FM for television transmission was the highly successful use of FM for a 500 megacycle television relay link reported by RCA Communications, Inc.,[6] and other work done utilizing FM for airborne television transmission.

In view of the above uncertainty, it was decided that the safest course would be to employ amplitude modulation for this application. It is emphasized, however, that this finding should not be construed as necessarily applying in the future, particularly at higher frequencies and with the application of more thorough development of FM for television.

The decision to transmit both side bands was based purely upon the desire to avoid the additional weight and complications of a side band filter in the airborne transmitter. Under many conditions, vestigial side band operation of the receiver was utilized by proper detuning of the receiver circuits.

Negative transmission (i.e. the tips of the synchronizing pulses corresponding to maximum carrier amplitude) was chosen for Project Ring for the same reasons which led to its choice for commercial television broadcasting—principally because interference produces, in such operation, less objectionable effects in the received picture than with positive modulation.

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† FM—frequency modulated or modulation.

* AM—Amplitude modulated or modulation.

[5] Murlan S. Corrington, "Frequency-Modulation Distortion Caused by Multipath Transmission", *Proc. I.R.E.*, Vol. 33, No. 12, pp. 878-891, December, 1945.

[6] F. H. Kroger, Bertram Trevor and J. Ernest Smith, "A 500-megacycle Radio-Relay Distribution System for Television", *RCA REVIEW*, Vol. V, No. 1, pp. 31-50, July, 1940.

CHOICE OF POWER SUPPLY

The presence in a television receiver of power supply hum of a frequency different from the field repetition rate usually makes interlaced scanning operation difficult. Considerable trouble from this source was encountered in the early tests with the PBY-4 installation at Banana River. In that case, the efficiency, and therefore the speed, of the gas engine driving the airborne power supply would change with altitude in such a way that the power supply frequency would be several cycles per second lower than the nominal 60-cycle value by the time the aircraft had reached an altitude of 5000 feet. At the same time, the local ground station power supply, also nominally 60 cycles, would vary above and below this value at different times of the day. This resulted in a situation where the airborne and ground power supply frequencies would often be different by several cycles per second. Under these conditions, a good interlace was always obtained on the monitors in the aircraft. The airborne synchronizing generator and therefore the vertical scanning frequency was locked in with the airborne power supply; consequently, any residual power supply hum in the pickup or monitoring equipment did not show up in the airborne video monitors as it produced a steady and almost unnoticeable pattern on the television picture. (The vertical scanning frequency automatically remained synchronous with the airborne power supply alternator even though the latter varied in frequency). Considerable trouble with interlace was experienced at the ground station due to this power supply frequency difference because the electrical filtering and shielding of the modified commercial television receivers used as monitors was inadequate for operation from a power source not synchronous with the vertical scanning frequency. Trouble from this source was minimized by additional filtering and shielding of the monitor circuits and kinescopes against power supply hum effects.

Obviously, use of pure direct current power supply sources for the airborne and ground equipments would have been an ideal solution to this problem. However, such a solution for the airborne equipment would have been uneconomical because of the amounts of power required and the number of different direct current voltages needed. An all-direct-current power supply system using individual dynamotors operated from 28 volts also would weigh more and be less efficient than an equivalent alternating current system.

A practical solution to the airborne power supply problem, in the case of the equipment specially designed for Project Ring, was found in the use of a combination of 28-volt direct current and 115-volt 400 to 2400 cycles alternating current power sources. The 28-volt direct current power is used for the filaments and heaters of the tubes in the pickup and transmitting equipment. The 115-volt 400- to 2400-cycle alternating current is used for all plate and bias supply rectifiers. The vertical scanning frequency is not locked in with the alternating current power supply, but is the 567th sub-multiple of a stable vacuum tube oscillator operating at 22,680 cycles per second, (twice the line scanning frequency). The vertical scanning frequency (field frequency) therefore remains very close to 40 cycles at all times regardless of the frequency of the airborne power supply. Adequate power supply filtering and careful magnetic shielding of the pickup and other cathode-ray tubes is effective in reducing the residual power supply hum to a sufficiently low value to allow satisfactory interlaced scanning operation. The use of 28 volts direct current for all filaments, except the video monitors and the oscilloscope, was found to be very helpful in reducing power supply hum trouble.

It was desirable to design the ground station receiving and monitoring equipment for transportable operation using a basic power supply of 28 volts direct current. All ground station equipment filaments, except those of the kinescope anode supply rectifiers, are operated from 28 volts direct current. The television receiver has a built-in dynamotor supplying plate and bias voltages, while the 4500-volt direct current kinescope anode supply is obtained from rectified pulses derived from a step-up winding on the horizontal deflection output transformer. The filament of the rectifier for the 4500-volt supply is fed with current at the horizontal deflection frequency supplied by a separate winding on the horizontal deflection output transformer. Plate supply for the viewing monitors may be obtained from a 28-volt to 350-volt direct current dynamotor. The 8500-volt direct current kinescope anode voltage for each viewing monitor is obtained from a self-contained single phase rectifier of conventional design operated from 115 volts 400 cycles, the 400-cycle voltage being supplied by a 28-volt direct current inverter. Adequate filtering and shielding of the 400-cycle rectifier eliminates hum troubles from this source. The ground station equipment may also be operated from 60 cycle alternating current using an electronically-regulated plate supply rectifier for the viewing monitor, a 28-volt rectifier and storage battery for the filaments and a 28-volt direct current to 115-volt 400-cycle alternating current inverter for the 8500-volt anode supply rectifier.

AIRBORNE PICKUP EQUIPMENT DESIGN

The division of the airborne equipment into a number of separate units was dictated by the following factors:

(a) The necessity of providing small packages of reasonable weight and size that could be conveniently removed from the aircraft for servicing and quickly re-installed.

(b) The desirability of designing monitors, power supplies, etc., as interchangeable identical units so that in case of failure of a single monitor or power supply during flight, the rest of the system would still be operable. (Also, a single spare of each type unit could be carried in the aircraft for installation in case of failure during flight.)

(c) The division of the system into a number of relatively small units interconnected by suitable cables provides flexibility in installations in aircraft where space is at a premium, and makes possible more uniform weight distribution.

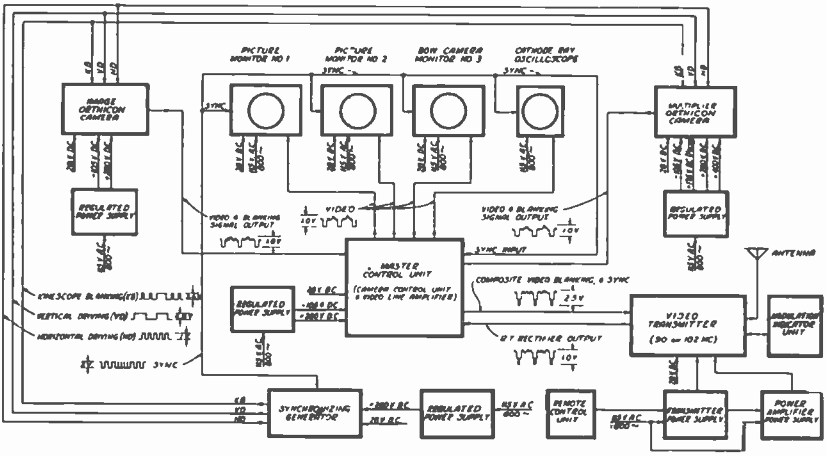

Fig. 6—Block diagram of airborne television equipment.

The block diagram of Figure 6 shows the units comprising the Ring airborne pickup and transmitting system. The video monitors are interchangeable units as are the regulated power supplies (except the multiplier orthicon camera power supply which provides two additional voltage outputs). Identical interconnecting plugs are used on both types of cameras and on both types of regulated power supplies for the video equipment. Plug connections are arranged so that any combination of camera types can be plugged into the system. When two image orthicon cameras are used, all video power supplies are units of the same type.

The block diagram, Figure 6, also shows the video and synchronizing pulse interconnections between the various units. The synchronizing generator provides four pulse outputs with amplitudes and waveforms as indicated on the diagram: vertical driving, horizontal driving, kinescope blanking, and synchronizing signals. The vertical and horizontal driving pulses are fed via 72-ohm coaxial cables to the two cameras where they energize the vertical and horizontal beam scanning circuits. The vertical driving pulses have a frequency of 40 cycles per second and the pulse duration is 6 per cent of one cycle. The horizontal driving pulses have a frequency of 11,340 cycles per second and a duration of 8 per cent of a horizontal scanning period. The kinescope blanking signals are added to the picture signals in each television camera and consist of a composite wave comprising 40-cycle pulses having a duration of 8 per cent of a field period and horizontal blanking pulses (11,340 cycles) having a duration of 18 per cent of one horizontal scanning period. The kinescope blanking pulses are timed to start earlier and last longer than the corresponding driving pulse components. Distribution of the blanking signals is via 72-ohm coaxial cables. The coaxial cable carrying the synchronizing signal output loops through the monitors and oscilloscope which are bridged across the line and is finally terminated in the master control unit. The wave form of the synchronizing signal is as shown on Figure 5.

The feeding of synchronizing signals separately to the video monitors and the oscilloscope has the advantage that these units may be used to monitor video signals which do not contain synchronizing signals. This is the case when the monitors or oscilloscope are switched to the camera video outputs directly, since synchronizing signals are added to the video signals only in the line amplifier which feeds the video transmitter. The monitors and oscilloscope are equipped with "internal-external" synchronizing switches, however, and may be synchronized from the incoming video signal by throwing the switch to the proper position when the incoming signal contains synchronizing pulses. The latter type of operation is desirable as a check on the proper video-to-synchronizing signal amplitude ratio when checking the transmitter output by means of the radio frequency detector provided.

It will be noted in Figure 6 that the master control unit receives video signals from the two cameras as well as from a radio frequency detector. The detector is coupled to the antenna transmission line and extracts and rectifies a small percentage of the transmitter output energy for monitoring purposes. This detected signal is fed to the master control unit via a terminated 72-ohm coaxial line. The control unit provides video monitoring outputs to the cathode-ray oscilloscope and the video monitors. It also houses a line amplifier for feeding the output of either camera to the transmitter. Synchronizing signals are added to the output of this line amplifier.

By means of push buttons provided on the control unit, any monitor or the oscilloscope may be independently switched to Camera No. 1 output, Camera No. 2 output, the control unit line amplifier output, or the output of the radio frequency detector. The front panel of the control unit is also the location of all electrical controls for the two cameras.

While the control circuits used are conventional, the video switching arrangement is thought to present novel features and will be described in some detail. This can best be done by reference to a simplified block diagram of the video and switching portion of the master control unit, Figure 7. It will be noted that Camera No. 1 video output passes through four monitoring networks in cascade, the last network being terminated in a resistance of 72 ohms in the form of an output gain control potentiometer. The Camera No. 2 output, line amplifier output and radio frequency detector output also pass through sets of monitoring networks in cascade. By means of groups of mechanically-interlocked push button switches indicated in vertical rows on Figure 7, the No. 1 monitor, No. 2 monitor, or the oscilloscope can be switched to any of these four outputs independently, the interlocks being arranged to prevent any monitor or the oscilloscope from being bridged across more than one network at a time. Only three input

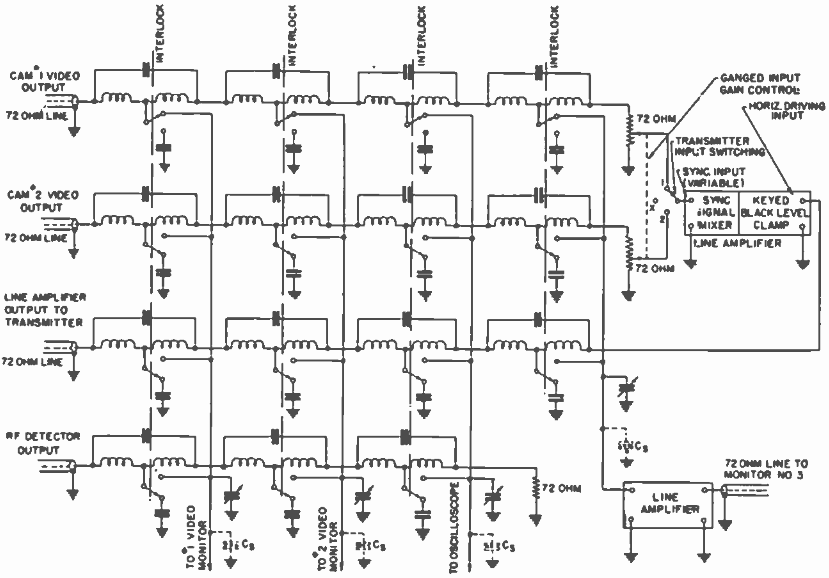

Fig. 7—Master control unit. (Simplified video block diagram.)

points are provided for No. 3 (utility) monitor as there is no occasion to use this monitor to look at the radio frequency output during operations. A line amplifier with bridging input and 72-ohm output is provided so that No. 3 monitor can be placed at any desired location in the aircraft and fed from a 72-ohm cable.

The operation of the monitoring filters is shown in more detail in Figure 8. It will be noted that during normal operation, with the switch in position *A*, the shunt capacitance consists of C_2 and C_3 in parallel. When the switch is thrown to position *B* to feed the video signal to a monitor, an equivalent capacitance to C_3 consisting of C_4 and C_5 in parallel replaces C_3. The capacitance C_5 represents the value of the stray capacitance in the connecting lead and input to the monitor, while C_4 is a trimmer capacitor adjusted to take care of different lengths of monitoring cables in various installations. This monitoring network is a constant resistance, bridged-tee type utilizing negative mutual inductance. Values which give constant resistance, approximately zero attenuation and linear phase are:

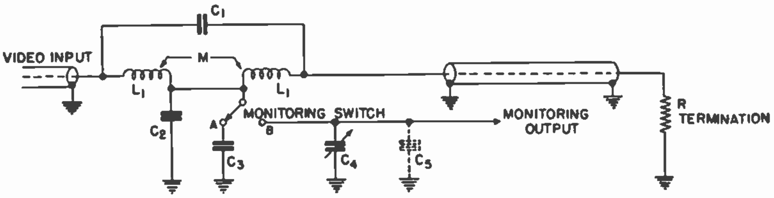

Fig. 8—Monitoring network.

$$L_1 = \frac{0.158R}{f}$$

$$M = \frac{-0.050R}{f}$$

$$C_1 = \frac{0.029}{Rf}$$

$$C_2 + C_3 = \frac{0.315}{Rf}$$

Where R = surge impedance of the cable

$f =$ top video frequency component considered
M, C_1, C_2 and C_3 as in Figure 8.

A further requirement is that the shunt resistance of the monitoring tap be high compared to the surge impedance of the line. This requirement is easily met in the usual monitoring input which has a direct-current resistance of 250,000 ohms or more and an equivalent shunt input capacitance, including several feet of coaxial input cable, of 100 micro-microfarads or less. This type of monitoring filter has been successfully used in the commercial television operations of NBC for the past several years. It has economical advantages over the use of a low-impedance video bus for monitoring purposes. It can also be designed to fit in a small space, the overall dimensions of the shield housing for the filters of Figure 7 being 1½ x 6 x 12 inches approximately.

As indicated on Figure 7, the line amplifier feeding the transmitter also incorporates a keyed "black level clamp" circuit.[7] This circuit performs the function of maintaining the tips of the video blanking signals at a constant direct-current level at the grids of the cathode follower tubes feeding the transmission line. Low frequency surges caused by switching the input of the line amplifier from one camera to another are thereby ironed out and prevented from affecting the transmitter. While this is not a new idea, it will be described here in some detail since it is also used in the television cameras to discriminate against microphonic disturbances.

Figure 9 shows the basic circuit of the keyed black level clamp. In operation, the diodes D_1 and D_2 of V_3 are periodically rendered conducting by the application of keying pulses supplied by V_2. The keying pulses are of line frequency and are timed to occur during the "back porch" portion of the video blanking pedestals. Application, simultaneously, of positive pulses to the plate of D_1 and negative pulses to the cathode of D_2 provides a low resistance discharge path for the coupling capacitor C_1. The potential on the grid side of C_1 is therefore brought back to a fixed value ($-C$ in this case) at the end of each scanning line. By adjustment of R_3 and R_4 to provide keying pulses of equal amplitude and some adjustment of the relative values of R_1 and R_2, the keying pulses can be balanced out so that only a small residual pulse signal appears at the grid of V_4. Provided that the peak-to-peak video signal at the point where the clamp is applied amounts to several volts, the residual unbalanced pulse signal can be made so small as to have negligible effect on the output of V_4.

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[7] K. R. Wendt, U. S. Patent No. 2,299,945, October, 1942.

CAMERA DESIGN

As already indicated, two types of television cameras were designed for Project Ring. The PH-536/AXS-1 camera utilizes the Image Orthicon,[8] a super-sensitive pickup tube producing a satisfactory picture with 1/100th the light required by the iconoscope. The PH-537/AXS-1 camera utilizes a developmental type orthicon similar to the type 1840, but incorporating an electron multiplier signal amplifier. The sensitivity of this tube is intermediate between the iconoscope and the Image Orthicon. The use of an electron signal multiplier, with the resulting increase in effective sensitivity, makes this tube useable over a greater dynamic light range than is possible with the type 1840.

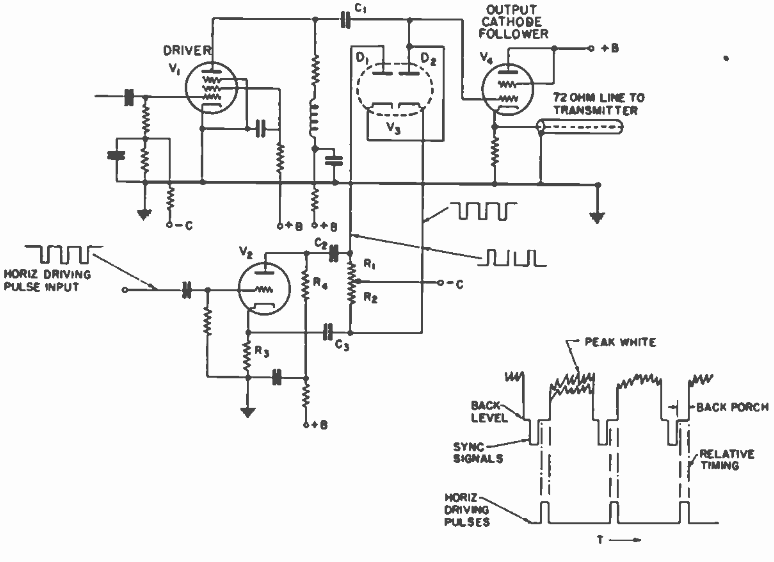

Fig. 9—Keyed black-level clamp circuit.

Photographs of these two camera types are shown in Figures 10 and 11.

Functionally, these cameras differ from the usual commercial designs in that they incorporate within a single housing all the tubes and components needed to produce a complete video signal (except the synchronizing signal generator). Their video outputs have a peak-to-peak voltage level of 1.0 volt into a 72-ohm load and are complete with video blanking pedestals. The chief advantage of this arrangement is that there is a saving in overall weight of the equipment as compared

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[8] Albert Rose, Paul K. Weimer and Harold B. Law, "The Image Orthicon—A Sensitive Television Pickup Tube", *Proc. I.R.E.*, Vol. 34, No. 7, pp. 424-432, July, 1946.

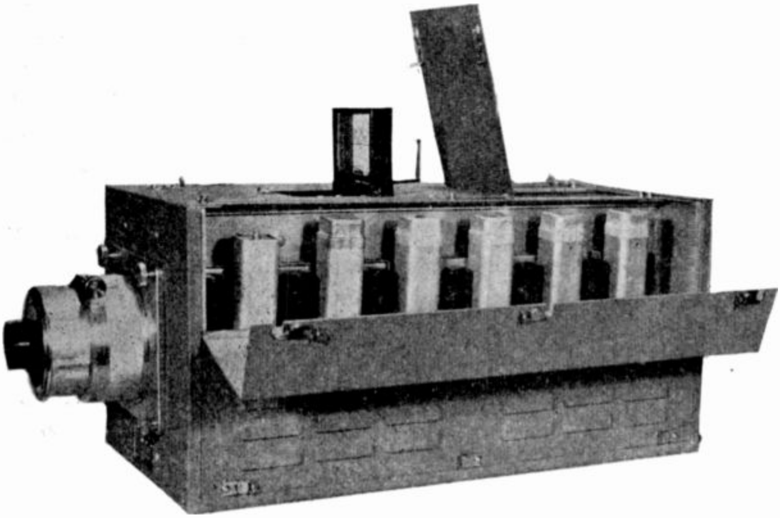


Fig. 10—Image orthicon camera with viewfinder. (Hinged cover opened to show video amplifier—top access door open.)

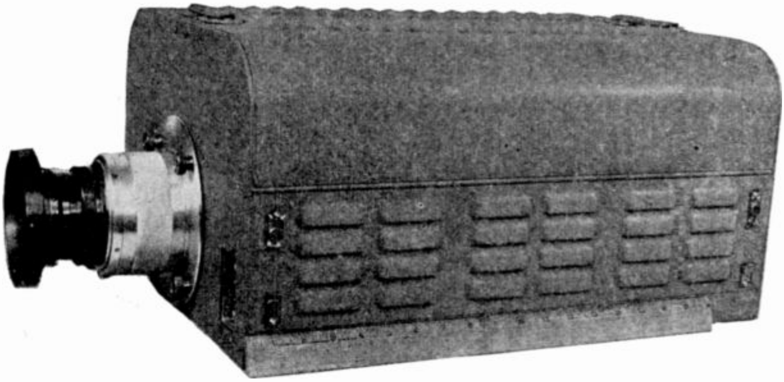


Fig. 11—Multiplier Orthicon camera with 7½ inch f/2.5 lens.

to the commercial practice of locating camera deflection amplifiers and the main video amplifier remote from the camera proper. The image orthicon camera is 10⅜ x 9¾ x 21¼ inches in size and weighs 46 pounds, less lens and lens mount. The camera utilizes 21 tubes, of which nine are dual types. The Multiplier Orthicon camera is 10¼ x 15½ x 20¾ inches in size and weighs 74 pounds. This camera utilizes 21 tubes, of which eight are dual types.

Since the subjects to be televised are always at a distance, in normal operation, the camera lenses are locked at infinity focus for this application. A simple viewfinder such as the combination ball sight and plano-convex lens with cross-hairs, as shown in Figure 10, may be used. A cathode-ray-tube viewfinder using a green zinc-orthosilicate fluorescent screen with a green optical filter has also been found to be convenient for airborne television use. The No. 3 (utility) video monitor has been used for this purpose in connection with the nose position camera where operating space is restricted.

In addition to special selection of non-microphonic tubes for the first video amplifier stages and careful design of the vibration isolators on which the cameras are mounted, keyed black level clamp circuits are used to eliminate microphonic signals from the camera video outputs. The early video amplifier stages of both types of cameras are purposely designed to have a drooping gain characteristic for signals below 500 cycles. This is followed in each case by a stage incorporating a black level clamp at a point where the video signal is at a sufficiently high level so that microphonic disturbances produced by subsequent stages are negligible. Video stages following the black level clamp, which effectively "restores" the low frequency components of the signal, are designed for uniform response from 20 cycles up to the top video frequency.

Another circuit artifice which counteracts microphonics is the use of a value of signal load resistor which is high compared to the effective shunt capacitance existing at the output of the pickup tube electron multiplier. The resulting loss in video signal output and the phase shift at high frequencies is compensated for in one of the higher level video stages by use of a "high-peaker" circuit. This results in a greater video signal input to the thermionic amplifier at low and medium frequencies than if the load resistor were chosen for more uniform response over the entire video band. A form of "high-peaker" compensation used in the Ring equipment is shown diagrammatically in Figure 12. Since both the electron multiplier and the amplifier tube V_1 can be considered as constant-current generators, and since the gain versus frequency characteristic of the intermediate video stages is uniform, the overall response of the system is proportional to the product of the impedances of the reactance arms consisting of R_1 and C_1 in parallel and R_2 and L_2 in series. These reactance arms became inverse networks ($Z_1 Z_2 = A^2 =$ Constant) when the values of L_2 and R_2 are chosen so that:

$$R_1 C_1 = \frac{L_2}{R_2}$$

When this is done, the phase and amplitude distortion produced by the R_1C_1 combination at the input to the amplifier can be exactly compensated over a wide band of frequencies within the limitations imposed by resonance of L_2 with the interstage shunt capacitance C_2. As a practical matter, satisfactory compensation can be obtained by choosing L_2 small enough so that the parallel resonant frequency of L_2 and C_2 is well above the top video signal frequency of the system. A ratio of resonant frequency to top video frequency of 1.5:1 was found to be satisfactory.

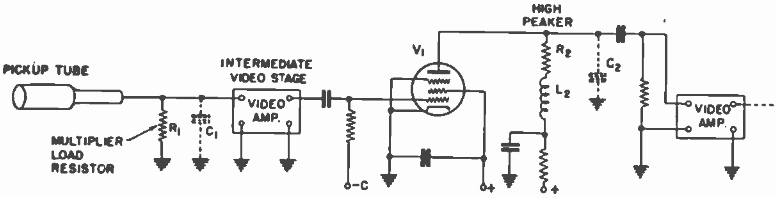


Fig. 12—Block diagram illustrating "high-peaker" correction network.

By use of such a compensating system, the orthicon low-frequency-signal output fed to the grid of the first thermionic amplifier tube can be increased many times over the value obtained with a flat amplifying system. This results in a worthwhile discrimination against microphonic disturbances originating in the thermionic amplifier portion of the video system.

AIRBORNE VIDEO TRANSMITTER

The video transmitter was based upon a preliminary design worked out for this project by RCA Laboratories Division. It was designed and constructed to provide for either of two types of installation. The low-power installation—100 watts average carrier or 400 watts peak carrier output—is comprised of the transmitter unit, transmitter power supply unit (units on left side of Figure 13), remote control box, modulation indicator unit, antenna coupling unit, antenna, shock mounts, wavemeter, test meter and cables. The installed weight is 200 pounds and the power required is 14 amperes at 28 volts direct current and 14 amperes at 115 volts 400/2400 cycle alternating current. Cooling air at the rate of 200 cubic feet per minute is required. The high power installation—350 watts average carrier or 1400 watts peak carrier output—consists of the video transmitter, power amplifier and two power supplies shown in Figure 13 plus the associated equipment, cables and antenna. The installed weight is 400 pounds (30-pound cathode-ray oscilloscope not included). The power required is 40 amperes at 28 volts direct current and 45 amperes at 115 volts 400/2400 cycle alter-

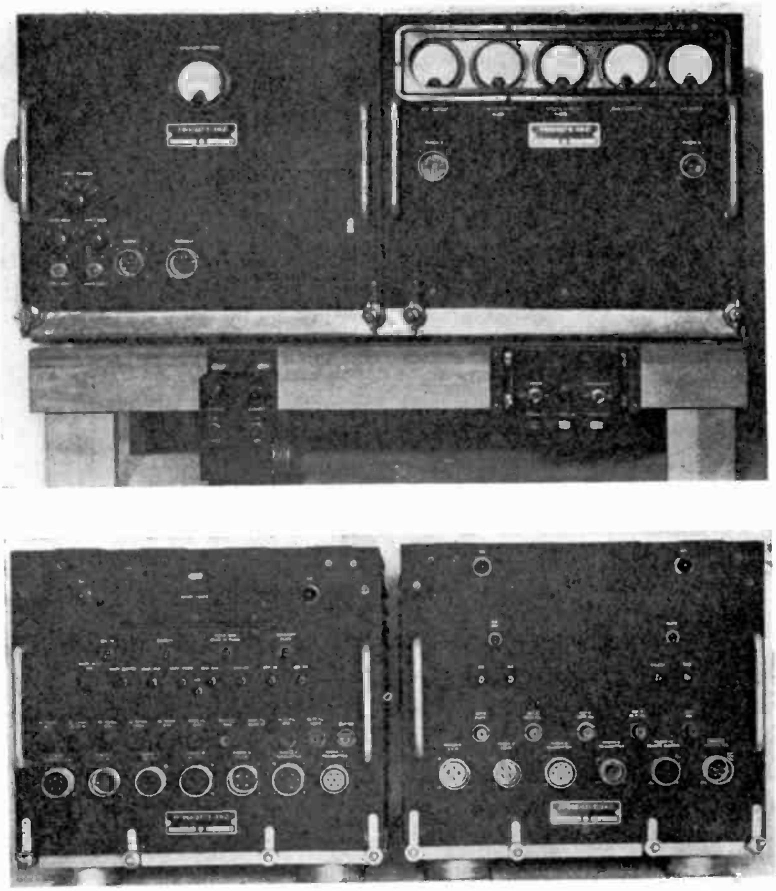


Fig. 13—The video transmitter and linear power amplifier. (These are shown at the top, mounted together as a unit on a special shock mounting. The remote control box and modulation indicator unit are shown mounted below the transmitter. The lower two units are the transmitter and power amplifier power supplies.)

nating current. Cooling air at the rate of 800 cubic feet per minute is required. The air for cooling the transmitter and power amplifier is supplied in flight by an air scoop mounted on the fuselage of the aircraft. An auxiliary cooling blower is used for ground testing. The high-power transmitter is shown in Figure 13.

The radio frequency section of the transmitter contains the following tube and circuit arrangements: A type 826 oscillator tube is operated in an ultra-audion (colpitts) type oscillator circuit, at a plate potential of 750 volts, at either 90 or 102 megacycles. The oscillator is very loosely coupled to two type 3E29 pentode tubes, connected in parallel push-pull, operating as a buffer stage. The buffer drives a grid-modulated stage through a four-terminal inductively-coupled band-pass circuit. The secondary of this network is heavily damped to provide a band-pass of 85 to 107 megacycles and also to improve the voltage regulation of the buffer stage. Provision of this wide bandpass eliminates the need for retuning for operation on either of the two carrier frequencies, 90 or 102 megacycles. Four type 4E27 pentode tubes, connected in parallel push-pull, are operated at a plate potential of 1100 volts as a grid-bias amplitude-modulated Class C stage. The output of this modulated Class C amplifier unit may be fed to an antenna system through an antenna coupling unit or may be used to drive the linear power amplifier unit used in the higher power installation.

The first stage of the video modulator section of the transmitter consists of two type 6AG7 pentodes with the plates connected in parallel. Input connections to each of the grids are provided so that synchronizing signals may be fed to one grid, and video signals to the other as required in some applications. When being fed mixed video and synchronizing signals from the pickup equipment described here, the two modulator inputs are operated in parallel. A video pre-emphasis network in the first video stage provides for the selection of several degrees of high-frequency peaking as may be required to compensate for sideband attenuation in the modulated radio-frequency section of the video transmitter. The second video stage consists of a type 3E29 pentode tube operating in a conventional video amplifier circuit. The video modulator stage contains two type 3E29 tubes operated in parallel. The video output signal of the modulator, approximately 120 volts peak-to-peak, is applied in series with the grid bias supply of the amplitude-modulated radio-frequency amplifier. High-frequency compensation is effected in the video stages by conventional means, i.e., combination (series shunt) peaking and LCR (inductance-capacitance-resistance) networks in the cathodes.

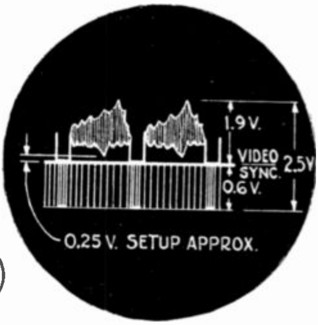
The Class B linear power amplifier unit contains two type 827R air-cooled screen-grid tubes connected in push-pull and operated at a plate potential of 2400 volts. The coupling to this stage is effected through a heavily damped four-terminal band-pass network. An LCR two-terminal network is inserted in series with the grid of each 827R tube. The combination of the tube impedance and this equalizing network provides a reasonably constant input impedance to the 827R stage between 85 and 107 megacycles. An LCR network is also placed in series with each filament lead of the 827R tubes. The capacitive reactances of these networks are adjusted to equalize the filament lead inductances over the range of 85 to 107 megacycles, thus effectively placing the filaments at radio frequency ground potential. The output of the power amplifier stage is inductively coupled to the 50 ohm transmission line and antenna. A 35Z5 diode tube connected as a detector and bridged across the transmission line provides a video monitoring voltage that can be used to check the modulation and overall performance of the transmitter. A direct-current milliammeter connected in the diode rectifier load circuit indicates the transmission line voltage or relative power output of the transmitter. Meters are also provided to read total grid, screen and plate currents of each tube of the transmitter.

The power supply for the low-power section of the transmitter contains five full-wave rectifiers and supplies the following direct-current voltages: 350 volts to the video modulator section, 350 volts to the 3E29 buffer stage, 1100 volts to the oscillator and 80 volts for bias. Xenon gas type 3B25 rectifiers are employed in all rectifier circuits except for the 5Z4 full wave high vacuum rectifier used for the bias supply. The Xenon gas rectifiers are preferred to the mercury vapor type for this application, since they will operate in any position and also over the wide temperature limits encountered in aircraft installations.

The ripple filtering circuits are conventional with the exception of the 1100-volt supply which contains a 50-ohm constant impedance network. This LCR network has constant impedance to all modulating frequencies up to 5 megacycles. The power supply unit also contains a thermal 40-second delay relay tube for timing the application of high voltages and the necessary fuses, interlock circuits, and contactors.

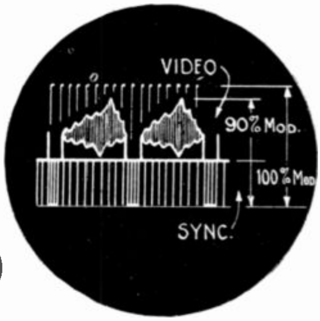
The power-amplifier power supply consists of eight 3B25 tubes connected in a bridge rectifier circuit and supplies 2400 volts for the plates and 1400 volts for the screens of the 827R tubes. A 50-ohm constant impedance network is used for filtering.

The modulation indicator unit, see Figure 13, is used in connection with a cathode-ray oscilloscope to give a visual indication of the percentage of modulation of the transmitter. The video signal from the radio-frequency-detector is fed to the modulation indicator unit and thence to the cathode-ray oscilloscope. The modulation indicator unit[9] contains a mechanical vibrator which short-circuits the detector load circuit at a rate of 300-400 times per second. The time of short-circuit of the video line (zero voltage) corresponds to zero power output of the transmitter. The tips of the synchronizing signals, as seen on the oscilloscope fed by the detector, correspond to maximum carrier amplitude, as the transmitter is operated with an input level such that the



14A

OUTPUT OF CONTROL UNIT LINE AMPLIFIER FEEDING TRANSMITTER AS SEEN ON OSCILLOSCOPE USING 20 CYCLE SWEEP



14B

MODULATION INDICATOR OUTPUT AS SEEN ON THE OSCILLOSCOPE USING 20 CYCLE SWEEP.

Fig. 14—Signal level and modulation check using oscilloscope.

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[9] 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.

tips of the pulses just begin to be compressed in the output of the modulated stage. The percentage of modulation of the carrier may therefore be quickly estimated visually on the oscilloscope screen as shown in Figure 14B. The dots at the top of the trace in Figure 14B occur when the contactor is closed and correspond to zero carrier while the tips of the synchronizing pulses at the bottom of the trace correspond to maximum carrier.

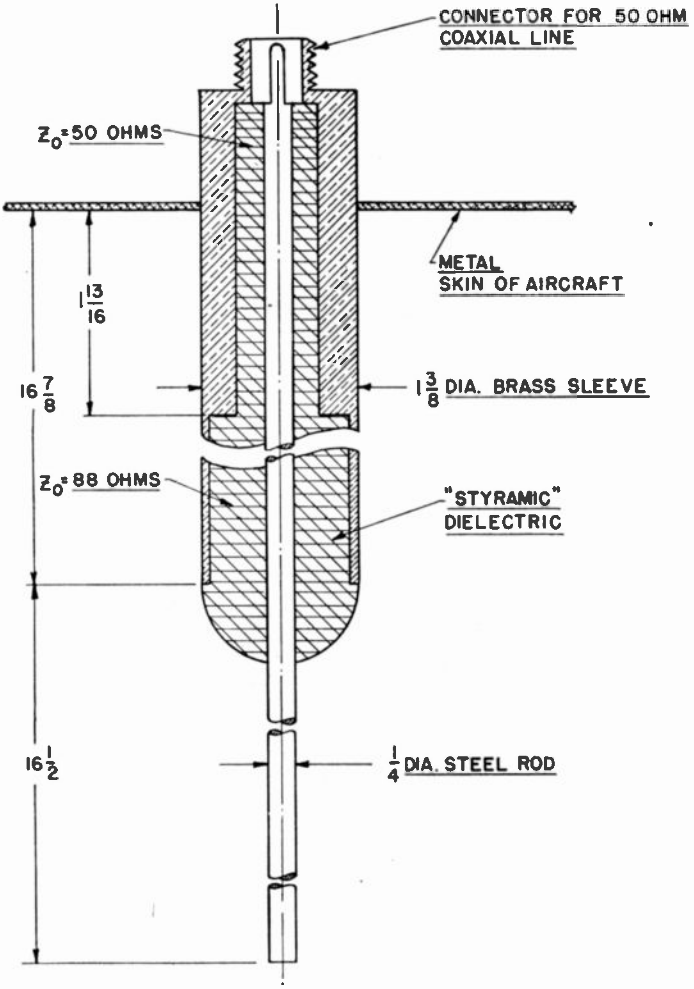
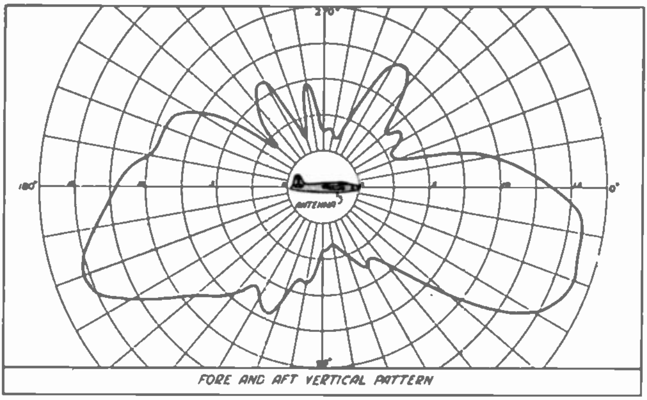
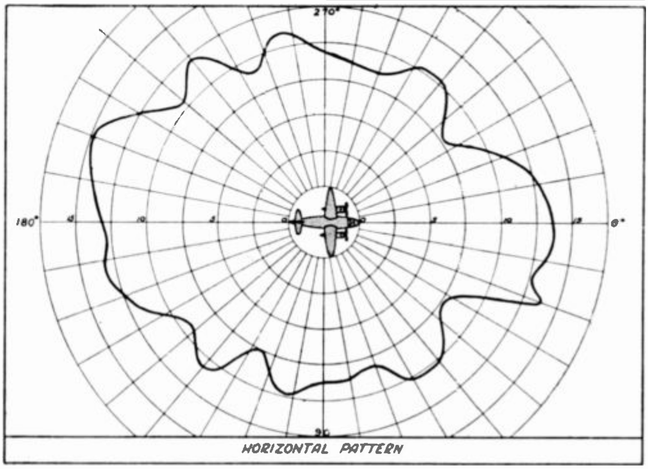


Fig. 15—Transmitting antenna.

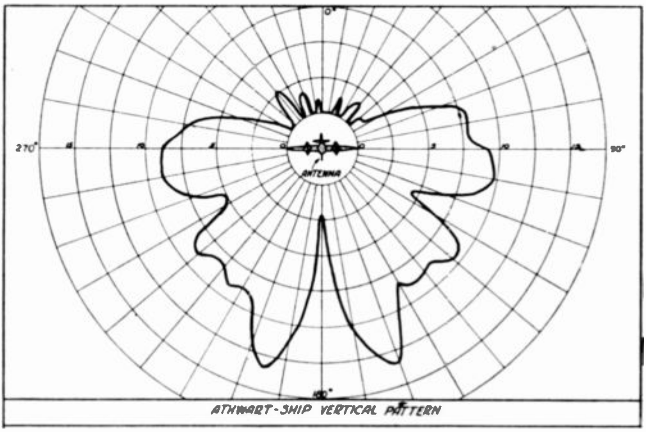
The transmitting antenna follows a design by Mr. P. S. Carter. This antenna, shown in Figure 15, provides satisfactory operation on carrier frequencies of 90 or 102 megacycles without readjustment.



FORE AND AFT VERTICAL PATTERN



HORIZONTAL PATTERN



ATHWART-SHIP VERTICAL PATTERN

Fig. 16—Measured radiation pattern of relative field intensities of a vertically polarized quarter-wave antenna mounted below forward bomb bay of JM-1 aircraft model, scaled 30 to 1 and measurements made at 3,000 megacycles.

The location of the antenna on the aircraft was determined on the basis of scale model tests performed by the engineers at Rocky Point, Long Island. Figure 16 shows a plot of data obtained in this way for the location finally chosen. As indicated, this provides a reasonably uniform horizontal radiation pattern.

AIRCRAFT INSTALLATION

The installation of the Ring equipment in a Navy JM-1 "Martin Marauder" aircraft, as used in developmental flight tests, is shown in Figures 17 and 18. The total installed weight including the two television cameras, control and monitoring equipment, the transmitter, video and transmitter power supplies, power and video cabling, cable junction boxes, equipment shock mountings, an assortment of camera lenses, spare tubes, meters, tools, etc. is 1400 pounds. These figures do not include the engine-driven generators and their voltage regulators.

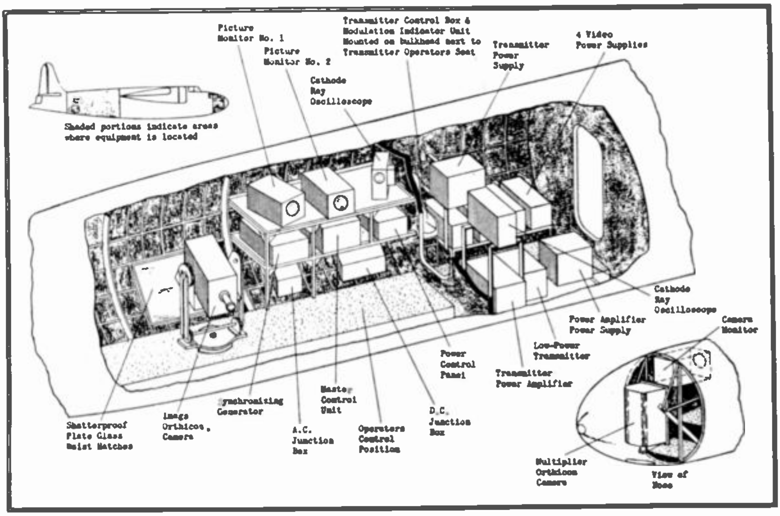


Fig. 17—Phantom view of the location of the various units in the JM-1 aircraft. The transmitter is located in the after bomb bay, and the antenna is located just forward of the transmitter beneath the fuselage.

Power for operating the equipment in flight is obtained from two Navy type NEA-8, 7.25 kilovolt-ampere 115-volt, 400- to 2400-cycle alternators, one mounted on each aircraft engine. Direct-current power at 28 volts for operating the video and transmitter filaments is obtained from two 200-ampere generators, one on each engine, operated in parallel. The power required for the video pickup and control equipment is 2.2 kilovolt-amperes at 115 volts 400 to 2400 cycles and 19 amperes at 28 volts direct current. The transmitter requires approximately 5 kilovolt-amperes at 115 volts 400 to 2400 cycles and 40 amperes at 28 volts direct current. The transmitter and video equipments are supplied from separate alternators. These alternators, operated on separate engines, are not synchronous but owing to adequate power supply filtering no trouble has been experienced from beats between the frequencies of the two power sources. At normal cruising speed, the alternator frequencies are in the vicinity of 1000 cycles.

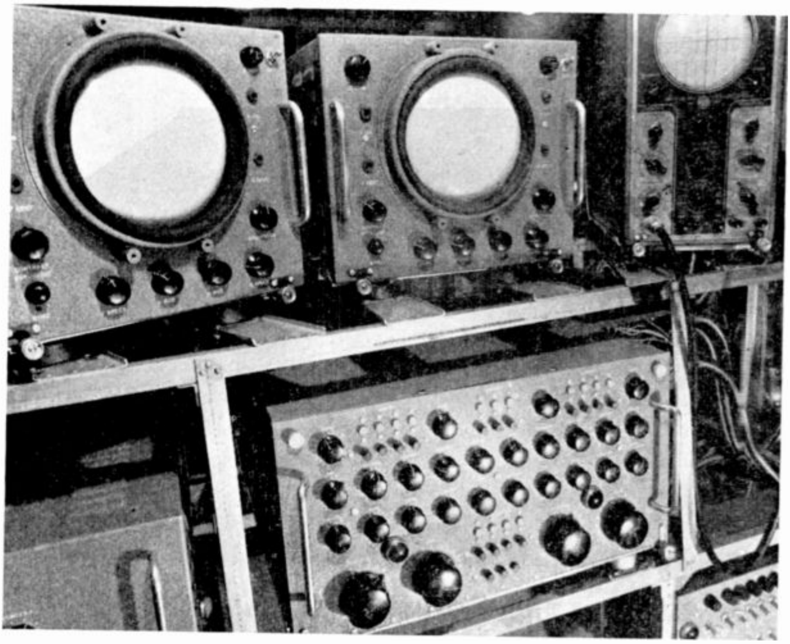


Fig. 18—Photograph of the master control position in the aircraft. (The picture monitors and cathode-ray oscilloscope (with calibration unit attached) are shown mounted above the synchronizing generator, master control unit and video switch panel. This position provides complete adjustment and control of the transmitted picture—including "on the air" monitoring of the picture, check of per cent of synchronizing signal transmitted and measurement of the modulation percentage of the transmitter.)

Power changeover switches and a power connection plug are provided for operation from an auxiliary power sources for pre-flight testing and maintenance when the aircraft is on the ground.

GROUND STATION EQUIPMENT

The R-90/UXR-2 television receiver shown in Figure 19 is continuously tunable over a range of 76 to 116 megacycles and accepts a video band extending 5 megacycles above and 5 megacycles below the carrier. The receiver sensitivity is such that an input signal of 35 microvolts with 40 per cent modulation produces an output signal with unity signal-to-noise ratio. The signal-to-noise ratio is expressed in terms of the ratio of peak-to-peak video signal output to peak-to-peak output noise. A half-wave vertical dipole antenna matched to a 72-ohm coaxial line is normally used to feed this receiver. While the design of this receiver follows conventional superheterodyne principles, it incorporates a very important feature in the form of a fast-acting automatic volume control operated from the peak value of the detected synchronizing signals. This automatic volume control serves to iron out the fluctuations in received signal which occur due to addition or cancellation of the direct wave from the airborne transmitter by waves received over an indirect path.

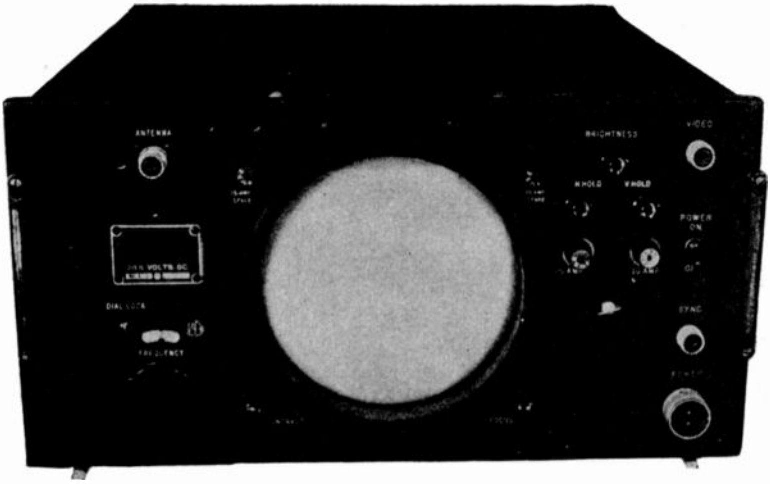


Fig. 19—Type R-90/UXR-2 television receiver.

The receiver provides an output of 1 volt peak-to-peak composite video signal at an impedance of 72 ohms for feeding one or more of the 12-inch viewing monitors shown in Figure 20.

TEST RESULTS

Numerous flight tests of the Ring equipment installed in a JM-1 aircraft were conducted at the Naval Air Station, Willow Grove, Pennsylvania and at the Naval Air Test Center, Patuxent River, Maryland, between November 1944 and July 1945 when Navy altitude and acceptance tests were completed.

In one test to determine the range of the transmitter, an acceptable television picture was received over a path length of 205 miles with the plane carrying the transmitter flying at 23,000 feet over Putnam, Connecticut and the receiver located at the Naval Air Station at Willow Grove, Pennsylvania. A vertical half-wave dipole at a height of 50 feet above the ground was used at the receiving end. A monoscope was used as a video signal source in the aircraft for this test so that the maximum transmission range could be determined independently of conditions of weather and visibility.

On another occasion with the aircraft flying over Philadelphia at an altitude of 10,000 feet, observers at the Naval Air Test Center at Patuxent River, Md., 120 miles away, were provided with views of the city buildings, railroad yards, ships in the Delaware River and other objects of interest. Using a 20 inch f/10 lens on the Image Orthicon camera, which provides a viewing angle of only 3.4 degrees, telephoto views of the Philadelphia-Camden Bridge showing moving traffic in sufficient detail to distinguish between automobiles, trucks and buses were received at Patuxent. Alternate overall views provided by the

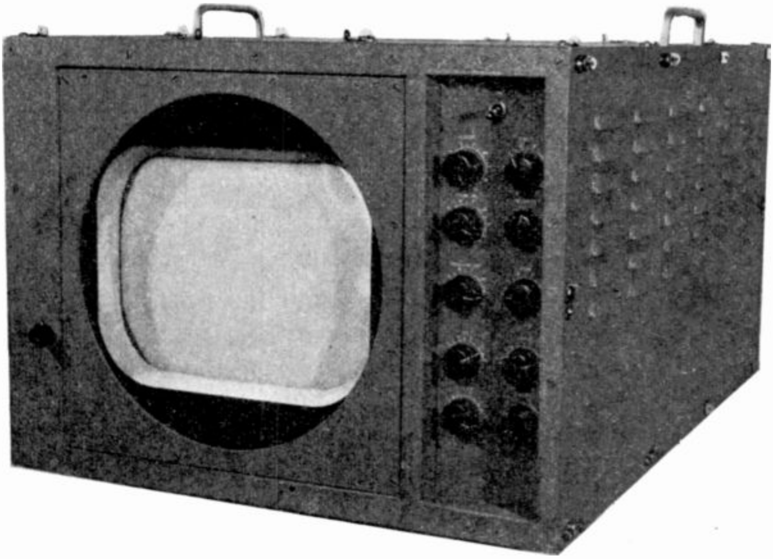


Fig. 20—Type ID-86/UXR-2 12-inch viewing monitor.

Multiplier Orthicon camera equipped with a 7.5 inch f/2.5 lens (17.5 degree viewing angle) were also transmitted. After circling the Philadelphia-Camden area, the aircraft proceeded to Washington, D. C., via Baltimore, Maryland, providing television views of ships, oil refineries, etc., arriving over the Capitol at an altitude of 17,500 feet. Using the Image Orthicon tube with the 20 inch f/10 lens, at this altitude, moving trucks, buses and automobiles were easily distinguishable. Circling down to 5,000 feet over Washington views of the National Airport and the Naval Air Station at Anacostia were transmitted to Patuxent, a distance of 55 miles, in clear enough detail so that expert observers were able to tell the number and types of planes parked on the airfields.

The altitude from which objects on the ground can be picked up by the television cameras is naturally controlled by conditions of illumination and visibility. The extent to which weather conditions govern the operating altitude is similar to that existing in the case of aerial photography. As in aerial photography, various optical filters, such as the Wratten No. 23, have been found helpful in improving picture contrast in the presence of ground haze. The type LM-15 Image Orthicon tubes used in these tests had considerable sensitivity in the near infra-red light region. The use of a Wratten No. 89A infra-red filter produced improved results with this tube in dealing with ground haze under some conditions.

ACKNOWLEDGMENT

The design, development and testing of the Ring equipment was the cooperative effort of a number of NBC television engineers under the direction of Mr. O. B. Hanson, Vice-President and Chief Engineer, Mr. R. E. Shelby, Director of Technical Development and Mr. G. M. Nixon, Assistant Director of Technical Development. Among the engineers taking an active part in the work were Messrs. R. M. Fraser, C. L. Townsend, Edward Wade, E. Stolzenberger, R. A. Monfort, W. L. States, E. C. Wilbur, A. W. Protzman, C. W. Turner and W. C. Resides. Preliminary transmitter design drawings were supplied by Mr. T. L. Gottier of RCA Laboratories Division; final transmitter development and construction were handled by Messrs. T. J. Buzalski, W. L. States and A. L. Hammerschmidt of NBC. Mr. F. J. Somers was the project engineer on design, development and engineering of the Ring equipment. Mr. H. P. See was in charge of installation, field testing and technical liaison with the Navy Department. In addition to the participation of the engineers already mentioned, substantial assistance was obtained from a number of groups and individuals in various

divisions of RCA. Administration of the several contracts under which the project was carried out was by the Government Development Section of RCA Victor Division; these contracts were NXss-20596, NXsr-47375 and NXsr-66811 between Radio Corporation of America and the U. S. Navy. Lt. Comdr. L. R. Moffett, USNR, was in active charge of the project for the Bureau of Ships of the Navy Department.

MINIATURE AIRBORNE TELEVISION EQUIPMENT*†

BY

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Summary—A developmental television camera, designed especially for airborne applications and using the image orthicon, is described. This camera is part of a complete airborne television transmitter system weighing 50 pounds. The transmitter has a power output of eight watts in the 260-to-380-megacycle range. Experimental results in guiding a medium-angle bomb with the aid of the miniature equipment are given.

INTRODUCTION

DURING the course of the World War II, it became apparent that extended application for television might be found if the transmitting equipment could be made smaller and lighter than the then-existing field equipment. One particular application for a miniature television equipment was a new medium-angle guided-bomb type of missile developed by the Douglas Aircraft Company, Inc., and known as the "Roc". The "Roc bird" is shown in Figure 1. After a preliminary study of miniature tubes and other components, it was decided to develop miniature television equipment for this project. The system was to consist of a small cylindrical camera unit placed in the nose of

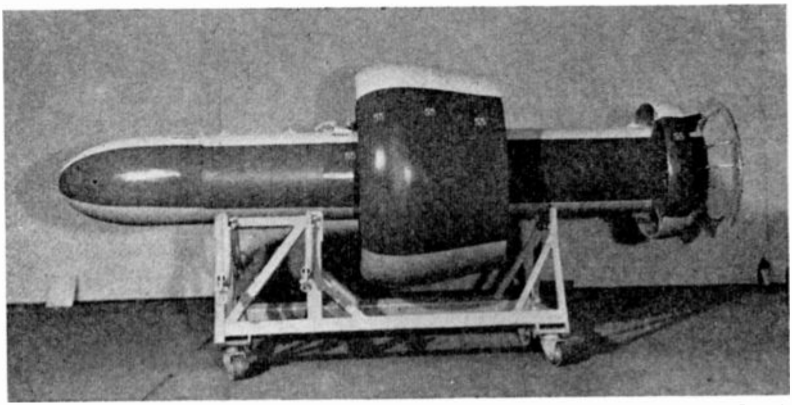

Fig. 1—The "Roc bird" (Courtesy of Douglas Aircraft Company, Inc.)

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* Decimal Classification: R583 × R560.

† Reprinted from *RCA REVIEW*, September, 1946.

the missile, a small transmitter and power supply placed in the after part, and a dipole antenna placed on the rear of the missile.

Some of the preliminary work was based upon tube operating conditions so severe that the life of the tubes would be materially shortened. The object was to get maximum performance for a short time with minimum apparatus, since, in use, the entire unit was expended after a few minutes of service. This design consideration was applied in the case of numerous expendable electronic apparatus, but due to the complexity of the equipment, it was decided that the advantage of the slight weight reduction, obtainable by severely overworking the tubes, was offset by the disadvantages of frequent tube replacements during development and the difficulty of adjusting and testing the equipments with the identical tubes to be used in service.

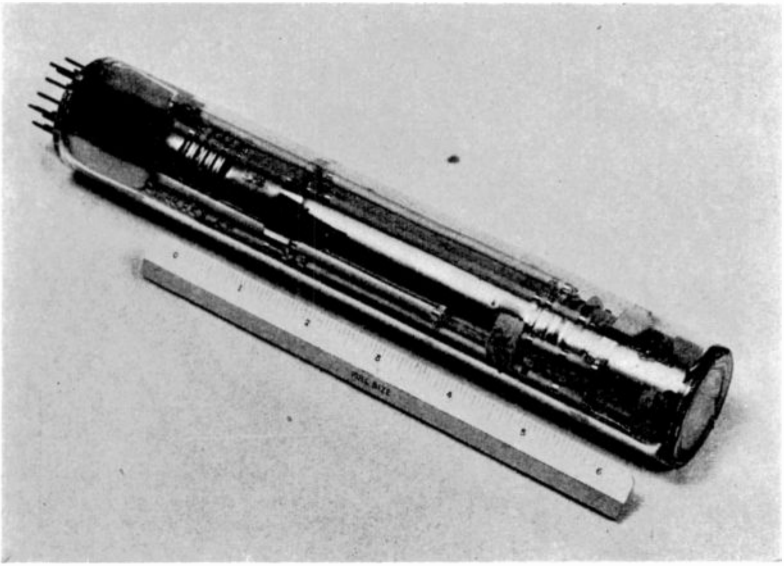

Fig. 2—The Mimo tube.

THE MINIATURE IMAGE ORTHICON

Since the space for the television camera unit was limited, a special pickup tube was developed for this project.[1] The tube is shown in Figure 2. It is called the "Mimo" tube (miniature image orthicon). The tube is 1½″ in diameter and 9″ in length.

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[1] P. K. Weimer, H. B. Law, and S. V. Forgue, "Mimo—Miniature Image Orthicon", *RCA REVIEW*, Vol. VII, No. 3, pp. 358-366, Sept., 1946.

THE CAMERA UNIT

The camera unit is shown at the right in Figure 3. An airborne iconoscope camera of earlier design is shown at the left for comparison of their sizes. The requirement for a cylindrical camera unit was met by mounting the tubes and components on three disc-shaped chassis surrounding the focusing coil and camera tube, as shown in Figure 4. The chassis, focusing coil, deflecting coils and alignment coils are assembled on and in a steel tube which also supports the lens mounting. The cylindrical case of the camera unit is so constructed as to be airtight, providing normal atmospheric pressure for the circuits regardless of altitude. The required controls pass through special vacuum-tight bushings in front of the camera unit case. The lens end of the case is covered with a flat disc of optical glass treated on both sides with non-reflecting film.

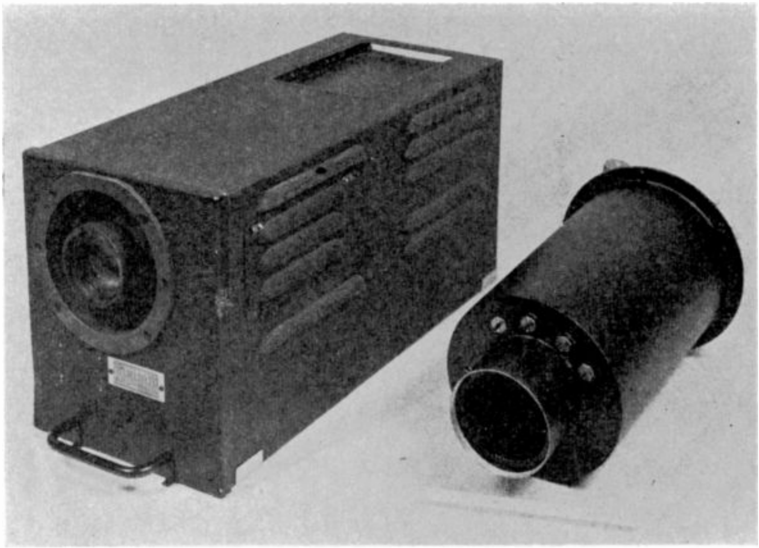

Fig. 3—The BC-1212 iconoscope camera and the Mimo camera.

The schematic diagram of the camera unit is shown in Figure 5. The video amplifier uses type 6AK5 tubes. It has a frequency characteristic which is approximately flat to 4 megacycles. The third video-amplifier stage has a conventional "high peaking" grid input circuit which compensates for the high-frequency attenuation of the orthicon output circuit. The grid of the fourth video-amplifier stage is "clamped" by a 6AL5 duo-diode to black level; thus the low-frequency components, lost in the small coupling capacitors preceding this stage, are reinserted.[2] Also, the clamping removes amplifier microphonics with the exception of those components near or above the line frequency of 14,000 cycles. Pulses for the clamp circuit are obtained from the balanced horizontal pulses appearing across the horizontal deflecting coils. Video blanking is added in the plate circuit of the fourth video stage. The cathode output stage acts as a clipper, setting a level of 0.3 volts, thus tending to limit the video output to 0.6 volts peak-to-peak.

The vertical deflection system, which operates at 40 cycles, consists of a 3A5 blocking oscillator and discharge tube (tube No. 8) and another 3A5 tube (No. 7) with both sections operating in parallel as the final amplifier. Blocking oscillators are used for driving both vertical and horizontal scanning circuits. The oscillator transformers are constructed with small mu-metal cores. The vertical speed of 40 cycles remains constant to approximately 1/10 cycle for supply voltage variations from 21 to 24 volts.

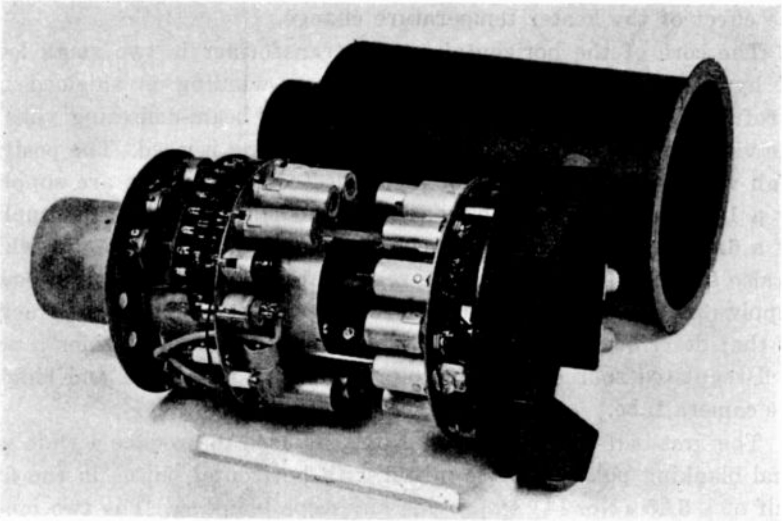

Fig. 4—Mimo camera with case removed.

The 14-kilocycle horizontal deflection circuit consists of a 3A5 dual-triode (tube No. 1) operating as a blocking oscillator and as a discharge tube, a 25L6 power pentode (tube No. 2) as the final amplifier and one-half of a 3A5 dual triode (tube No. 14) as a horizontal deflection regulator. This regulator, which uses the 50 volts across an NE-2 tube as a reference value, increases the plate voltage of the horizontal dis-

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[2] R. D. Kell and G. C. Sziklai, "Image Orthicon Camera", *RCA REVIEW*, Vol. VII, No. 1, pp. 67-76, March, 1946. (See page 70).

charge tube and thus tends to increase the deflection if the +335 voltage decreases. This cancels the reduction of deflection which would occur otherwise and provides substantially constant amplitude of horizontal deflection. This regulator also serves to keep the high voltages derived from the deflection-flyback voltage constant.

The horizontal oscillator has a regulated power supply, but its heater supply varies directly with the primary power sources. Under this condition the deflection frequency tends to increase with decreasing battery supply voltage. To counteract this effect a special compensating bias arrangement is used in the grid circuit. The grid leak is supplied with a positive voltage obtained partly from the 26-volt battery and partly from the regulated +150 volts. A drop in the +26 volts tends to cause the speed to decrease. The bleeder resistor for combining the two voltages is proportioned to make this effect cancel the effect of the heater temperature change.

The core of the horizontal output transformer is two small loops of hypersil core material. The secondary winding is shielded and carefully balanced to reduce pickup from the beam-deflecting yoke to the video input. Resistance-capacitance damping is used. The positive high voltages for the image-orthicon electron multipliers are supplied by a 1654 rectifier. The photo-cathode and ring voltages are supplied by a 6AK5 (tube (No. 4) in a constant-voltage rectifier circuit which is also self-regulating, using the regulated +150 volts from the power supply unit as a reference. This circuit operates on the same principle as that described for the large orthicon camera[2], and provides a very well regulated source of —350 volts for the photocathode and ring of the camera tube.

The first half of a 3A5 (tube No. 6) is used to produce a wide vertical blanking pulse. This is mixed with horizontal pulses in the first half of a 3A5 (No. 14) to provide kinescope blanking. The two pulses are also mixed in the second half of a 3A5 tube (No. 6) to provide orthicon blanking. Three volts of orthicon blanking are supplied to the target at a direct-current level variable from about —1½ to +1½ volts.

The video input circuit picks up an appreciable amount of undesired horizontal pulse voltage, mostly from the target blanking. This is partially neutralized by mounting near the input capacitor a terminal with a few volts of the opposite-polarity horizontal pulse on it, obtained from the clamp circuit. The remainder of the pickup is removed by blanking. The focusing coil and the alignment coil, connected effectively in parallel, are supplied with current from an Amperite regulator.

Horizontal and vertical sync (synchronizing) pulses are supplied

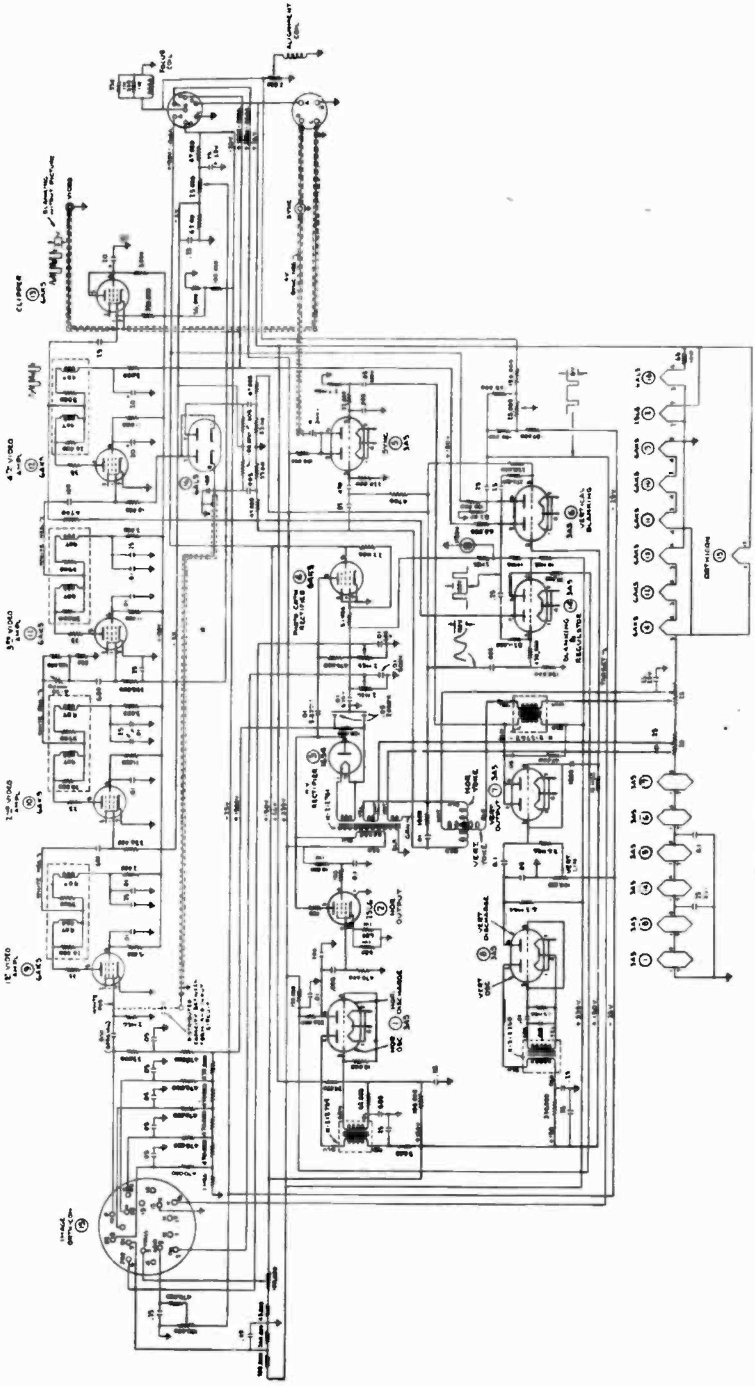


Fig. 5—Mimo camera diagram.

through the two halves of 3A5 tube No. 5. Both pulses are delayed in their respective grid circuits, providing a "front porch" on each blanking pulse. Since the plates, which are in parallel, operate at a low voltage, the tube tends to limit the horizontal pulses during the vertical sync pulse. Four volts peak-to-peak of sync signal appears on the output when it is coupled into a 150-ohm load.

The cathode of the image orthicon is at ground potential. Voltage for its bias adjustment is obtained from the negative supply of the dynamotor. The "wall" voltage for the tube is provided by the regulated +150-volt supply.

The camera unit has ten potentiometer adjustments. Four of these may be adjusted with the sealed cover in place. They are the image-orthicon grid bias, "wall" voltage or beam focus, photocathode or image focus, and video gain. The other adjustments are the two centering controls, the photocathode ring voltage, vertical linearity, alignment-coil current, and target voltage. No adjustments are provided for the deflection amplitudes and speeds, these being permanently adjusted during test.

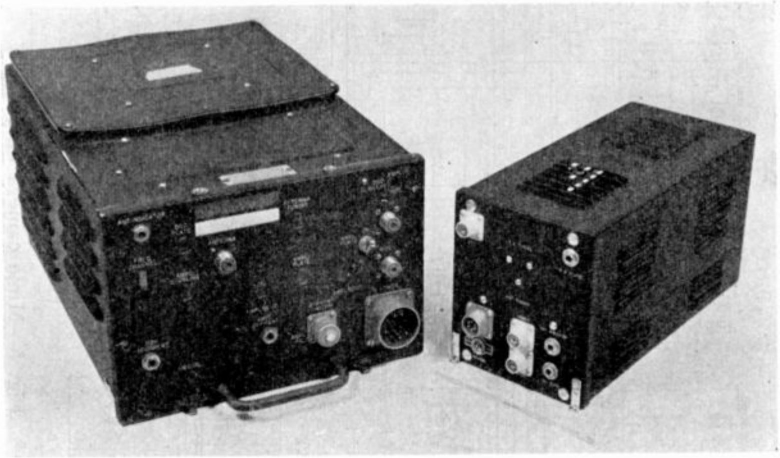


Fig. 6—The BC-1212-T3 and the Mimo transmitter.

THE TRANSMITTER

The Mimo transmitter unit is shown at the right in Figure 6. An earlier airborne television transmitter is also shown to indicate the relative size. The schematic diagram of the transmitter is shown in Figure 7. The transmitter is tunable between 260 and 380 megacycles and has a power output of 8 watts. Figure 8 is a side view of the transmitter, showing the relative location of the oscillator and the power amplifier. Figure 9 is a view of the bottom of the chassis, showing the master oscillator. Figure 10 shows the top of the chassis, with the power amplifier and modulator tubes.

The master oscillator is a 2C43 lighthouse tube, with a resonant-line tuning circuit and Colpitts feedback. The feedback capacitor is adjusted for the proper plate current, which may be measured across a 10-ohm resistor in the cathode. A tuned link circuit feeds the grid circuit of the push-pull power amplifier. The link circuit is fed through the chassis and it has a short-circuiting bar which is ganged with the short-circuiting capacitor of the oscillator. The power amplifier consists of two 2C43 lighthouse tubes in push-pull neutralized with a dual ceramic trimmer. The amplifier is grid modulated by two 6V6 beam tetrodes in parallel. The video and the sync signals coming from the camera unit are mixed at the grid of the first video amplifier (6AK5).

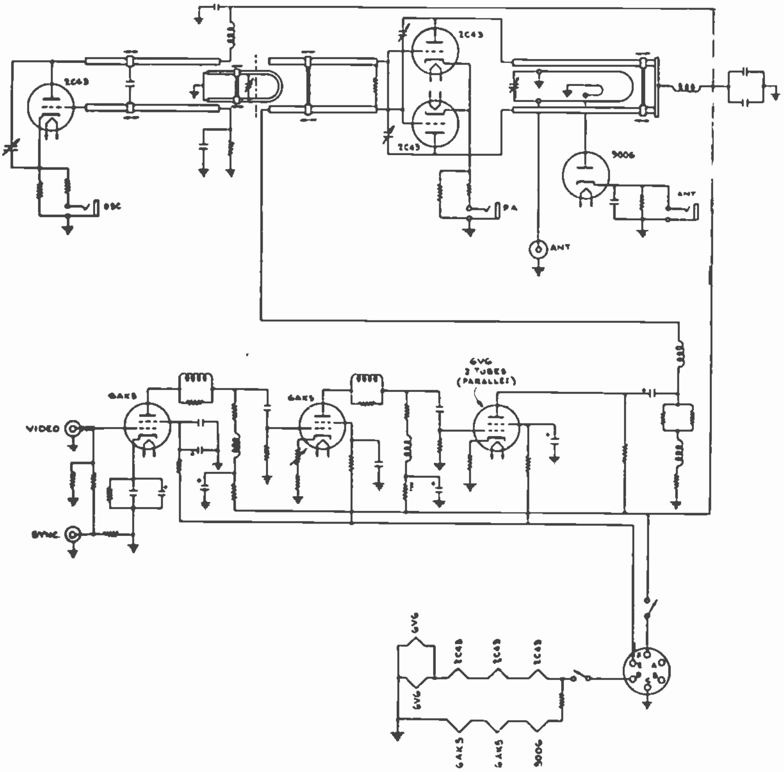


Fig. 7—Mimo transmitter diagram.

Each input terminal is terminated by a 150-ohm resistor and the 4-volt peak-to-peak sync signal is attenuated fifteen times before mixing. This circuit also attenuates the undesired flow of the video signal toward the camera, thus preventing cross-modulation which would cause parts of the picture signal to act as spurious sync pulses.

The second video amplifier has an "unbypassed" 1000-ohm variable cathode resistor acting as a gain control. With the gain control fully closed, the 0.7-volt peak-to-peak video input from the camera provides approximately 50 volts peak-to-peak across the plate lode of the 6V6 tubes, corresponding to a modulation in excess of 90 per cent. The video amplifier is flat out to 4 megacycles.

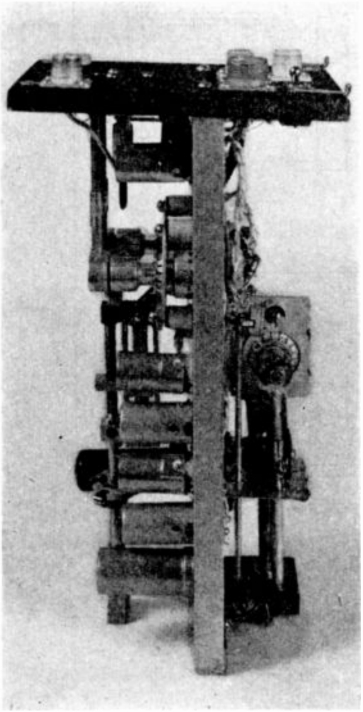


Fig. 8—Mimo transmitter chassis, side view.

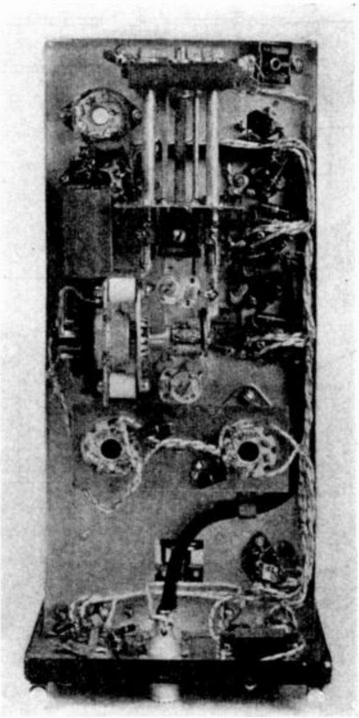


Fig. 9—Mimo transmitter chassis, bottom view.

A 9006 diode coupled to the antenna provides a monitor signal or a direct antenna tuning indication to a plug-in meter. The plate and grid currents of the power amplifier can also be measured by means of a plug-in meter.

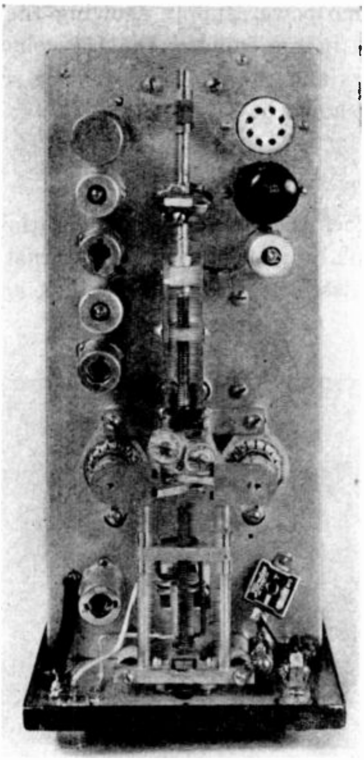


Fig. 10—Mimo transmitter chassis, top view.

POWER SUPPLY UNIT

The power supply is shown on the right-hand side of Figure 11. This may be compared in size with the dynamotor and junction box of the earlier airborne equipment in the same photograph. Figure 12

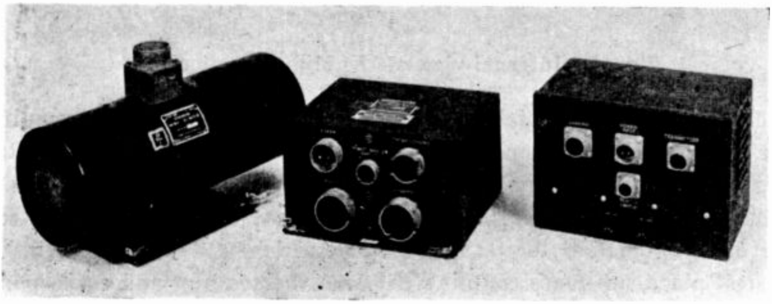


Fig. 11—The BC-1212-T3 dynamotor and junction box and the Mimo junction box including the dynamotor.

is a view of the Mimo power supply showing the Amperite regulator for the focusing coil in the camera, the 150-volt regulating tube, the electrolytic capacitors and the dynamotor. The schematic diagram of the power supply is shown in Figure 13.

ANTENNAS

The project involves the simultaneous operation of two radio links, namely, the Mimo link by which the picture signal is transmitted from the "Roc bird" or missile to the control plane, and the radio link by which the control signal is transmitted to the missile. This requires four antenna installations which will be referred to as Mimo-Roc, Mimo-plane, control-plane, and control-Roc, respectively.

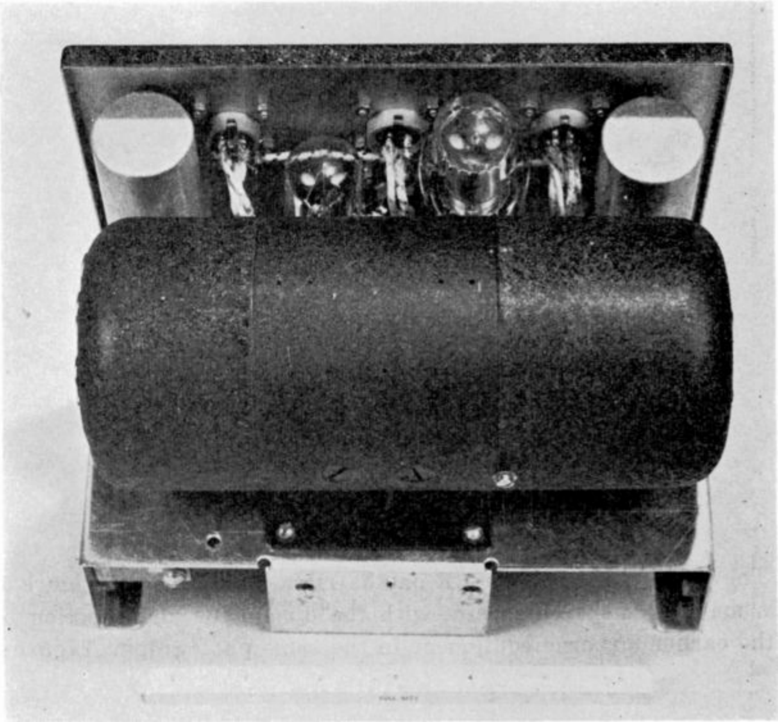


Fig. 12—Internal view of the Mimo junction box.

The desired characteristics of the antennas are prescribed by certain operational and tactical conditions which will be discussed. In the first place, adequate coupling between the sending and corresponding receiving antennas must be maintained for every likely position of the "Roc bird" with respect to the launching plane from the time of dropping until impact. Since the plane is always roughly at the rear of the bird, the Roc antennas should have maximum radiation in this general direction to provide a favorable signal-to-noise ratio. Also, these antennas should have negligible radiation in a forward or downward direction in order to avoid strong signals being reflected from the ground. This is particularly true of the Mimo antennas because a television picture is inherently very susceptible to multipath reception. Also, in this case, the effect of multipath reception would be made worse by the Doppler effect due to the high velocity of the bird.

Obviously the antennas on the plane should have their maximum radiation in a downward direction in order to bracket the Roc missile during its fall to earth. Also, adequate coupling between sending and receiving antennas should be retained even when the plane turns. It is not expected that the directivity of the plane antennas can be of any value in excluding multipath reception.

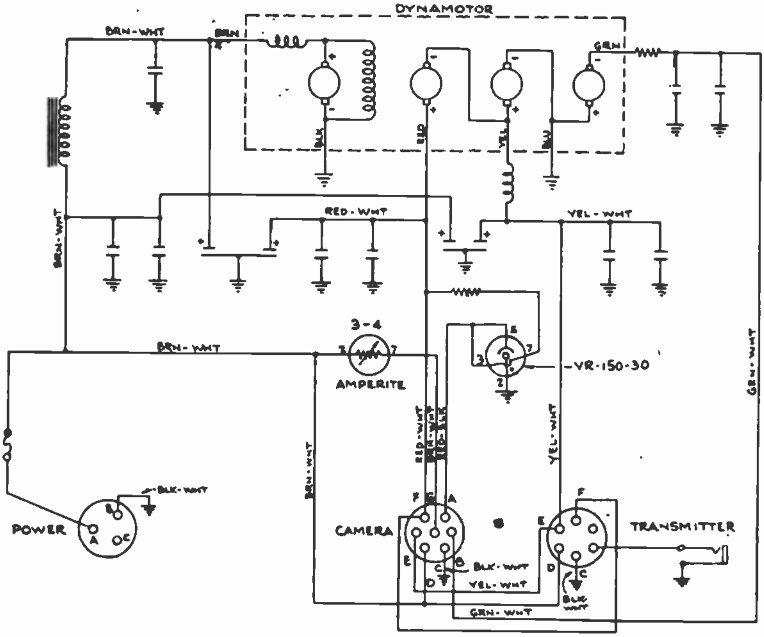


Fig. 13—Mimo power supply diagram.

Secondly, it was required that it be possible to adjust the Mimo antennas to match the impedance of the lines at any spot frequency in the band (260 to 380 megacycles) and have a several-megacycle bandwidth in order to transmit a picture of adequate detail.

Both the directivity requirements and the fidelity requirements are somewhat less severe for the control link than for the Mimo because the signal transmitted is of a simpler character. However, the control frequency (84 megacycles) is several times lower so that more space would normally be required to obtain a prescribed antenna performance. At the start, the required size of the control antenna presented a serious problem in the "Roc bird" because of the danger of interfering with the aerodynamic performance of the missile. After considerable study, a satisfactory solution of the problem was found by insulating the brake ring, which is primarily part of the aerodynamic equipment, and using it as the control-Roc antenna. This brake ring, however, affects the radiation pattern of the Mimo-Roc antenna. On this account it was found necessary to devise a method of effectively grounding the brake ring at the Mimo frequencies, and at the same time insulating it at the radio-control frequencies. Figure 14 shows the Mimo dipole transmitting antenna and the brake ring used for the receiving control antenna on the rear of the "Roc bird."

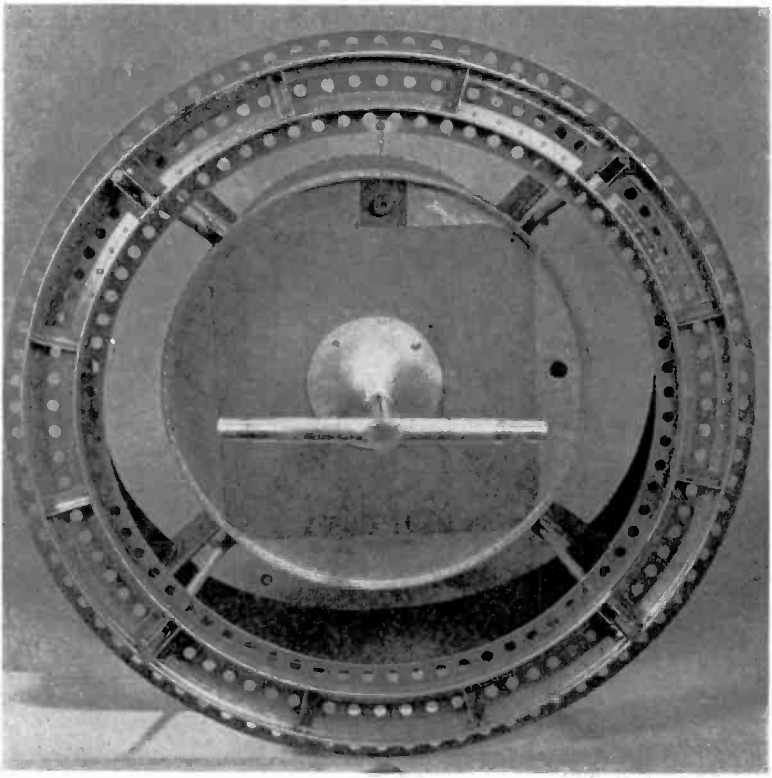


Fig. 14—The brake ring and dipole antennas on the "Roc bird."

The supporting rods of the brake ring are made up as coaxial lines which are a quarter wavelength long at the Mimo frequency, and thus ground the brake ring to the body of the "Roc." The impedance of these lines is high at the control frequency, and consequently insulates the brake ring from the body of the "Roc" at this lower frequency. Figure 15 shows the radiation pattern of this assembly in the elevation and azimuth planes. The reduction of secondary lobes from the Mimo antenna was of great importance in reducing the interference that reception of the radiated energy could cause in the control receiver. The brake ring is connected at a point 90 degrees with respect to the Mimo antenna, as shown in Figure 14, thus reducing the coupling between the two antennas to very low value.

The Mimo-plane antenna consists of two dipoles, identical in construction with those used on the "Roc." They are mounted on the underside of the plane with centers spaced 0.417 of a wavelength at mid-band, and oriented so that the extended axes of the dipoles intersect at 90 degrees at a point equidistant from the two dipoles. The physical layout is shown in Figure 16. The dipoles were fed equally and in phase.

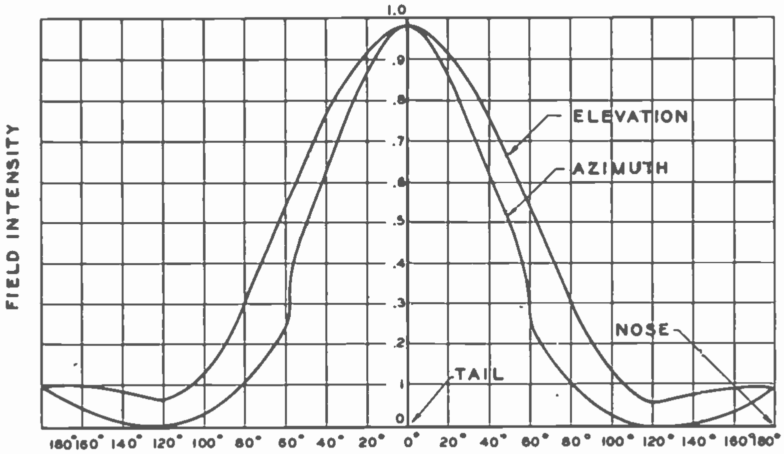


Fig. 15—Radiation patterns of the "Roc" dipole at 310 megacycles.

Comparison of Weights of Earlier Airborne Television and Mimo Equipments

While photographs give a fair indication of the size reduction accomplished in the Mimo design, it may be interesting to note that the weight of the total equipment was cut in half. The weights of the various components of the two systems are as follows:

	Earlier Airborne Television Equipment	Mimo 3
Camera Unit	33¼ lbs.	20 lbs.
Transmitter	26¾ lbs.	7 lbs.
Power Supply Unit	21½ lbs.	15 lbs.
Shock Mounts, Cables, etc., approx.	18½ lbs.	8 lbs.
Total	100 lbs.	50 lbs.

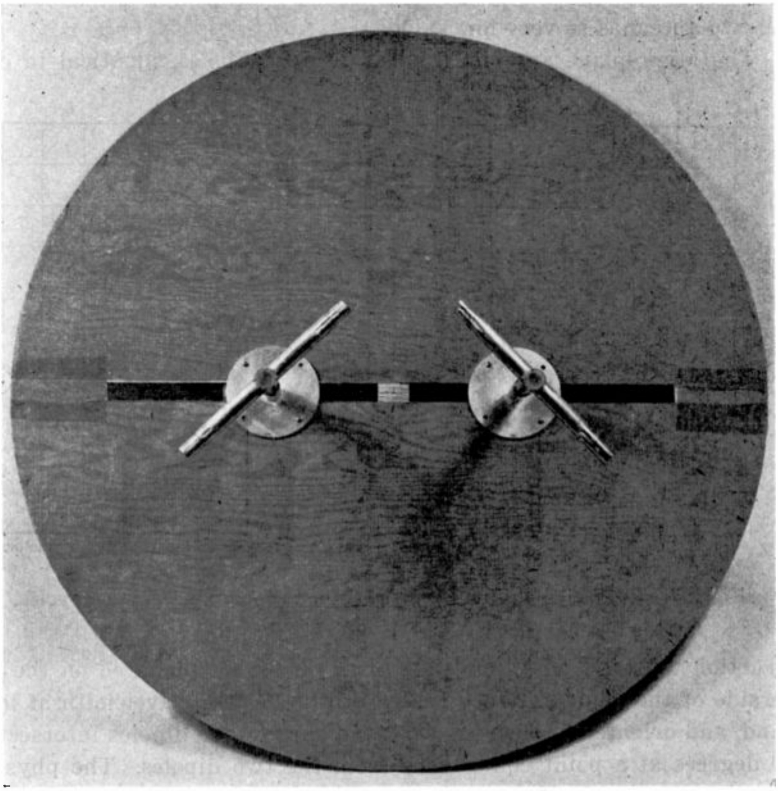


Fig. 16—Two dipoles as Mimo-plane antenna.

PERFORMANCE

In the early flight tests, made in the East, the camera and transmitting equipment was mounted in an AT-11 type aircraft. The signal was received on the ground. The camera unit was mounted in the plastic nose of the plane on a tiltable platform. The transmitter and power supply were in the rear of the plane and the antenna was mounted on the underside of the plane. A single horizontal dipole mounted on a wire screen was used as the receiving antenna at the ground station. On several occasions very satisfactory signals were obtained from a distance of 20 miles with the plane at 10,000 feet altitude. A useful signal was obtained at the same altitude even 40 miles away, when the nose of the plane was tipped down, thus taking greater advantage of the antenna directivity.

In actual drops of the "Roc" missile, of course there were no distances of such magnitudes involved. There were, however, several other difficulties encountered in obtaining good signals from the missile. The most important among these difficulties was a predominant Doppler effect, due to the multipath reflections from the ground (in spite of the directivity of the antenna) and the high speed of the "Roc."

When the source of a radiation is moving, an apparent wavelength will be observed according to the relation:

$$\lambda' = \frac{c \pm v_t}{f} \tag{1}$$

where λ' is the apparent wavelength, c is the velocity of propagation, v_t is the velocity of the radiating source, and f is the frequency. The sign is positive when the path increases, and it is negative with a reduced path. Since

$$f = \frac{c}{\lambda} \tag{2}$$

the frequency f' observed at the receiver is:

$$f' = \frac{c \pm v_r}{\lambda'} \tag{3}$$

where v_r is the velocity of the receiver. Substituting the value of λ' from (1), we have:

$$f' = \frac{c \pm v_r}{c \pm v_t} f \tag{4}$$

Assuming two straight paths, one increasing as the bird travels from the plane, and one reflected from the ground decreasing as the bird approaches the ground, two signals will be received producing a beat frequency according to the relation:

$$\Delta f = f_1' - f_2' = f \left(\frac{c + v_r}{c + v_t} - \frac{c + v_r}{c - v_t} \right)$$

$$\Delta f = f \frac{2c\, v_t}{c^2 - v_t^2}, \tag{5}$$

Since $c^2 >> v_t^2$,

$$\Delta f \approx f \frac{2v_t}{c}, \tag{6}$$

assuming a missile velocity of approximately 500 miles per hour, or 224 meters per second and a carrier frequency of 300 megacycles.

$$\Delta f = 448 \text{ cycles per second.}$$

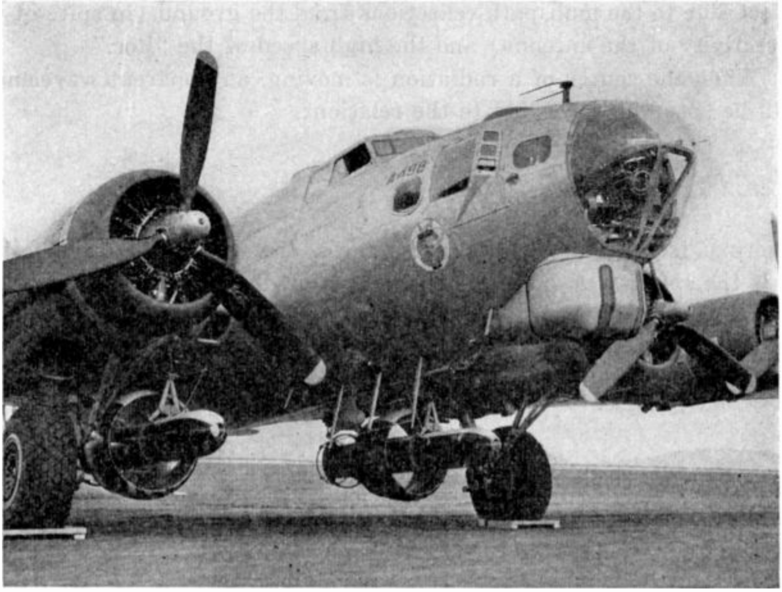


Fig. 17—Two "Roc birds" on a B-17 plane (Courtesy of Douglas Aircraft Company, Inc.)

Another difficulty encountered was the microphonics caused by the control surfaces. The microphonic problem, while considerably less serious than the Doppler effect, aggravated the situation. Since these two difficulties produced similar effects in the form of horizontal bars, the causes were not easily separated. In the course of actual drops these difficulties were reduced to the extent that very satisfactory pictures were obtained for guiding purposes.

Figure 17 shows two "Roc birds" attached to a B-17 plane. Figure 18 shows a photograph of a test target in the western United States.

Figures 19-21 are enlargements from 16-millimeter moving picture films taken from the television picture received at the control plane. In this particular drop the plane flew at 19,200 feet, which was 15,000 feet above the target area when the missile was dropped. The vertical bars caused by the Doppler effect are noticeable in all three pictures, particularly in Figure 21, which is taken less than a second before impact, but they do not destroy the value of the information. The reproductions through the 16-millimeter motion picture film enlarging and finally the printing process destroy much of the detail and clarity of the picture appearing on the monitor, but even from these reproductions, the value of the information for guiding the missile may easily be seen.

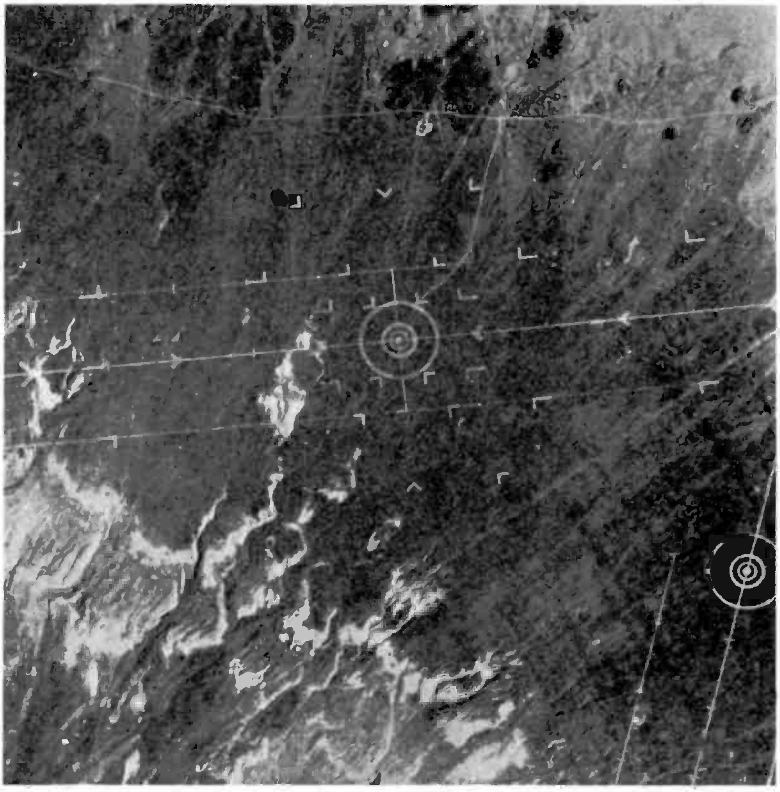
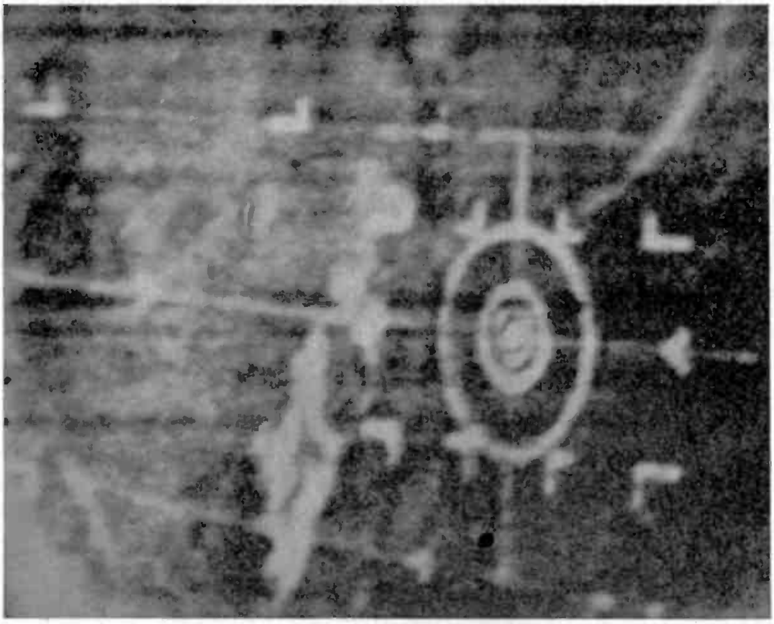


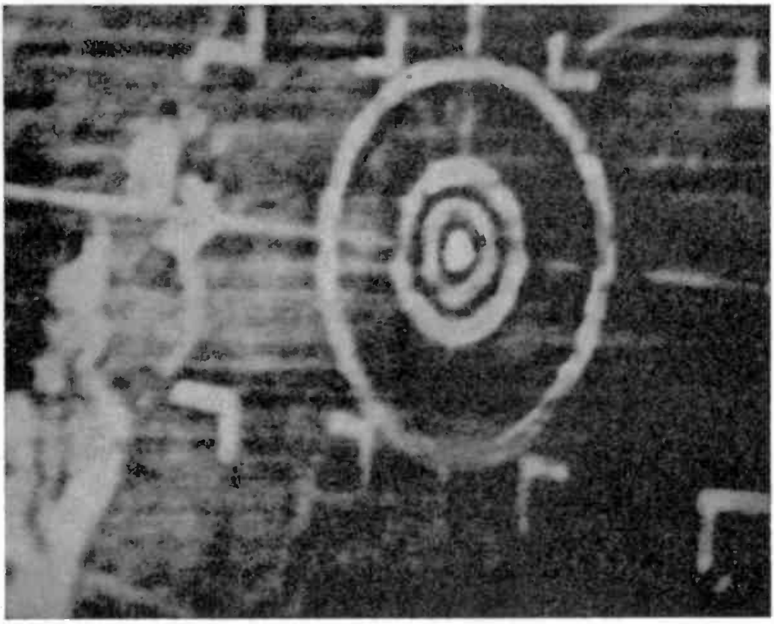
Fig. 18—Photograph of a test target.

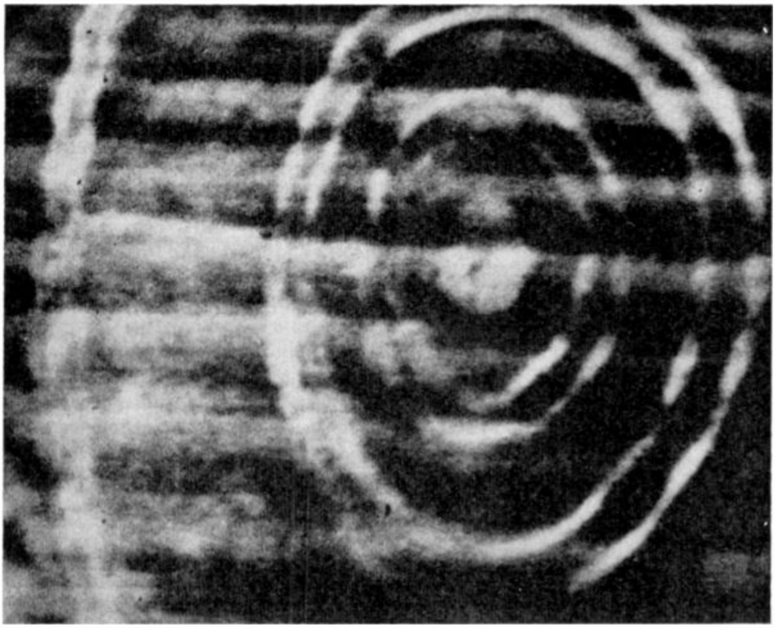
ACKNOWLEDGMENT

Acknowledgment is made to the engineers of the Douglas Aircraft Company, Inc., and RCA Laboratories Division who cooperated on the



Figs. 19-21—Progressive photographs of the monitored picture.





project. In particular, special credit is due Messrs. R. R. Thainer and K. R. Wendt, who helped in the design as well as the tests, to M. A. Jackson of NBC, who assisted in the drop tests, and to Dr. G. H. Brown and Mr. J. Epstein, who designed the antenna systems. The work described in this article was done in whole or in part for the Office of Scientific Research and Development under Contract OEMsr-441 with Radio Corporation of America.

MIMO-MINIATURE IMAGE ORTHICON*†

BY

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Summary—A miniature image orthicon, known as the "Mimo" tube, has been developed for use in airborne television equipment. Its reduced size and power requirements permit a substantial reduction in the dimensions and weight of the pickup-tube camera. The Mimo incorporates an improved mounting technique and employs additional fine mesh screens in front of the photocathode and target for the purpose of shaping the electric fields and simplifying operation. The resolution and signal-to-noise ratio of the Mimo approach that of the larger image orthicon at high light levels under carefully controlled conditions. Performance considerations as a function of the tube size are discussed.

INTRODUCTION

MILITARY applications for a miniature television camera have prompted the development of a pickup tube considerably smaller than any of the tubes in commercial use. At the same time, the aim was to approximate the performance of the larger tubes. Of all of the well-known types of pickup tubes, the image orthicon because of its high sensitivity and high signal level output was most suited for scaling down. Accordingly, a miniature image orthicon called the "Mimo" tube has been designed for use in airborne television equipment.[2] Figure 1 shows a comparative photograph of the Mimo tube and an image orthicon.#

The mechanism of operation of the Mimo tube is essentially the same as the image orthicon whose cross-sectional diagram is shown in Figure 2. The optical image is projected on the semi-transparent photocathode laid on the inside surface of the glass envelope. The resulting photoelectrons are focussed by the uniform magnetic field, and they land at high velocity on the thin, semi-conducting glass target. Since the secondary emission ratio of the glass is greater than unity, a positive charge pattern is built up on the glass corresponding to the light and

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* Decimal Classification: R583.6.

† Reprinted from *RCA REVIEW*, September, 1946.

[1] A. Rose, P. K. Weimer, and H. B. Law, "The Image Orthicon—A Sensitive Television Pick-up Tube", *Proc. I. R. E.*, Vol. 34, pp. 424-432, July, 1946.

[2] R. D. Kell and G. C. Sziklai, "Miniature Airborne Television Equipment", *RCA REVIEW*, Vol. VII, No. 3, pp. 338-357, Sept., 1946.

Throughout this paper the term "image orthicon" will refer only to the tube described in Reference 1.

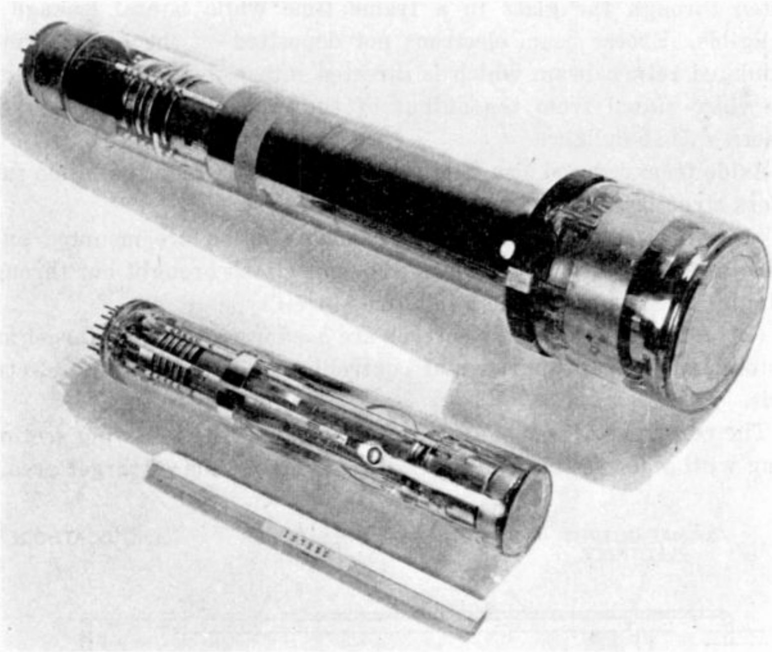

Fig. 1—Comparison of the Mimo tube with an image orthicon.

shade in the optical image. The target screen collects the secondary electrons from the glass and serves to limit the maximum potential to which the glass may rise. A low-velocity beam scans the other side of the glass target and deposits sufficient electrons in the positive areas to drive the glass down to the potential of the thermionic cathode of the gun. The conductivity of the glass is so chosen that the charge is con-

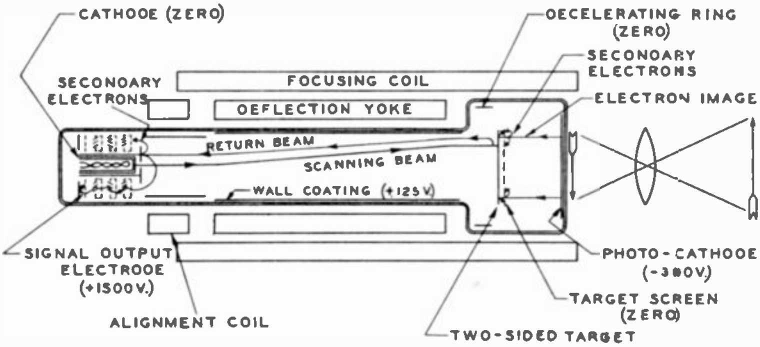

Fig. 2—Cross-sectional diagram of the image orthicon.

ducted through the glass in a frame time while lateral leakage is negligible. Excess beam electrons not deposited on the glass form a modulated return beam which is directed into a five stage multiplier. The video signal from the output of the multiplier is fed into the camera video amplifier.

Aside from reduced size, the principal ways in which the Mimo tube differs structurally from the image orthicon are:

(1) All electrodes including the image section are mounted on a single assembly with all the electrical connections brought out through a single 18-lead stem of the miniature button type.

(2) Additional fine mesh screens are used in front of the target and photocathode for the purpose of controlling the shape of the electric fields.

The results of these changes are described in the following sections along with a discussion of performance as a function of target area.

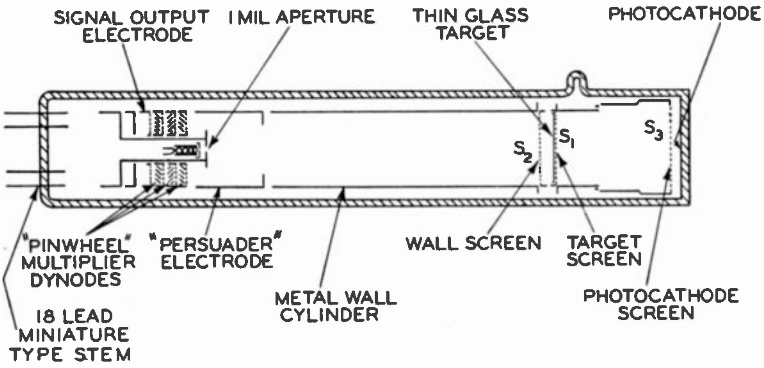

Fig. 3—Cross-sectional diagram of the Mimo tube.

Structural Details of the Mimo

A cross sectional drawing of the Mimo tube is shown in Figure 3. The overall length is 9″ (as compared to 15¼″ for the image orthicon) and the maximum diameter has been reduced from 3″ to 1½″. These dimensions allow a substantial reduction in weight of copper, and power required for the focusing and deflection coils, as well as the use of a smaller lens.

In the type of assembly used in the Mimo tube the metal wall cylinder is of thin nichrome and replaces the platinum coating used in the image orthicon. Ceramic tubing supports all electrodes (except within the gun) and the target connections are made to the stem by means of wire leads pushed through the ceramic tubing. Mandrels are used to align the cylinders during assembly, making possible a more accurate alignment than if the target structure were mounted independently as was done in previous tubes.

A gap of 180 degrees between two ceramics is left between the target structure and the wall for convenient insertion of the glass target and wall screen just prior to sealing. A spring contact to a metal button on the inside of the glass envelope connects the photocathode to the proper lead in the stem.

A glass envelope of uniform diameter is used for the Mimo tube in order to take advantage of the single unit construction without requiring the additional sealing operation over the target. This fact, combined with the elimination of the leads at the shoulder, greatly simplifies the glass blowing operations. The molded miniature type stem which requires no basing is extremely convenient.

It was found that the performance of the multiplier was unaffected by scaling it down from 1½″ to 1″ diameter. The persuader electrode is tied electrically to the first stage instead of being brought out on a separate connection as before. The gun is made slightly smaller in diameter and the defining aperture reduced to about 1 mil.

Vibration tests showed the Mimo to be structurally quite strong. One tube was found to be operable after having been subjected to an acceleration of 25 g's.

USE OF THE WALL AND PHOTOCATHODE SCREENS

The availability of fine mesh screens of high transmission and uniformity have made practical the use of screens for controlling the electric fields in front of the target and photocathode. These screens, labelled S_2 and S_3 in Figure 3, are mounted directly on the wall and target structures, replacing the separate "decelerating ring" and "photocathode ring" of the image orthicon. Unlike the target screen, S_1, they are positioned far enough from a nodal plane of the electron stream that their meshes are not superimposed on the transmitted picture.

One advantage of using screens in this manner is that good focus at the edges of the picture and freedom from distortion are automatically assured without requiring separate adjustment of ring voltages. Furthermore, the screens permit high fields in front of the target and photocathode without requiring high voltages. Uniform landing of the low-velocity beam at the edges of the target is easily obtained, and the position of the deflection coil for best landing is less critical.

Another important advantage gained by the use of the wall screen is the elimination of the multiplier shading control found in the image orthicon. The screen compels the electric field in front of the target to become more nearly parallel to the magnetic field. This reduces the translational effect on the beam which is the major cause of the scanning of the first stage of the multiplier by the return beam. The consequence of this reduction in scan is that the requirement of uniform gain of the first stage is somewhat less stringent. As a result there is no need to adjust the persuader voltage for controlling uniformity of gain. The persuader electrode of the Mimo is connected internally to the first stage lead.

It should be pointed out that the screens result in some loss in signal-to-noise ratio (in some cases as much as 30 per cent). Also, the two extra screens are potential sources of spurious signal. The wall screen is the more critical of the two because the beam passes through it twice. In spite of the fact that both the scanning beam and the return beam are out of focus when passing through this screen a spurious interference pattern simulating a mesh appears under certain conditions. This pattern can be minimized by proper spacing of the wall screen and the target.

In the airborne application for which the tube was designed, the advantage of the screens in simplifying operation considerably outweighed the accompanying disadvantages.

PERFORMANCE AS A FUNCTION OF SIZE

The active target area of the Mimo tube is slightly more than one quarter of that used by the image orthicon. This reduction affects performance from the standpoint of resolution, signal output and optics of the camera lens.

1. *Resolution*

Assuming that the resolution is limited electron-optically only by chromatic aberration and the stiffness of the beam at the target, it follows that a reduction in size of the tube, while keeping the voltage constant, should have no effect on the number of television lines which may be transmitted. The higher fields in the smaller tube should reduce the spot size in proportion to the change in dimensions. Actually, other less fundamental factors enter in to determine resolution — factors whose contributions are not as readily scaled down with tube size.

Loss of resolution by target leakage, for example, may arise from the volume conductivity of the glass or from the surface conductivity caused by a conducting coating of caesium on the glass. The first cause depends mainly on the resistivity and thickness of the glass and may be practically eliminated by using thinner glass in the small tube.

(Targets as thin as 0.05 to 0.1 mil were used in the Mimo tube). However, caesium leakage, when present, will deteriorate resolution to a greater extent with a target of small dimensions.

The superposition of the target screen upon the transmitted picture requires a finer mesh screen for the Mimo tube. An improvement in maximum resolution resulted when a screen of 500 meshes per linear inch was replaced by a screen of 1000 meshes per linear inch.

Cross talk in which the stray deflection fields disturb the paths of the photoelectrons in the image section might be expected to scale down proportionally with tube size. However, in the Mimo tube the deflection coil, for compactness, has been placed relatively closer to the target than in the image orthicon. This makes the cross talk a more critical problem. An effective solution is the use of iron wire wound over the deflection coil in combination with a cylindrical copper shield over the image section, but the position of the copper shield is quite critical. Alternative methods of reducing cross talk are an iron ring on the end of the deflection coil or a "bucking coil" over the image section fed by a small fraction of the horizontal deflection voltage. The cross talk from the horizontal deflection coil is more persistent than from the vertical coil. The shortened storage time of the image orthicon type of target, when the light is raised, rapidly erases the effect of "vertical" cross talk but has no effect on the "horizontal" cross talk until extremely high lights are reached.

The resolution required of the Mimo tube for the airborne television project was 250 lines at high lights, and this was easily met. (See Figure 4 and Figure 5). A number of tubes when carefully set up under experimental conditions with high light illumination showed more than 500 lines. The high limiting resolution of the scanning beam is evidenced by the fact that by under-scanning the target (to remove the video amplifier frequency band limitation) the individual wires of the 1000 mesh target screen can be resolved. This is equivalent to 2000 television lines per inch. Separate tests have shown that under ideal conditions the image section is also capable of equal resolution. The contrast ratio near the limiting resolution is, of course, very low.

The limiting resolution of the small tube is enhanced by the use of the wall and photocathode screens as well as by the use of a smaller defining aperture in the gun. However, it should be pointed out that high light resolutions approaching that of the image orthicon can be attained only if great care is taken in selection and adjustment of the tube.

2. *Signal Output*

At very low light levels, where full storage occurs, the signal output at the target is independent of tube size. This assumes that the camera lens aperture is adjusted to give the same depth of focus.

At high lights the area of the target is important in determining signal output. For a "close spaced" target (i.e. glass-screen spacing less than one picture element) the signal output varies as the target area and signal-to-noise ratio varies as the diameter. For "wide spaced" targets (i.e. glass-screen spacing wide compared to a picture element) the signal output varies more nearly as the diameter of the target and signal-to-noise ratio as the $(\text{diameter})^{\frac{1}{2}}$. The target spacing of the Mimo is of the order of two mils which is about the same as in the image orthicon. Because this is somewhat intermediate between "close" spacing and "wide" spacing, the drop in signal-to-noise ratio is bracketed by the above limiting cases.

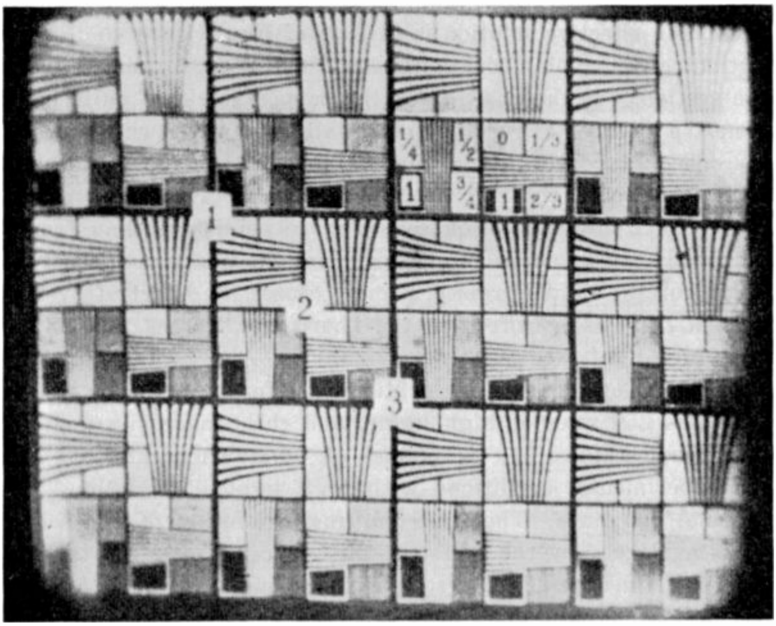

Fig. 4—Photograph of a test pattern transmitted by the Mimo tube.

Another factor influencing the variation of signal with target area at high lights is the degree of overlapping of the beam spot in adjacent lines. Some overlapping does occur in the Mimo tube (in spite of the high *limiting* resolution quoted above), and this would contribute to enhanced signal at high lights owing to the recharging of the target between successive scans.

3. *Choice of Lens*

The first consequence of the smaller photocathode of the Mimo tube is that a shorter focal length lens may be used for the same angle of view. This results in a saving in space although a faster lens is required. If, in addition, the lens diameter is also scaled proportionally, so that the numerical aperture remains unchanged, increased depth of focus is obtained at the expense of operating sensitivity.

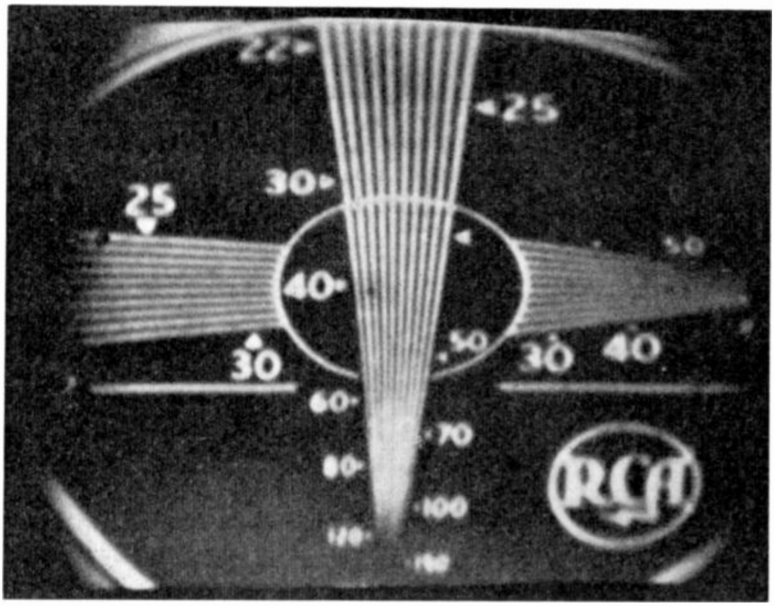

Fig. 5—Enlarged section of a test pattern transmitted by the Mimo tube. (The optical pattern was projected on the photocathode at normal size while the scanning amplitude was reduced to cover only the center portion of the target. This procedure tests the resolving power of the tube by reducing the limitations of the amplifier frequency response as well as cross talk in the image section. The numbers on the pattern should be multiplied by ten to give the resolution in television lines.)

At high light levels, in which range the signal output is substantially independent of scene brightness and in which range the Mimo was mostly used, lens speed is of no importance. Here the shorter focal length lens is an unalloyed gain in conserving space.

The size of the image projected on the photocathode of the Mimo is approximately the same as that of one frame of a 35-millimeter motion picture film. Thus a wide choice of lenses for the Mimo is at hand. An f/2.0 lens was used in the camera but light conditions in the airborne application were such that the lens was often stopped down to as small as f/22.

CONCLUSIONS

A useful television pick-up tube of reduced size has been developed for airborne television purposes. This tube retains the high sensitivity and stability under adverse lighting conditions which have characterized the image orthicon. At the same time changes in design have been incorporated which make for simplified operation of the camera. It is believed that the Mimo tube represents a first step toward the development of a television camera which approaches the miniature photographic camera in convenience and portability.

ACKNOWLEDGMENTS

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TELEVISION EQUIPMENT FOR AIRCRAFT*†

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Summary—The design considerations involved in the development of lightweight television equipment for airborne use are discussed in Part I of this paper. Following this is a description of Block I television equipment developed in accordance with these considerations. Flight testing of Block I equipment brought to light several difficulties peculiar to the transmission of television signals from aircraft. In Part II a number of these difficulties are discussed as well as methods developed for minimizing them.

PART I

DEVELOPMENT AND DESIGN OF LIGHTWEIGHT TELEVISION EQUIPMENT

THE first development work on television transmitting apparatus for use in aircraft was undertaken in 1936 under the direction of R. D. Kell. This was a direct outgrowth of Dr. V. K. Zworykin's memorandum to David Sarnoff of April, 1934.[1] Equipment was built using the 1850 type iconoscope and was installed and tested in a Ford trimotor airplane. Results obtained with this equipment clearly demonstrated the potential usefulness of television for the armed services. However, it was apparent that this equipment, although much smaller than other commercial equipment of the same type, was still appreciably larger and heavier than was considered desirable.

The advent of the 1848 iconoscope, a smaller tube than its predecessor, the 1850, made possible the design of television cameras of greatly reduced size. Commercial equipment using this tube and designed specifically for field pickup use was introduced in 1939. All the equipment associated with the camera was built into suitcase-type units, making it convenient to transport.

At this time both the Army and the Navy began to take a very serious interest in the military possibilities of television and equipment quite similar to the suitcase-type commercial design was constructed for airborne use in 1940. This equipment has been described in previous literature.[2] While this apparatus was useful in supplying a need for experimental and training equipment it was not particularly suit-

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* Decimal Classification: R583 X R520.

† Reprinted from *RCA REVIEW*, December, 1946.

[1] V. K. Zworykin, "Flying Torpedo with an Electric Eye," *RCA REVIEW*, Vol. VII, No. 3, pp. 293-302, Sept., 1946.

[2] C. J. Marshall and L. Katz, "Television Equipment for Guided Missiles," *Proc. I.R.E.*, Vol. 34, No. 6, pp. 375-401, June, 1946.

able for military use since it had been designed primarily to meet commercial standards.

In 1940 a program was undertaken to develop television equipment specifically for use in military aircraft and having the following design objectives:

(1) Light weight
(2) Compactness
(3) Reasonably low power drain from a 12-volt direct-current source
(4) Reliable line-of-sight range up to 10 miles
(5) Resolution capability close to commercial standards
(6) Unattended operation possible at the camera and transmitter
(7) Overall ease of installation, operation and maintenance.

Naturally, a certain amount of compromise was necessary in order to produce results in reasonable agreement with the above requirements. Some of the fundamental decisions made in order to achieve these results are listed below:

(1) It was decided to place the iconoscope with its associated circuits and the transmitter in a single unit instead of two separate units. Although some difficulty was anticipated because of feedback from the transmitter into the camera circuits it was felt that advantages were to be gained with regard to overall size and weight considerations as well as in simplification of the interconnection problem.

(2) The dynamotor power supply was made a separate unit so that a dry battery supply might be substituted for the dynamotor under certain operating conditions.

(3) The frame frequency was made approximately 40 cycles and the line frequency approximately 14,000 cycles without definite relationship between the two as is necessary for commercial interlaced scanning. This decision greatly simplified the problem of developing suitable synchronizing and other line and frame frequency pulses. The choice of 40 cycles for frame frequency was a compromise. A lower frequency would have resulted in objectionable flicker and a higher frequency would have resulted in lower overall resolution for the available bandwidth. The line frequency of 14,000 cycles was chosen to give 350 scanning lines. This resulted in approximately 275 lines resolution in the vertical direction and 350 lines in the horizontal direction. Equal values of resolution could have been obtained by using a higher frequency for horizontal scanning and a consequent increase in the number of lines. However, it

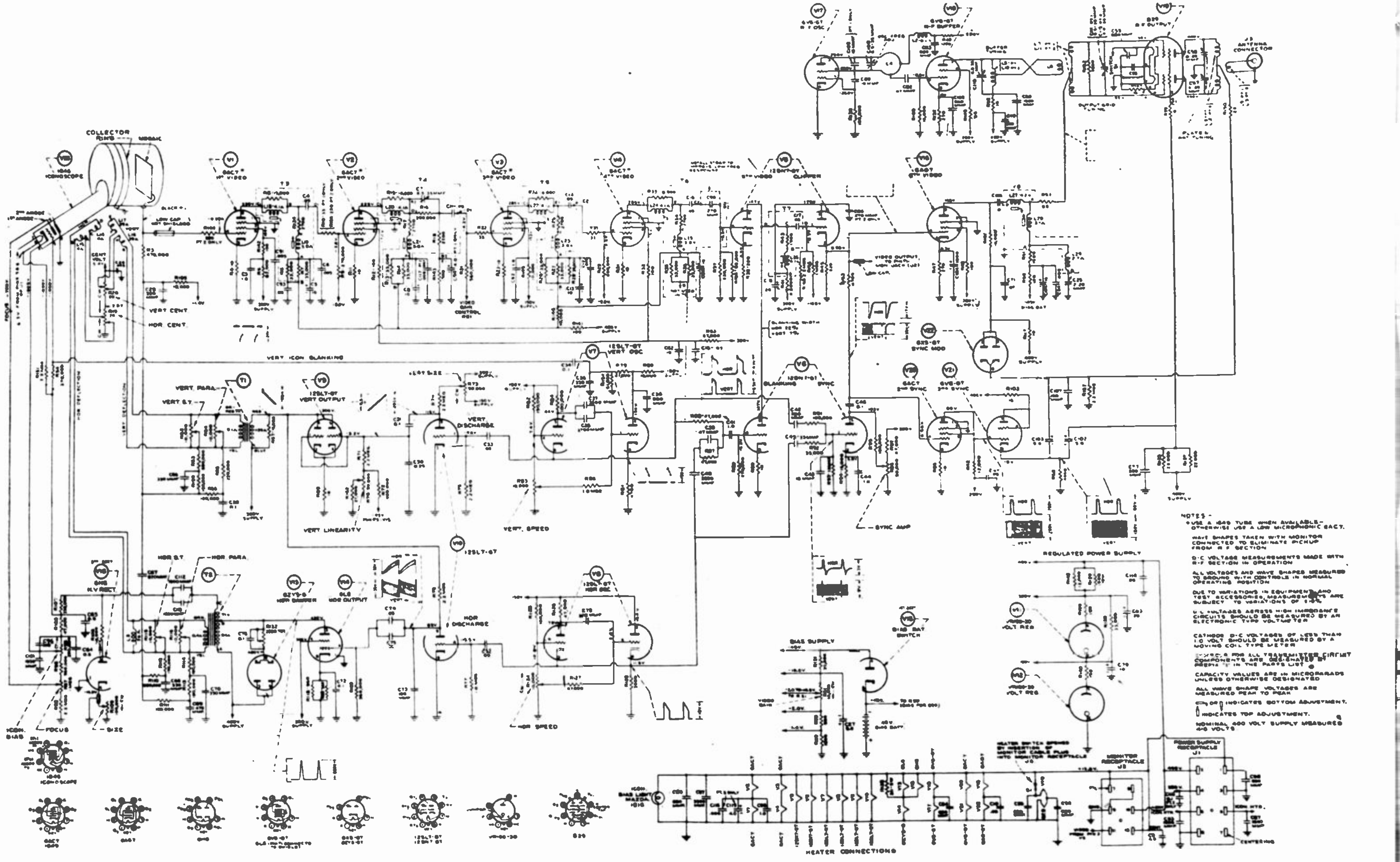

Fig. 4—Camera transmitter—schematic diagram.

NOTES -

*USE A 1846 TUBE WHEN AVAILABLE— OTHERWISE USE A LOW MICROPHONIC 6AC7.

WAVE SHAPES TAKEN WITH MONITOR CONNECTED TO ELIMINATE PICKUP FROM R F SECTION

D-C VOLTAGE MEASUREMENTS MADE WITH R-F SECTION IN OPERATION

ALL VOLTAGES AND WAVE SHAPES MEASURED TO GROUND WITH CONTROLS IN NORMAL OPERATING POSITION

DUE TO VARIATIONS IN EQUIPMENT AND TEST ACCESSORIES, MEASUREMENTS ARE SUBJECT TO VARIATIONS OF ±10%

D-C VOLTAGES ACROSS HIGH IMPEDANCE CIRCUITS SHOULD BE MEASURED BY AN ELECTRONIC TYPE VOLTMETER

CATHODE D-C VOLTAGES OF LESS THAN 1.0 VOLT SHOULD BE MEASURED BY A MOVING COIL TYPE METER

SYMBOLS FOR ALL TRANSMITTER CIRCUIT COMPONENTS ARE DESIGNATED BY PREFIX "T" IN THE PARTS LIST

CAPACITY VALUES ARE IN MICROFARADS UNLESS OTHERWISE DESIGNATED

ALL WAVE SHAPE VOLTAGES ARE MEASURED PEAK TO PEAK

OR INDICATES BOTTOM ADJUSTMENT.

INDICATES TOP ADJUSTMENT.

NOMINAL 400 VOLT SUPPLY MEASURED 410 VOLTS

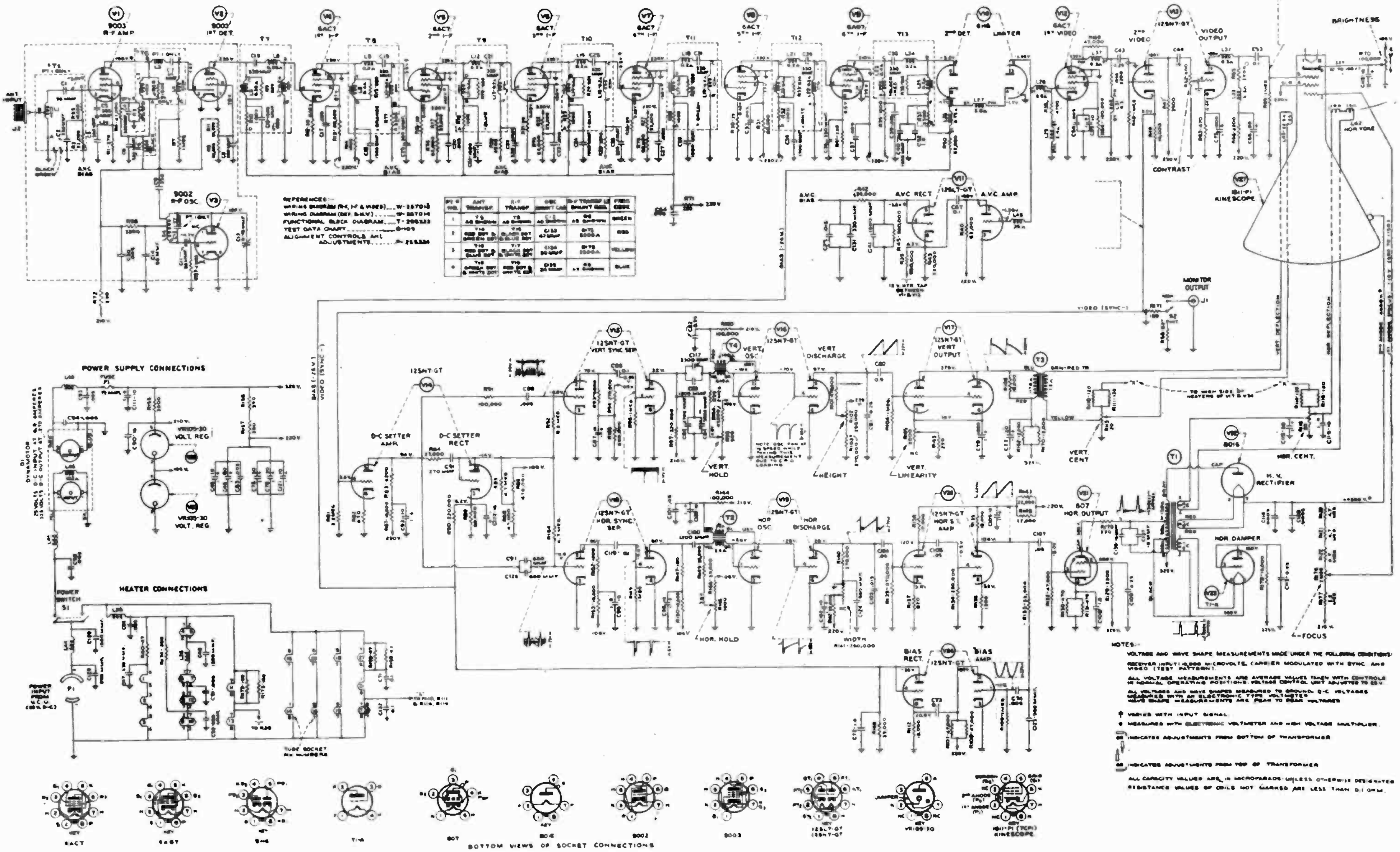

Fig. 10—Receiver—schematic diagram.

was considered desirable to keep this frequency somewhat low so that the horizontal scanning circuits would require less power and would have lower dissipation.

(4) The vertical synchronizing pulses were made approximately one and a half lines long so that only one or at the most two horizontal synchronizing pulses would be lost during the vertical synchronizing period. It was considered unnecessary to add sufficient tubes and circuit elements to place slots in the vertical pulse as is done with the standard Radio Manufacturers Association synchronizing signal. The timing of the leading edge of the vertical synchronizing pulse was also made coincident with the leading edge of the vertical blanking pulse. This simplification of the synchronizing signals was made possible by the choice of sequential rather than interlaced scanning and resulted in a considerable saving in tubes and circuit elements.

(5) The overall video band width of the system was made 4.5 megacycles and both the transmitter and receiver were designed for double sideband operation. Obviously, single-sideband operation would have resulted in a more complex transmitter. The decision to make the receiver accept both side bands was based on the desirability of eliminating tuning controls in order to simplify receiver operation. The band width chosen was considered to be a fair compromise between the resolution capability of the 1848 iconoscope, normally somewhat better than 350 lines, and the number of video and intermediate-frequency amplifier stages required in the camera, transmitter and receiver in order to obtain adequate amplification. Field tests with earlier types of television equipment indicated that the resolution capability of the proposed system was entirely adequate for most of the military applications proposed.

(6) A transmitter frequency was chosen in the neighborhood of 100 megacycles. This made possible efficient tuned circuits with lumped constants of relatively small size and also an antenna structure not considered to be excessively large.

(7) Picture monitoring facilities were not included with the camera transmitter unit since unattended operation was expected. Instead a separate monitor unit employing a 7-inch kinescope was designed so that it could be connected to the camera transmitter with a single cable connection. When this was done, the *B*+ supply to the transmitter output tube was automatically transferred to the monitor circuit. When this connection was

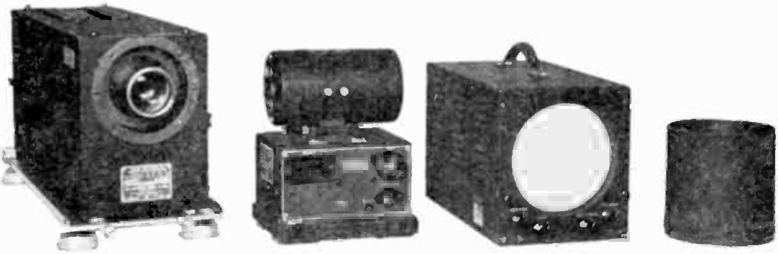

Fig. 1—Block I equipment at transmitting location.

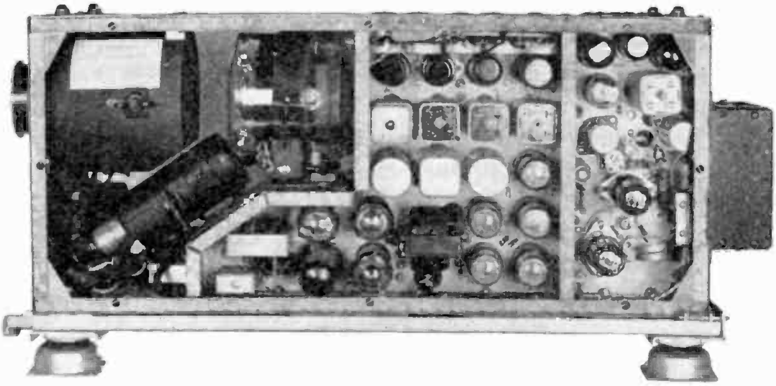

Fig. 2—Camera transmitter—tube side.

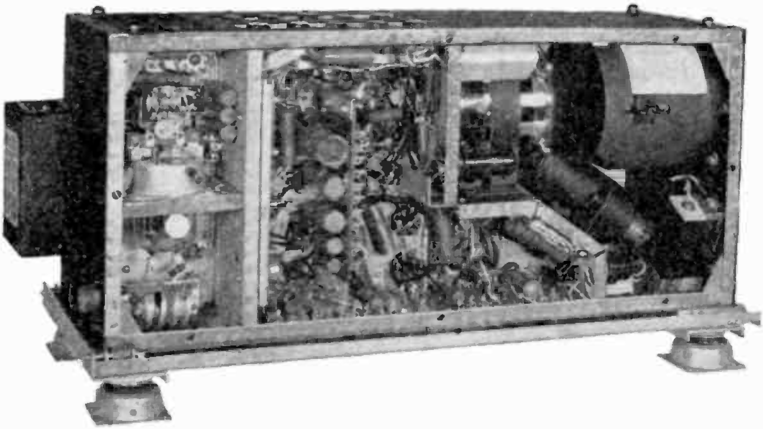

Fig. 3—Camera transmitter—circuit side.

made, the power drain on the dynamotor remained approximately the same as with the transmitter load.

Other design considerations involved in the development of this equipment will be discussed in the detailed descriptions of each unit which follow. The three units necessary at the transmitting location are shown in Figure 1. Figure 2 presents a side view of the camera-transmitter showing most of the tubes. As may be seen in the photograph the type of construction used resembles an "I" beam with closed ends and results in a light-weight but very rigid unit. The center chassis section on which most of the tubes are mounted is welded to the "wrap around." Both side covers of the unit, which are removable, are fastened in place by six "Dzus" type fasteners. Special spring contacts are placed on all the outside flanges of the case to insure a good ground between the covers and the case. The front compartment houses the lens mounting and iconoscope; the central portion of the case contains the video amplifier and deflecting circuits; and the transmitter is located in a narrow section at the rear of the unit.

Ten control knobs are recessed in the top of the case. These, together with the power switch on the back of the unit may be considered as "operating" controls. Their functions are as follows:

(1) Horizontal Sawtooth Shading
(2) Vertical Sawtooth Shading
(3) Horizontal Parabola Shading
(4) Vertical Parabola Shading
(5) Iconoscope Horizontal Centering
(6) Iconoscope Vertical Centering
(7) Iconoscope Size
(8) Iconoscope Focus
(9) Video Gain
(10) Iconoscope Bias

Five additional controls are accessible after removal of the right-hand cover. These controls require adjustment only when associated tubes are changed. The five controls are arranged along the partition separating the transmitter section from the rest of the unit. Reading from top to bottom they are:

(1) Synchronizing Amplitude
(2) Iconoscope Vertical Size
(3) Iconoscope Blanking
(4) Vertical Speed
(5) Horizontal Speed

The power from the dynamotor is brought in through a plug in the rear of the unit. A coaxial connection for a 75-ohm radio-frequency output line from the transmitter is also located at the rear of the unit.

Figure 3 shows the other side of the case where most of the circuit elements are placed. In spite of the large number of components most of them are readily accessible. Electrically there are four main groups of circuits making up this unit. These are:

1. The master scanning oscillators and deflecting circuits
2. The iconoscope with its associated video amplifier, shading, blanking and synchronization mixing circuits
3. The transmitter circuits
4. Power supply circuits

Considering first (1) above, this comprises that part of the schematic shown in Figure 4 (facing page 434) associated with tubes V6, V7, V8, V9, V10, V13, V14 and V15. Cathode-coupled multivibrators were chosen as master oscillators for both line and frame frequencies because they produce pulse wave shapes which can easily be transformed into synchronizing and blanking pulses. Across the common cathode resistor of this type of oscillator is found a positive pulse with steep vertical sides, a decreasing amplitude during the pulse itself and a constant voltage level between pulses. The wave shape of the voltage on the output plate of the oscillator is similar to that on the cathode but it has the opposite polarity. It may be noted that a relatively high value of cathode resistor is used in the multivibrator circuits in order to obtain sufficient pulse voltage on the cathodes for satisfactory operation of the mixer circuits. This requires a positive bias on the multivibrator grids.

An adjustment of this bias voltage on the output triode of the multivibrator provides a convenient speed control. The circuit elements of the multivibrators were so chosen that the resulting widths of the pulses appearing on the cathodes have a duration such that these pulses are suitable for use as kinescope blanking signals. Narrower pulses obtained by differentiating the cathode pulses are used in developing the synchronizing signal. The same cathode pulses are also fed to the usual type of discharge-tube sawtooth generating circuits in order to create the necessary scanning waveforms. The pulse appearing in the plate circuit of the vertical multivibrator is used as a source of blanking signal for the iconoscope grid.

Resistance mixing circuits are used to combine the horizontal and vertical pulses from the multivibrator cathodes in the grid circuit of the blanking amplifier (Section A of tube V6), and the resultant is added to the video signal in the common plate circuit of the video amplifier tube V5, Section A, and tube V6, Section B. The amplitude of these pulses on the grid of the tube V5, Section B, is sufficient to drive the grid beyond cut-off. Thus no video signal is transmitted during the pulse time with the result that the kinescope screen of the receiver can be made black during this period and the retrace of the cathode ray beam will not be visible. The synchronizing pulses also obtained from the pulses appearing on the multivibrator cathodes, are much narrower than the corresponding blanking pulses because of the insertion of small coupling capacitors in the mixing circuit at the grid of tube V6, Section B, which pass only the steep wave front of the pulses. These short pulses appear in the plate circuit of the tube and are added to the combined video and blanking signal appearing at the cathode of the video amplifier tube, V5, Section B.

The function of the 18-micromicrofarad, additional shunt capacitance from grid to ground of the synchronizing pulse amplifier tube V6, is to delay slightly the leading edge of the horizontal synchronizing pulses with respect to the leading edge of the horizontal blanking pulses thereby producing a narrow but adequate "front porch" or delay period. The function of this delay is the same as that in the standard Radio Manufacturers Association signal—to prevent video signals near the leading edge of the blanking signal from changing the timing of the synchronizing signals. The same capacitor also causes a slight delay of the leading edge of the vertical synchronizing signal with respect to the vertical blanking signal with the same result.

As noted previously, the necessary saw-tooth wave forms for deflecting the iconoscope cathode-ray beam are derived from the pulses appearing on the cathodes of the two multivibrators by impressing them on the grids of the two discharge tube sections of tube V10. The vertical saw-tooth is amplified in tube V9 and fed through a step-down transformer to the vertical deflecting coils of the iconoscope yoke. A horizontal saw-tooth appears in the plate circuit of Section B of tube V10. The plate supply for this circuit comes from the vertical output tube V9. Thus the horizontal saw-tooth is modulated by a vertical saw-tooth thereby producing the keystone correction made necessary by the construction of the iconoscope. This requires a greater amplitude of horizontal scanning current when the beam strikes that part of the mosaic nearest the electron gun. A control is not required on the amplitude of keystone correction since the circuit constants of the horizontal saw-tooth generating circuit were chosen to give the exact amount of correction necessary. This keystone-corrected horizontal saw-tooth is amplified in the horizontal output tube (V14), and the horizontal damping tube V13 supplies the necessary damping to suppress horizontal frequency transients. A step-down transformer in the plate circuit of tube V14 feeds the deflecting signal to the horizontal deflection coils of the iconoscope yoke.

In order to obtain a direct-current second anode voltage of approximately —1000 volts for the iconoscope, the plate transformer of the tube V14 is also provided with a step-up winding which feeds enough horizontal signal to the high voltage rectifier tube V15 to obtain the necessary voltage.

Since a horizontal size control would also cause a variation in the second anode voltage which would, in turn, require a readjustment of vertical size, it was considered more desirable to keep the horizontal scanning current in the yoke at a fixed value and provide an adjustment of the second anode voltage supply.

This adjustment changes the deflection amplitude of horizontal and vertical scanning circuits simultaneously so that a quick check of scanning amplitudes can be made by decreasing the second anode voltage slightly which should normally bring the frame of the iconoscope mosaic uniformly into view at all four edges of the picture. The second anode voltage is made adjustable by inserting a variable resistor in the cathode lead of the rectifier tube V15.

The second major division of the camera transmitter mentioned in (2) above is the iconoscope and its associated video amplifier consisting of tubes V1, V2, V3, V4, and V5.

The iconoscope output resistor has a relatively high value in order to improve the signal-to-noise ratio at low frequencies, thereby reducing microphonics and ripple. This causes a drop in the high-frequency content of the video signal which is compensated for by the "high peaker" coupling circuit between tubes V2 and V3.

All the video stages except the one associated with V5 were made using both a series and a shunt inductance in the amplifier plate circuits in order to equalize the high frequency response. The single inductance in the plate circuit of the first triode section of V5 was adjusted to produce a broad resonance in the middle frequency range in the neighborhood of 2.5 to 3.0 megacycles in order to equalize the dip in response resulting from the other four circuits in cascade. In this manner an overall flat response was obtained to approximately five megacycles. The low-frequency phase and amplitude response, affected mainly by the choice of coupling capacitors, grid leaks, and plate-filter by-pass capacitors, is such that the amplifier is capable of passing a 40-cycle square wave with very little distortion. In order to eliminate some of the disturbing effects in the picture caused by microphonics especially at the lower frequencies, a small series coupling capacitor (270 micromicrofarads) is placed in series with the grid of the video amplifier V5. This is normally short-circuited with a short length of wire which can then be cut if noise and vibration conditions are especially severe. The consequent reduction in low-frequency response attenuates most of the usual microphonic frequencies but also results in some loss of picture intelligence. The most objectionable effect is the introduction of rather long horizontal light streaks after dark objects in the picture.

The potentiometers and associated circuits in the low-impedance side of the two deflection transformers provide shading signals which are mixed with the video signal in the iconoscope output circuit. These signals compensate to a certain extent for the spurious signal generated by the iconoscope itself (black spot signal). Four compensating signals are generated; horizontal and vertical saw-tooth wave forms, and horizontal and vertical parabola wave forms. Since the two deflecting output circuits are balanced to ground (by having the connection between pairs of deflecting coils grounded through the low-impedance centering circuits) the mid-position of the four potentiometers corresponds to zero shading signal. Moving the potentiometers to one side of this point provides one polarity of shading signal, moving them to the other side provides the opposite polarity.

The waveform of the signal appearing at the vertical transformer secondary is a reasonably linear saw-tooth so that a single differentiation circuit consisting of R65 (220,000 ohms) and C28 (0.1 microfarad) supplies a satisfactory vertical parabola. The 330-micromicrofarad capacitor (C68) is for the purpose of by-passing horizontal frequency pulses introduced into the vertical circuits through coupling in the deflecting yoke. Since the voltage waveforms appearing on the secondary side of the horizontal output transformer are pulses, it is necessary to use a single differentiating circuit to obtain saw-tooth waveforms and two of these circuits in series to obtain parabolic waveforms. The series resistors in the mixing circuit used to combine all four different waveforms are so chosen that the maximum voltages derived from all shading potentiometers are approximately the same.

The radio-frequency oscillator and buffer circuits are also shown in the schematic (Figure 4). V17 is the radio-frequency oscillator, the tank circuit of which consists of L4, C108, and C109. C109 may be removed for change of frequency and C108 is variable to permit tuning to the exact frequency desired within the range. L4 is tapped in two places, the center tap going to the plate supply voltage through L2 and R40, the other tap being the oscillator output and feeding through C82 to the buffer grid. The voltage supply is connected through the circuit of L4 to both the plate and the screen grid. This oscillator has the control grid and screen grid closely coupled together so that the tube oscillates as a low "mu" triode. The output is taken from L4 through C82 to the control grid of the buffer stage V18. This buffer stage is a radio-frequency amplifier with a tuned plate circuit consisting of L5 and C110. Connected to L5 is a link circuit which has a coupling coil L6 feeding to the radio-frequency output stage, V19.

The radio-frequency output and modulation circuits are also shown in Figure 4. The coupling coil L6 couples to L7 which is the grid tuning circuit for the radio-frequency output tube V19. This is a double section tube having essentially two beam-power structures within one envelope. L7 has a center-tap connection for grid-modulation and bias for V19. The bias supply, from a 45-volt dry battery, passes to L7 through the output circuit of the last video amplifier which contains the video signal, synchronization and blanking. Consequently, the grid voltages on V19 will be modulated with these various voltages. The plates of this tube are connected to their respective ends of L8. L8 is part of the output tank circuit which consists of L8, C58 and C57. C58 and C57 are tuning condensers ganged together and controllable by one adjustment. The heater circuit of V19 is supplied directly from the 12.5-volt power supply through a switch, shown as S4, which opens the heater circuit when the monitor cable plug is inserted. L9 is a radio-frequency output coil coupled to L8 and is connected to the antenna output terminal.

To insure that synchronizing pulses will always be transmitted at full power, further modulation is accomplished in the plate circuit of V19 for the synchronizing pulses only. The output from V20 consists of synchronizing pulses applied to the grid of V21. The plate current for V19 passes through R117, V22 and R150 to the mid-tap of L8. V22 is a half wave rectifier tube through which the plate current of V19 flows. Condenser C103 is connected from the plate circuit between V22 and R150 to ground through a 4,000-ohm resistor. This 4,000-ohm resistor is also in series with the cathode of V21. Between synchronizing pulses, V21 is not drawing plate current and consequently C103 will be charged to full plate voltage. When a synchronizing pulse arrives at the grid of V21 it will draw considerable plate current, causing current to flow through R68, making the lower end of C103 approximately 180 volts positive. This 180 volts added to the approximately 400 volts already on C103 will raise the plate voltage of V19 to about 580 volts. The condenser C103 supplies V19 with power during the short interval that the synchronizing pulse continues, as V22 will not allow a discharge of this condenser back through the power supply.

Consequently, during the period of a synchronizing pulse, V19 is driven completely to the limits of operation and definite modulation of synchronizing pulses is secured. In order to make the rise of plate voltage completely effective, C102 is placed in this circuit. This raises the screen grid voltage on V19 in much the same manner as the plate voltage was raised. Therefore, by having both grid and plate modulation for the synchronizing pulses, it is possible to hold the receiver in synchronism under severe conditions of noise and interference.

The power supply unit consists of a dynamotor, a power-control relay and two filter systems and is shown schematically in Figure 5. All voltages required for operation of the camera transmitter and monitor are obtained through connections to this unit. The dynamotor, which comprises the means for converting the 12-volt primary power into the high-voltage direct-current potential required for operation of the camera-transmitter and monitor, is mounted on the top of the unit, thus providing adequate ventilation and ease of access for routine service of the rotating equipment.

All cable connections required for normal operation of the equipment are completed by means of an 8-connector receptacle (camera-transmitter connections) and a 2-connector plug (power connection). A third receptacle is provided for connection of a remote power control switch if such is required.

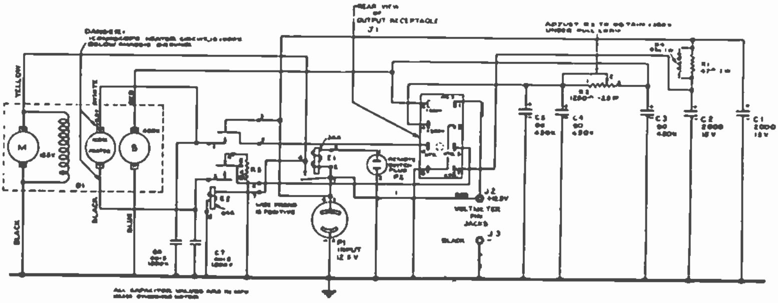

Fig. 5—Power supply—schematic diagram.

The dynamotor is designed for operation from a 12.5 volts direct-current supply. Two secondary windings and commutators are provided; one insulated for 1000 volts delivering 6.3 volts direct-current to the filament of the iconoscope and the other delivering 400 volts direct-current to the camera-transmitter and to the input of a filter system. Approximately 280 volts is supplied to the camera-transmitter from the output of this filter. The iconoscope centering voltage is derived from the heater supply but is filtered to keep ripple on this supply from affecting the deflecting circuits.

In the camera transmitter unit itself are located two regulator tubes

V11 and V12 in series. These tubes provide a constant voltage supply to those circuits which are sensitive to slight changes in anode voltage. Such circuits include the scanning oscillator circuits, the screen grids of the video amplifier tubes, the blanking and synchronizing amplifier tubes and the clipper tube V5. The bias supply is derived from a 45-volt "C" battery. This was chosen primarily to supply the transmitter output tube with a low impedance bias source so that grid current would not change the bias voltage. A bleeder is used to provide the necessary bias voltages required for the video amplifier. This bleeder is supplied through one diode section of tube V15 so that the drain on the battery is automatically removed whenever the heater supply is removed. The second-anode supply for the iconoscope is derived from the horizontal output transformer and was discussed in detail in the description of the scanning circuits.

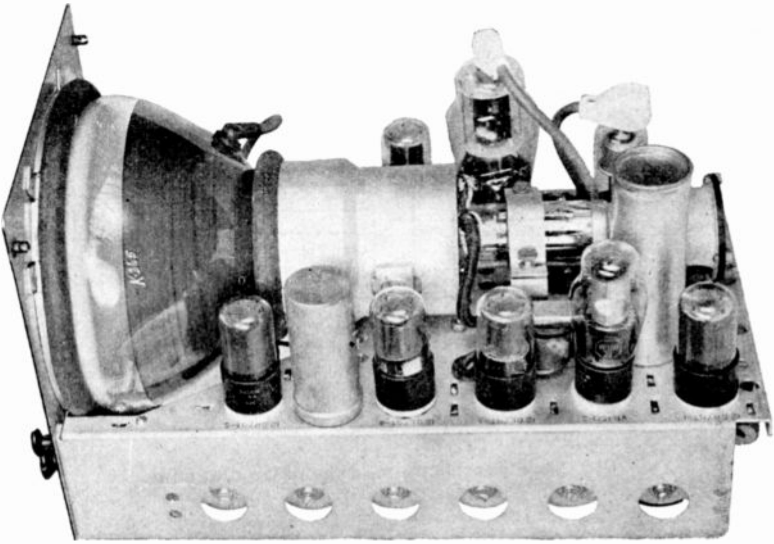

Fig. 6—Monitor—with cover removed.

The monitor designed for use with the camera transmitter is shown with the transmitting equipment in Figure 1. In Figure 6 it may be seen again with the cover removed. The monitor unit consists of a type 1811-P1, 7-inch kinescope, together with its associated deflection circuits and video amplifier. Connection of the monitor to the camera transmitter automatically opens the heater circuit of the Type 829 output tube in the transmitter (V19), thus compensating for the power required by the monitor unit, and maintaining normal $B+$ voltage.

Four controls required for normal operation of the monitor unit are located on the front panel, below the screen of the kinescope. These controls, progressing from left to right are, respectively, the "horizontal hold" control, the "focus" control, the "brightness" control and the "vertical hold" control. The "width", "height", "vertical linearity", "vertical centering" and "horizontal centering" controls, being used infrequently, are screwdriver adjustment. They are arranged from front to back along the right-hand side of the unit in the order named. The monitor is provided with a removable light shield.

To obtain access to tubes and circuit components, the chassis can be removed from the case after loosening the three "Dzus" fasteners on the front of the unit and also removing the power cable.

The monitor was designed specifically for operation from a composite picture signal of the type used to feed the transmitter modulator and having approximately one volt peak-to-peak amplitude. It contains a video amplifier, separating circuits, line and frame frequency oscillators and their associated deflecting circuits. A schematic diagram is shown in Figure 7.

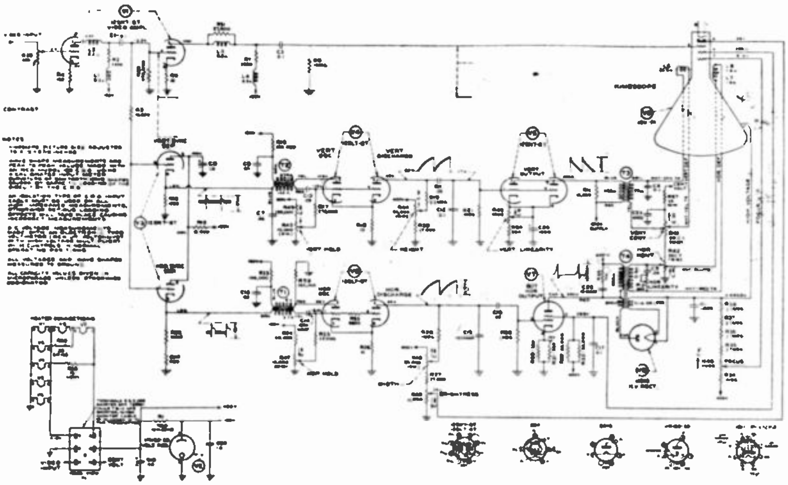

Fig. 7—Monitor—schematic diagram.

A 12SN7GT (V1) forms a two stage video amplifier which receives a video signal from the monitor output of the transmitter unit and feeds the signal to the kinescope grid. From the plate of the first amplifier stage the signal is fed to the synchronizing separator tube, a 12SL7GT (V3). There is sufficient synchronizing signal amplitude at this point so that the video signal drives the grids of this tube beyond cut-off and only synchronizing pulses appear on the two cathodes from which the vertical and horizontal oscillator (V4 and V6) are fed. Both horizontal and vertical oscillators are conventional blocking oscillators except that speed controls are provided by adjustments of the amount of positive bias on the oscillator grid leak ground returns.

A resistance-capacitance damping circuit is used across the horizontal deflection transformer in place of a damping tube in order to save space, although the circuit is somewhat less efficient.

This same transformer also has a step-up winding and a filament winding which supplies heater and plate voltage to a Type 8016 rectifier, V8. The rectified direct-current output (approximately 4500 volts) is applied to the second anode of the kinescope. A potentiometer in a bleeder circuit on this supply furnishes an adjustable first anode voltage. A VR-150-30 regulator tube (V2) supplies a substantially constant 150-volt potential to the video amplifier and the vertical and horizontal oscillators, so that gain and scanning oscillator frequency will be reasonably constant regardless of change in supply voltage.

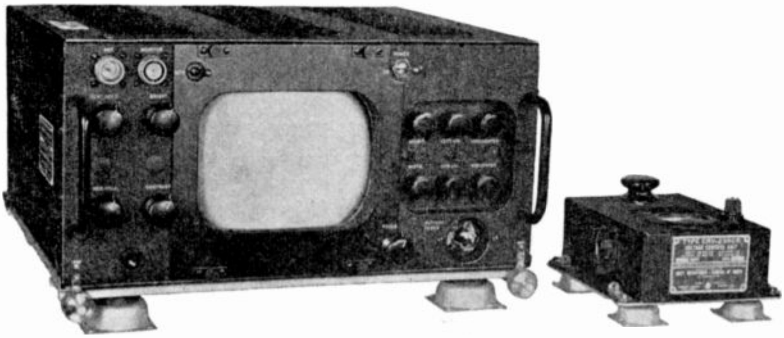

Fig. 8—Block I receiver and voltage control.

Figure 8 shows the receiver and a voltage control box designed specifically for adjusting the input voltage to the receiver. In Figure 9 the top cover of the receiver has been removed to show the general construction and arrangement of parts. It may be seen that in design features it is somewhat similar to the camera-transmitter unit.

The receiver is completely self-contained and includes its own power supply, designed for operation from a 12- to 14-volt storage battery or the equivalent. The entire unit mounts, by means of thumb-screws, on a shock-mounting base supplied with the equipment. Plate power is supplied by an internal dynamotor developing 330 volts (direct-current), when connected to a 12.5-volt direct-current source. All cable connectors and controls necessary for set-up and operation of the unit are brought out to the front panel. Battery connections are terminated at a polarized plug, mounted at the lower right-hand side of the panel. The antenna connection is completed through a connector mounted at the upper left-hand side of the receiver. The antenna circuit is designed to work out of a 75-ohm coaxial line.

Controls requiring adjustment under normal operating conditions are mounted on the front panel. Those controls requiring only infrequent adjustment are recessed below the level of the panel.

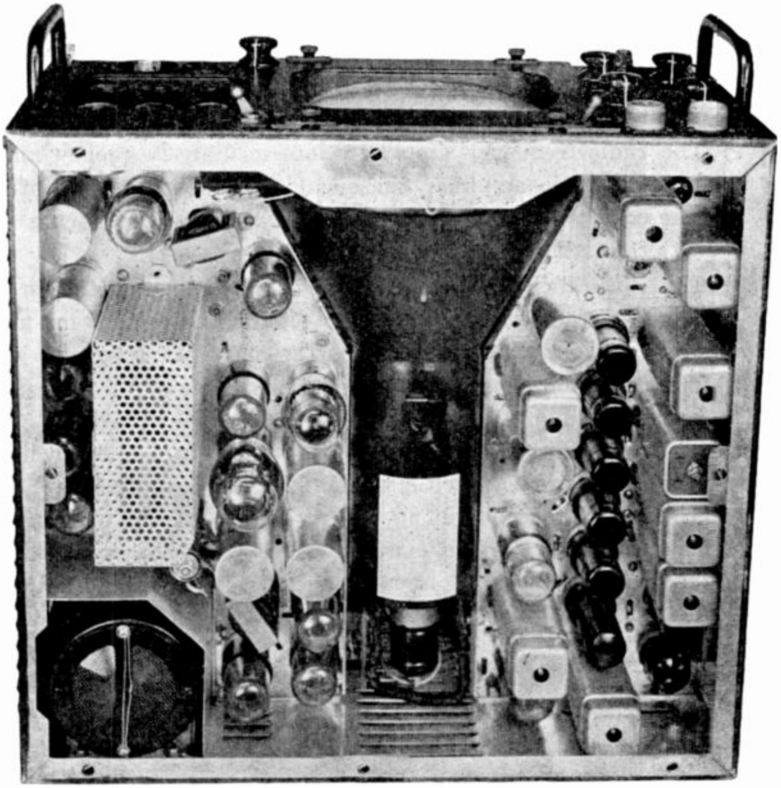

Fig. 9—Block I receiver—top cover removed.

The controls mounted on the panel are the "vertical hold" control located on the left end of the upper row of controls, the "brightness" control at the right of the vertical hold control, the "horizontal hold" control at the left of the lower row of controls, the "contrast" control at the right of the horizontal hold control, and the "focus" control at the right and below the kinescope. The power switch is mounted directly above the focus control.

The controls which are recessed are (starting at the left end of the top row); the "vertical size" control, the "vertical linearity" control and the "vertical centering" control, the "horizontal linearity" control and the "horizontal centering" control.

The connector plug located at the right of the antenna plug is used when it is necessary to send a signal to a remote monitor. When it is desired to utilize the monitor signal, the small switch mounted next to the monitor output plug must be switched to the monitor ouput position; otherwise, the switch should always remain in the "internal" position.

Electrically, the receiver may be separated into four main divisions:

(1) The radio-frequency amplifier, intermediate-frequency amplifier and video amplifier chassis is located at the left of the kinescope when the receiver is viewed from the wiring side with the face of the kinescope tube at the top.

(2) The seven-inch kinescope, Type 1811-P1 is mounted in the middle of the receiver. The kinescope is housed in a mu-metal shield, within which is mounted the magnetic deflecting yoke.

(3) The deflection chassis, mounted to the right of the kinescope, contains the circuits which separate the synchronizing signal from the video signal, the blocking oscillators and discharge tubes, the saw-tooth voltage amplifiers and output tubes. These are utilized for generating the saw-tooth current wave in the magnetic deflecting yoke which deflects the electron beam in the kinescope. The horizontal output circuit which provides the high voltage power supply for the kinescope is also located in this section.

(4) The dynamotor, located at the right rear corner of the deflection chassis supplies all the low-voltage direct-current power required for operation of the receiver.

A schematic diagram of the receiver is shown in Figure 10 (facing page 435). The radio frequency amplifier, Type 9003 (V1), the first detector, Type 9003 (V2), and the fixed frequency radio-frequency oscillator, Type 9002 (V3), are mounted on a small chassis. This chassis is so designed that it can be removed from the radio-frequency and intermediate-frequency chassis of the receiver unit and replaced with a unit designed to receive another carrier frequency within reasonable limits if that becomes desirable. The radio frequency unit is followed by a six-stage intermediate frequency amplifier (tubes V4, V5, V6, V7, V8 and V9). The first five stages of the intermediate frequency amplifier utilize Type 6AC7 tubes and the sixth stage a Type 6AG7 tube.

The coupling circuits in the intermediate frequency amplifier circuit are "constant K" type filter sections with resistance loading on the input or plate side, and a full shunt arm on the grid or output side.

The sixth intermediate frequency amplifier stage drives the second detector, one section of a Type 6H6 tube (V10). The other section of V10 is used as a limiter tube to limit the noise peaks coming through input voltages.

The picture signal output from the second detector, after going through the limiter, is amplified in the video amplifier tube, Type 6AC7 (V12), and the two sections of the Type 12SN7GT output tube (V13). The first half of V13 furnishes at its cathode an output signal for the external monitor and the video signal for the synchronizing pulse separating system. Its plate circuit furnishes video signal to the contrast control. The second half of V13 is used as an output tube, the signal on its grid being obtained from the contrast control, the plate circuit driving the grid of the Type 1811-P1 (C7466) kinescope (V27).

The automatic volume control amplifier and rectifier tube, Type 12SL7GT (V11), is also mounted on this chassis. Its purpose is to hold the output of the set at a constant level over a wide range of signal input voltages.

Mounted on the deflection chassis is the video signal amplifier and direct-current setting tube, Type 12SN7GT (V14), the vertical synchronizing separator and clipper tube, Type 12SN7GT (V15), the vertical oscillator and discharge tube Type 12SN7GT (V16), and the vertical output tube, Type 12SN7GT (V17). Tube (V17) furnishes saw-tooth deflection current to the vertical coils in the deflection yoke through the vertical output transformer.

Contained also in this chassis are the horizontal synchronization separator and clipper tube, Type 12SN7GT (V18), the horizontal oscillator and discharge tube, Type 12SN7GT (V19), the horizontal saw-tooth voltage amplifier tube, Type 12SN7GT (V20), the horizontal output tube, Type 807 (V21) which furnishes a saw-tooth current wave to the horizontal deflection coils through the horizontal-output transformer, the controlled damper tube, Type 71A (V31), and the high-voltage rectifier tube, Type 8016 (V22). This tube rectifies the "kick-back" voltage generated in the horizontal output transformer during the return line time to provide high-voltage (direct-current) required by the second anode of the kinescope.

The bias amplifier and rectifier tube, Type 12SN7GT (V24) which amplifies and rectifies the horizontal saw-tooth voltage to provide a negative bias voltage, and the two voltage regulator tubes, V25 and

V26 (Type VR-105), which provide regulated voltage to the plates of the several oscillators (where stability of frequency is important) and to the synchronizing separator and clipper tubes, are also located on this chassis.

Antennas for use with this equipment are not discussed herein, but will be covered in a separate paper at a future date. However, one of the frequently-used antenna types was a quarter-wave vertical rod antenna working against two quarter-wave ground rods extending on opposite sides at the base of the vertical radiator. A matching unit was used to transform the impedance of the antenna to 75 ohms in order to match the transmission line.

Tests of the overall equipment have already been described in some detail.[2] These tests involved extensive work over a long period by both government and company engineers. Although a number of operating difficulties were experienced (see Part II) which were corrected in later equipment, it was found that the results obtained were in substantial agreement with the original design objectives.

More than 500 equipments of this type were built, most of them by the production engineering group in charge of A. Wright and K. A. Chittick. Later this group redesigned the Block I equipment in order to make it more suitable for quantity production. The new design was called Block III equipment and was used by the armed services under actual combat conditions.

PART II

TRANSMISSION PROBLEMS IN AIRBORNE TELEVISION SYSTEMS

Introduction

Early flight tests of Block 1 equipment indicated that the demand for satisfactory performance under conditions of actual operation in military aircraft imposed extremely severe requirements on the operating characteristics of the equipment. Normally after take-off of the transmitting plane it was impossible to readjust any controls to compensate for changes in power-supply voltage, light conditions, signal strength, interference, temperature, and vibration. This frequently resulted in inferior performance because of poor picture quality, poor synchronization or both.

Investigation showed that some of the principal defects were caused by the following conditions:

(1) Microphonics.

(2) Power supply voltage variations.

(3) Interference caused by ignition-type noise or by other carrier frequencies, particularly those having radar modulation.
(4) Unstable synchronization.
(5) Variation in light conditions on pickup tube.
(6) Insufficient signal at receiver.
(7) Multi-path transmission and frequency modulation of the transmitter master oscillator.

Considerable time and effort has been devoted to these problems and some progress has been made toward satisfactory solutions. There follows an account of the work done to date.

1. *Microphonics*

There are two basic methods for approaching this problem:

(a) by mechanically isolating the unit involved from noise and vibration; and
(b) by providing electrical means for reducing the resultant effects in the units themselves.

Where especially severe conditions are encountered, a combination of the two methods is probably necessary in order to obtain satisfactory performance. It is hoped that the improvements developed in electrical circuits will make it possible to operate under normal conditions without elaborate mechanical isolation.

The most troublesome microphonics originate in those tubes in the video amplifier which are followed by maximum low frequency gain. There has been some improvement in performance resulting from better tube construction and it is hoped that even more rugged tube types will be available for future equipment. A considerable improvement in performance was also obtained by applying the clamp circuit principle used in pre-war orthicon equipment to the video amplifier. This circuit effectively allows the gain at low frequencies in the video amplifier, where microphonics are normally most troublesome, to be reduced almost to zero but does not interfere with the reproduction of low-frequency signal components. Circuits of various types employing this fundamental principle were tried and flight-tested and a relatively simple circuit easily adaptable to Block equipment was finally evolved.

In Figure 11 are shown modifications of some of the circuits in a Block I camera-transmitter unit. One of these modifications is the addition of a clamp circuit in the video amplifier. The double diode V23 receives pulses from the horizontal output transformer in such a manner that positive pulses are impressed on the plate of one diode and negative pulses on the cathode of the other diode. The negative pulses are also fed to the iconoscope grid in order to provide a constant reference voltage from the iconoscope during the horizontal blanking period. A complete explanation of the operation of the clamp circuit may be found in an article by C. L. Townsend.[3]

More effective shock mounts than the original design, were developed by the Robinson Company and contributed to a marked reduction in noise signal originating in microphonic tubes. However, it was apparent that under some conditions microphonics originated from noise reaching the tubes through the walls of the unit itself. It was possible to reduce this effect considerably by enclosing the unit in a separate container lined with sound absorbing material. In such cases it was also necessary to provide forced ventilation in the unit to prevent overheating.

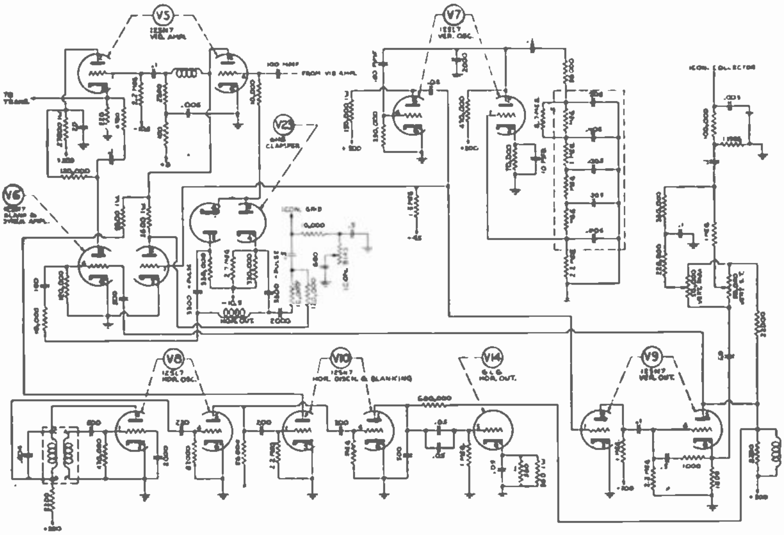

Fig. 11—Block I camera transmitter modifications.

2. *Power Supply Voltage Variations*

Occasionally the primary direct-current voltage source supplying the equipment was reasonably constant but frequently, especially in aircraft, the supply voltage was subject to considerable variation. This made the proper adjustment of equipment quite difficult since the voltage with the plane engine stopped or idling was usually very much lower than that obtained in flight. In order to overcome this difficulty some type of voltage regulator appeared to be essential. Automatic voltage regulator circuits, commonly used in other types of equipment

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[3] C. L. Townsend, "The Clamp Circuit," *Broad. Eng. Jour.*, Vol. 12, No. 2, pp. 5-8, Feb., 1945; No. 3, pp. 5-7, March, 1945.

for the $B+$ supply, were adapted for use in Block equipment. In addition to providing a constant output voltage over a considerable range of primary supply voltage, this circuit has the advantage of providing a low impedance source for the various electrical circuits in the equipment, thereby eliminating to a large extent the necessity for large electrolytic by-pass capacitors normally used for preventing coupling at low frequencies between various circuits.

Some undesirable effects caused by variation in the primary voltage supplying heater power to the equipment still remained. It was possible to reduce these effects considerably by the use of devices such as ballast lamps, which provide a relatively constant current for a wide range of input voltage. This combination of a regulated $B+$ supply and ballast lamps on the low voltage supply resulted in satisfactory performance but the efficiency of these circuits was rather poor because of the power loss in the regulator tubes.

Recently it has been possible to obtain dynamotor power supplies which will supply a constant output voltage for the $B+$ supply over a considerable range of primary voltage change. These have the advantage of being much more efficient than the other devices described but it may still be desirable to use a regulated $B+$ supply in order to obtain a low impedance source for preventing crosstalk. Since this $B+$ regulator circuit does not need to be designed to accommodate a large variation of input voltage, the tubes can be designed to operate much more efficiently than in customary regulator service. Another possibility is the use of a circuit quite similar to the normal regulating circuit which will take out low frequency disturbances and provide a low impedance source but which will not regulate the direct-current voltage itself.

3. *Interference*

On some of the first flight tests it was discovered that radar equipment operating near the Block equipment carrier frequency would completely eliminate the picture signal appearing on the receiver. This would occur when sufficient interfering signal reached the automatic gain control circuit at the receiver to decrease the gain of the intermediate-frequency amplifier enough so that none of the desired signal was visible on the kinescope. A long series of experiments and flight tests were made in an effort to reduce this effect as much as possible.

For radar-type interference it was noticed that while the amplitude of the signal was large, the energy content was quite small. This characteristic was used to improve the operation of the automatic volume control. Under normal conditions the automatic volume control is controlled by the amplitude of the synchronizing pulses appearing at the output of the receiver second detector. Since the energy content of the synchronizing pulses is normally much greater than that of the radar pulses it was possible to alter the constants of the automatic volume control circuit in such a way that it operated to a large extent on signal energy rather than peak amplitude. This type of circuit is called the low impedance type automatic volume control circuit and is illustrated in Figure 12. It is capable of permitting satisfactory operation of the receiver under interference conditions from radar at least 100 times as severe as was formerly possible. Another type of "low impedance" automatic volume control circuit is illustrated in Figure 13. The input circuit to the automatic volume control diode is tuned to the line frequency in order to discriminate between the synchronizing pulses intended for rectification and the interfering radar pulses. This circuit, however, showed little improvement over the preceding one since shock excitation of the tuned circuit by radar or noise pulses also contributed to the automatic volume control bias developed.

A still further improvement was made possible by introducing pulses from the horizontal output circuit into the automatic volume control circuit in such a way that signals can reach the automatic volume control detector only during a short interval corresponding to slightly more than the period of the horizontal synchronizing pulse. This circuit is called a "keyed automatic volume control circuit" and prevents all the interference occurring during the interval in which picture signal is transmitted from affecting the operation of the automatic volume control. An improvement of approximately 10 to 1 was noted over the normal low impedance automatic volume control circuit. This type of circuit is shown in Figure 14.

Suitable limiter circuits on the signal output to the automatic volume control circuit are also helpful in improving performance. However, these circuits require rather careful adjustment so that the limiting level is always somewhat higher than the amplitude of the synchronizing signal itself. This also means that the automatic volume control characteristic must be quite flat so that with any reasonable value of signal input the synchronizing level will not exceed the limiter adjustment.

4. *Instability of Synchronization*

Another effect of noise and interference was to disturb seriously the synchronization of the picture at the receiver. In many cases this resulted in a loss of picture intelligence much greater than that caused by the presence of the interference in the picture signal itself. During

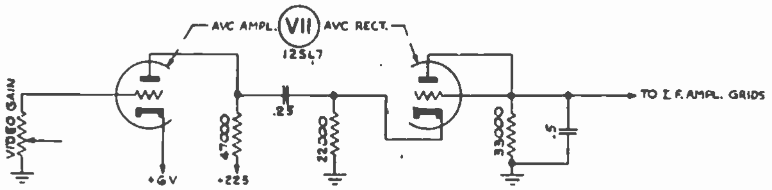

Fig. 12—Low impedance automatic volume control circuit.

flight tests it was frequently necessary for the operator at the receiver to adjust the hold controls in order to keep a stationary picture on the kinescope screen.

A fundamental difficulty with the first Block equipment lay in scanning oscillator instability. Both the line and frame frequency scanning oscillators were cathode-coupled multivibrators whose frequency was a function of a large number of variables including the supply voltage, ambient temperature, and all the resistors and capacitors used in the feedback circuit. This type of oscillator has the advantage that it produces pulse wave shapes which can be used directly for blanking and synchronizing signals. Various other types of relaxation oscillators were tried in an effort to find some whose frequency stability was superior to those in use. However, none of them exhibited a frequency stability comparable to that obtained from a sine wave oscillator.

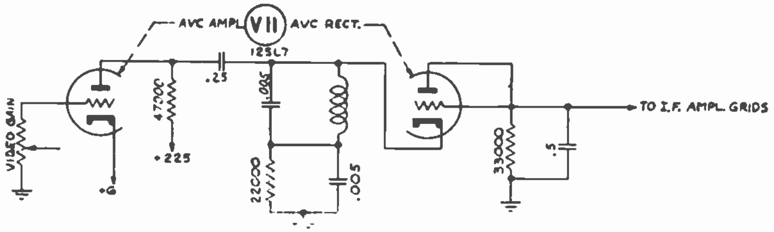

Fig. 13—Low impedance automatic volume control circuit with tuned circuit discrimination.

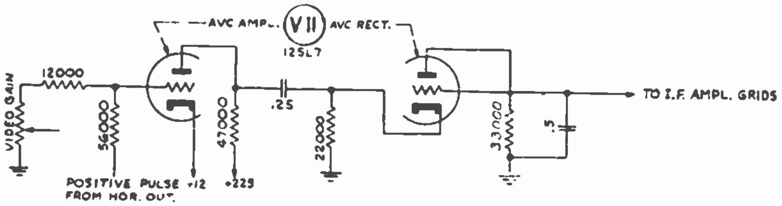

Fig. 14—Keyed low impedance automatic volume control circuit.

It was finally decided to use sine-wave type oscillators for both line and frame frequency scanning circuits and relatively simple shaping circuits were developed to obtain the necessary blanking and synchronizing pulses from the wave shape obtained from the sine wave oscillator. The horizontal oscillator was of the inductance-capacitance type with an adjustable powdered iron core for tuning the circuit to the proper frequency. The vertical oscillator was of the resistance-capacitance network type. This was used in place of a conventional inductance-capacitance oscillator because of the large size of the inductance required for a 40-cycle oscillator of the latter type. The stabilization of these scanning frequencies at the transmitter resulted in a considerable improvement in performance since the receiver synchronizing circuits did not have to compensate for any changes of frequency at the transmitter. It was also possible to adjust two different camera transmitter units to have exactly the same scanning oscillator frequencies so that a receiver could be tuned from one transmitter to another without requiring readjustment of the synchronizing hold controls. Figure 11 shows modifications made in the scanning oscillator circuits of a Block camera transmitter unit in order to permit the use of stable oscillators.

One-half of the 12SL7 (V8) is the horizontal frequency oscillator. The other half is an amplifier and clipper of the pulse-wave form appearing across the 2200-ohm resistor in the plate circuit of the oscillator. The plate circuit of the pulse amplifier is coupled to both triodes of the 12SN7 (V10). One triode is the horizontal discharge tube which develops a saw-tooth in its plate circuit for the horizontal output tube. The other triode is the horizontal blanking amplifier supplying blanking to the 12SN7 clipper tube (V5) in the video amplifier. One-half of the 12SL7 (V7) is the vertical scanning oscillator; the other half is an amplifier stage which amplifies a vertical pulse obtained by differentiating the plate voltage wave of the oscillator. One-half of the 12SN7 (V9) is the vertical discharge tube which converts the pulse to a saw-tooth wave form. The other half of (V9) is the vertical output tube which feeds the output transformer. Vertical shading voltages are also obtained from the plate and cathode voltages of this tube. Vertical blanking signal is obtained from one-half of 12SN7 (V6). In the plate circuit the horizontal and vertical blanking signals and the video signal are mixed together to be impressed on the connected to the horizontal and vertical output circuits.
clipper grid. The other half of (V6) is the synchronizing signal amplifier which obtains its grid voltage from differentiating circuits

However, even with this improvement severe noise and interference conditions caused the picture on the receiver to be unstable. Radar interference of insufficient intensity to interfere with the picture signal was enough to cause the picture edges to be quite irregular so that a considerable amount of picture intelligence was lost from that cause. To improve this condition the automatic frequency control principle was applied to the scanning oscillators in the receiver. This circuit uses a phase detector which operates in such a way that it maintains the pulse from the synchronizing signal and the pulse from the horizontal output circuit within a very short time interval of each other. The output of the phase detector operates on a reactance tube which changes the frequency of the scanning oscillator in such a way that the proper phase relationship is maintained. This circuit has a sufficiently long time constant so that noise pulses of short duration have practically no effect on the bias supplied to the reactance tube.

In this way the average frequency of the scanning oscillator at the receiver is made the same as the average frequency of the scanning oscillator at the transmitter. If, for instance, both horizontal oscillators maintain this constant frequency over a period greater than one vertical frame then the picture edges will automatically be exactly vertical. The longer the time constant of this circuit the better noise immunity it will have for low frequency interference. However, the long time constant circuit requires an appreciably longer time to come into synchronism if the signal is temporarily lost. Another difficulty noticed with the long time constant circuit on the horizontal oscillator is the lateral movement of the picture caused by rapid changes in the path length of the signal reaching the receiver. This happens only when the receiver is obtaining its incoming carrier signal over two separate paths, one of them usually a direct path and the other a reflection from the ground or some large object. The best value of the time constant of the horizontal automatic frequency control is a matter of compromise. Experiments indicate that an optimum value will permit the horizontal receiver scanning oscillator to stay in synchronism with the transmitting scanning oscillator even though a synchronizing signal may be lost for one complete frame and still allow a reasonably fast pull-in after temporary loss of synchronizing signal. The time constant for the automatic frequency control circuit controlling vertical oscillator speed is more difficult to adjust. In order to obtain superior noise immunity over the standard lock-in circuit the time constant must be made long compared to one vertical frame. This makes the recovery time after a temporary loss of signal too slow to be satisfactory. What is necessary is a circuit which has a long time-constant while the vertical oscillator is locked in but permits a short recovery time when the oscillator is out of step. One outstanding advantage of this circuit arrangement for both horizontal and vertical oscillators is that it practically relieves the operator at the receiving location from the necessity of adjusting the receiver hold controls once they have been properly adjusted.

Figure 15 shows the horizontal and vertical lock-in circuits of a Block 1 receiver modified to include automatic frequency control of both scanning oscillators. Referring to Figure 15, one-half of the 12SL7 (V24) is the horizontal sine-wave oscillator. The 6AC7 (V20) is a reactance tube operating in a conventional circuit. A change in the direct-current bias of this tube causes it to change its gain. Since the plate current of the tube is almost 90 degrees out of phase with the plate voltage due to the resistance capacity network in its grid circuit, the change in gain causes a change in frequency because of the change in reactive current through this tube which is in shunt with the oscillator tank circuit. The bias which operates the reactance tube and thereby changes its frequency comes from a 6H6 discriminator tube (V19) which receives signal from a synchronizing amplifier and horizontal synchronizing separator (V14). The plate circuit of the separator has a transformer which supplies a synchronizing signal of opposite polarity to the two discriminator diodes. A horizontal pulse from the horizontal output is introduced into the discriminator circuit in such a way that the same polarity pulse appears on both diodes. When the phase relationship between the incoming synchronizing pulses and the pulses from the horizontal deflecting circuit changes, the direct-current output from the discriminator changes. If the phase shift is in one direction, the direct-current output is positive; if in the opposite direction, the direct-current output is negative. This bias applied to the reactance tube tends to shift the oscillator frequency just enough to keep the synchronizing pulse and the horizontal output pulse in proper phase relationship. The time-constant of the resistance-capacitance network in the output of the discriminator determines the speed of response of the horizontal automatic frequency control system. If this time constant is slow, the oscillator will hold its frequency over longer periods of absence of synchronizing signal than is possible with a short time constant. However, if the oscillator is out of synchronism either at the time the receiver is turned on, or because of excessive noise or lack of signal, the time of recovery is much slower with the long-time constant circuit than with the other.

It may be noted that the horizontal discriminator circuit is not balanced with respect to ground. Test indicated that it was highly desirable to have the discriminator output nearly zero when no input

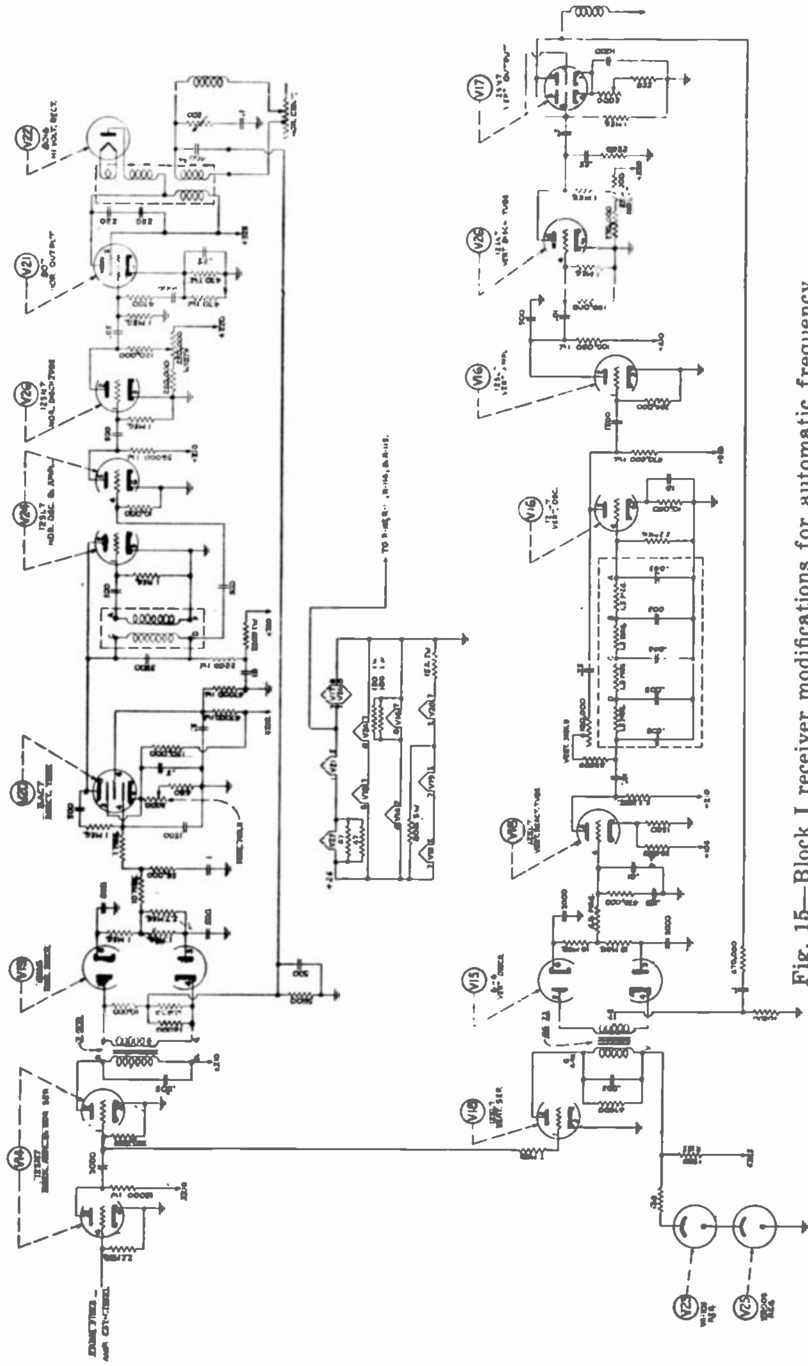

Fig. 15—Block I receiver modifications for automatic frequency control scanning oscillator control.

signal was present. Since a pulse is used from the horizontal output instead of a saw-tooth, the discriminator output will not be zero unless the circuit is unbalanced with respect to ground. If this condition is not fulfilled, the recovery or pull-in time is quite long because of the necessity for charging the one-microfarad capacitor through the 10-megohm resistor to a definite direct-current potential.

The speed control for the horizontal hold adjustment is a rheostat in the cathode circuit of the reactance tube (V20). A powdered iron plug in the oscillator coil acts as an auxiliary speed control. The cathode voltage adjustment also makes it possible to change the sensitivity of the automatic frequency control system. When the control is near the lowest value of cathode voltage possible, the sensitivity is greatest. With this adjustment the horizontal oscillator will lock in about 400 cycles above and below the 14,000 cycle normal frequency. At the highest value of cathode voltage, this range drops to about 100 cycles each way.

The automatic frequency control circuit for the vertical oscillator is quite similar to the circuit used for the horizontal oscillator. The oscillator is one-half of a 12SL7 (V16) with the resistance-capacitance feedback circuit mentioned before. Here the reactance tube changes the impedance to ground of one leg of the resistance-capacitance network. By changing the bias on the grid of one-half of the 12SL7 (V18) the plate impedance of the tube varies and therefore the impedance in series with the .01-microfarad capacitor to ground also changes, thereby changing the frequency. The vertical separator (V18) and discriminator (V15) have the same circuit arrangement as the equivalent horizontal circuits. Vertical synchronizing pulses appear on the two discriminator diodes of the 6H6 with opposite polarities and a vertical sawtooth from the vertical output has the same polarity on both diodes. The resultant direct-current voltage which is generated when the diodes are unbalanced tends to shift the oscillator frequency so that the correct phase relationship between the vertical synchronizing pulses and the vertical saw-tooth output is maintained. The time constant used here is again a compromise between the time required for the absence of synchronizing signal to cause the oscillator to fall out of synchronism and the time for recovery of synchronism once the oscillator is out of step. A choice is more difficult here than with the horizontal circuit because of the low frequency of vertical scanning. The time constant should normally be quite long compared to the time required for one picture so that the vertical automatic frequency control circuit, in effect, obtains its synchronizing information from a large number of vertical pulses. However, if this time constant is very long, and the vertical oscillator does lose synchronism so that the vertical blanking pulse appears near the center of the picture, a number of seconds may be required before the automatic frequency control circuit can reestablish normal synchronization.

The vertical hold control is a rheostat in a resistive element of the resistance-capacitance oscillator network. With a reasonably strong signal, the automatic frequency control circuit will keep the vertical oscillator in step over a range of plus or minus two cycles variation from the normal 40-cycle frame frequency.

5. *Variation in Light Conditions*

The difference in light conditions prevailing at even moderate altitudes compared to the light conditions on the ground is usually extremely great. The normal characteristics of a scene observed from an airplane during daylight include an extremely high light level and very low contrast, caused by either haze or smoke in the atmosphere. This complicates the problem of adjusting the equipment on the ground for satisfactory performance in the air. For the early tests of Block 1 equipment, a test bench was developed for setting up the camera transmitter unit on the ground and was arranged so that a slide with very little contrast could be projected on the iconoscope mosaic at a light level comparable to that obtained during flight. This, in general, improved performance considerably over that obtained when preliminary adjustments were made for a scene on the ground.

A number of experiments were made in an effort to improve the signal output from the iconoscope under high light level and low contrast conditions. It was discovered that many of the earlier iconoscopes would saturate at very-high light levels and would produce considerable noise output and very little signal. In many cases improved performance was obtained by a reduction in the aperture of the lens used in the camera. The effect of various types of filters was studied but the results were somewhat disappointing since the insertion of a filter which cut down the effect of haze also reduced the signal output from the iconoscope to the point where hiss noise from the first video amplifier tube became quite noticeable. In most cases it was found that a Wratten #25 orange filter would improve picture contrast without seriously increasing noise.

An investigation of the saturation phenomena in the iconoscope by the tube engineers resulted in a slight modification in the tube which permitted better operation at high light levels. A high-light test was also included in the iconoscope test specifications which would insure satisfactory operation.

Two methods for obtaining a more nearly constant signal output from the iconoscope were also investigated. One method was the development of an automatic volume control circuit which would change the gain of the video amplifier as a function of its signal output. This was somewhat difficult because of the spurious signal delivered by the iconoscope during the blanking intervals and the spurious signals commonly referred to as "dark spot" and "edge flare".

Two types of automatic volume control circuits were developed. In one circuit only the high frequency components of the video signal were used to operate the automatic volume control system. This operated satisfactorily as long as there was a sufficient amount of fine detail in the picture from which high frequency signals could be obtained. A circuit somewhat more satisfactory was made by keying the automatic volume control circuit in such a way that only the picture information in the central part of the picture was used to operate the automatic volume control. In actual flight tests these circuits showed very little, if any, advantage over the normal circuit because the usual light conditions normally required all of the video gain available consistent with a reasonably satisfactory signal to noise ratio. An automatic iris control was also suggested for this purpose and was actually developed, although primarily for use with the orthicon tube.

There were two other possibilities suggested for improving the performance of the iconoscope under flight conditions: (1) the possibility of introducing automatic shading circuits; and (2) the possibility of controlling automatically the beam current of the iconoscope itself. These investigations were discontinued because it was felt that the development of the small orthicon and the image orthicon would eliminate the necessity for circuits of this type.

6. *Low Signal Strength*

With a carrier power output of approximately 15 watts from the transmitter of Block I equipment the reliable operating range from aircraft to ground was approximately 10 miles. Operation beyond this range was possible but the signal would sometimes be lost because of the change in position of the plane. An increase in this range was considered very desirable but it was felt that considerable increase in transmitter power would be necessary for any appreciable improvement in performance. This would, of course, mean a much larger unit at the transmitter with a corresponding increase in weight and power drain. It was decided first to improve the signal-to-noise ratio of the receiver as much as possible before attempting to provide more transmitter power. It was also considered desirable to provide the receiver operator with a tuning adjustment so that interference effects from carrier frequencies close to the desired frequency could be minimized as much as possible. A new type of head-end tuner for the receiver was developed which had the following advantages:

(a) an improvement in signal-to-noise ratio of at least 2 to 1 over previous circuits;

(b) considerable improvement in the radio-frequency selectivity and prevention of excessive oscillator radiation by the inclusion of a radio-frequency stage; and

(c) better performance, as evidenced by flight tests of a receiver with the improved tuner, not only because of the improved sensitivity but because it was possible by slight changes of tuning to eliminate or greatly reduce interference near the edge of the channel. (This was impossible with the earlier fixed-tuned type receivers.)

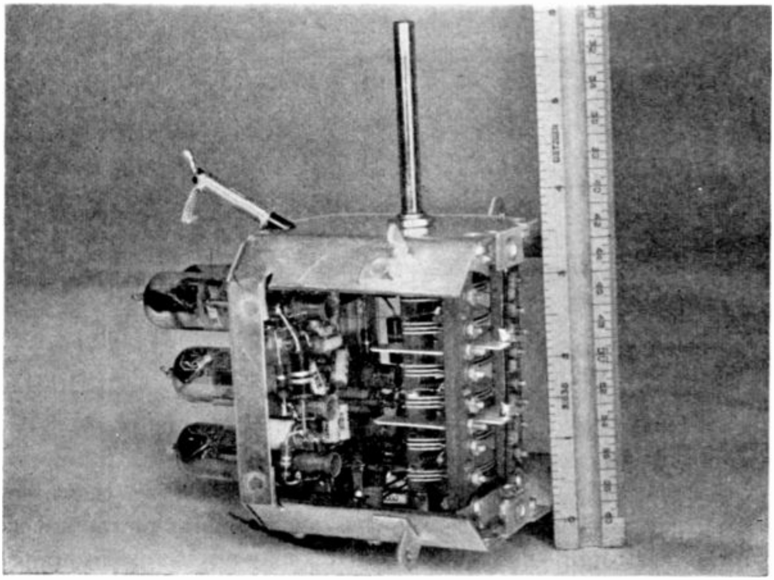

Fig. 16—Block I receiver tuning unit—top cover removed.

It is estimated that the addition of this tuner increased the effective operating range of the receiver to at least 25 miles. A photograph of one of these tuners is shown in Figure 16 and a schematic diagram is shown in Figure 17.

The general principle of this tuner, which has wide application, is that of connecting appreciable fixed inductance between the low potential side of the variable capacity and ground or virtual ground. Thus, inductance is added to the circuit as the variable capacitance is increased. This action occurs most effectively when the minimum capacitance is small compared to the distributed capacitance immediately external to the capacitor. Under this condition, the circuit current follows a shorter path when the variable capacitor is at minimum than when at maximum.

With ideal capacitor construction and with the largest useful auxiliary inductance, the maximum possible tuning ratio should increase from two to three by the addition of the inductance. In practice, of course, the maximum tuning range of 2 cannot be realized because of limited capacity range. Hence, the addition of inductance in the capacitor may increase the tuning range by more than the theoretical maximum factor of 3 to 2. In the actual application, the frequency

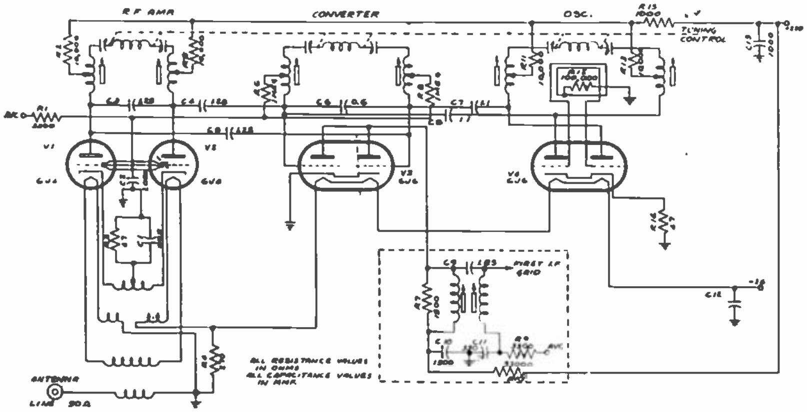

Fig. 17—Block I receiver tuning unit—schematic diagram.

coverage was doubled by the inductance. However, the tuning ratio has been limited to less than 2 in order to reduce the crowding at the low-frequency end of the dial when semi-circular plates are used. "Midline" or straight-line-frequency plates may also be used to improve dial linearity. Inasmuch as the tube capacitance at one end of the circuit is in series with the tuning capacitor at the other end, there is little to be gained by using a tuning capacitor with a greater maximum than two or three times the tube capacitance.

7. *Multipath Transmission*

One of the troublesome problems encountered in the transmission of television signals from one plane to another is the effect of two out-of-phase signals of nearly the same field strength reaching the receiver simultaneously. This frequently causes two pictures to appear simultaneously on the face of the picture tube. The one having lower intensity is sometimes called a ghost. Synchronization difficulty is also experienced because of the presence of two signals. Considerable improvement can be obtained by using directional antennas and by locating them in such a way that reflections from the ground are minimized.

The multipath problem becomes considerably more serious if frequency modulation is present on the transmitter carrier especially if the frequency modulation is caused by video signal. Peculiar interference patterns are caused by the combination of multipath transmission and transmitter frequency modulation which may be so strong as to destroy entirely the usefulness of the picture. When transmitting from plane to ground the difficulty usually arises because a second signal arrives at the receiving location, reflected from a large building or other similar object in the area near the receiving location. When the transmission path is from plane to plane the second signal at the receiver is the one arriving by virtue of a reflection from the ground. A similar effect can be obtained even though the receiver is at a ground location if the antenna is located high enough so that a ground-reflected signal will have a phase difference, compared to the direct signal, sufficient to cause trouble. How this may come about is explained briefly in the following paragraphs.

Consider the elementary propagation system consisting of a transmitter (T), a receiver (R), the propagation medium, and the earth's surface (E). In the simplest case, the signal is transmitted to the receiver, R, from the transmitter, T, over just two paths, one being the direct path TR and the other the indirect path, or path of reflection TER. The resultant signal at R is the vector sum of the two.

Obviously, the direct signal from T will arrive at R ahead of the indirect signal, the time difference, t, depending upon the heights of the transmitter and receiver as well as the distance between them. Thus, if a simple rectangular pulse of very short duration as compared to t is sent from the transmitter, two pulses will be received, spaced t seconds apart. Now let us assume that the transmitter is sending out a continuous signal of frequency f_1 and after a period of time the frequency is suddenly changed to f_2 for a time much less than t after which the frequency again returns to f_1. No amplitude changes occur at the transmitter during the transition from f_1 to f_2 to f_1. In t seconds after the pulse of frequency f_2 has left the transmitter two signals will be received, f_1 and f_2. Both will be demodulated at the receiver producing a beat note having a frequency $f_1 \pm f_2$ but only during the times when different frequencies arrive by the two paths. The frequency $f_1 + f_2$ is too high to be accepted by the receiver, but $f_1 - f_2$ may be a very low frequency falling within the video acceptance band. When this is the case, the beat note will appear superimposed on the picture signal.

In actual television practice, large repetitive changes in modulation occur mainly at the beginning and ending of the horizontal blanking pulses. This means that, when the reaction from power amplifier to oscillator is appreciable, there is a beat note of large amplitude during the horizontal return time, or fly-back, with consequent vertical bar patterns appearing in the picture. Often, the disturbance is further complicated by the presence of not just one but several paths of reflection so that a corresponding number of beat notes can appear simultaneously and for varying lengths of time.

Because of this difficulty it is extremely important to keep frequency modulation of the transmitter to an absolute minimum. A buffer stage between the transmitter oscillator and modulator is almost an absolute necessity. It is also important that mechanical vibration of the oscillator circuit elements be kept below a level which will cause excessive frequency modulation. In general, it seems that more than 1 per cent frequency modulation can cause noticeable effects in the picture.

* * *

The development of Block I television equipment and the subsequent investigations carried out in an effort to improve the performance were the collective work of a large group of engineers working under the direction of R. D. Kell, in charge of television research at RCA Laboratories Division, and G. L. Beers, Assistant Director of Engineering in charge of Advanced Development at RCA Victor Division. The development and field test of this equipment would have been at best extremely difficult without the whole-hearted interest and cooperation of NDRC, Military, and Naval personnel associated with this project. This paper covers work done in whole or in part under the following contracts: W535sc238, NOs-86775, PDRC-29, NXs3405-A, and OEMsr-441, all with Radio Corporation of America.

TELEVISION—A REVIEW, 1946*

BY

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Summary—After reviewing the history of television very briefly the four problems attending its commercialization are discussed. These problems—transmission, networking, reception and pickup—are treated in detail. Developments made since 1939 are enumerated and the results which have produced today's commercial television equipment are considered.

TELEVISION had its start as a public service during the year and a half before the United States entered the war. That called a stop to further progress. Television has marked time during the war period, but now sufficient time has elapsed since the close of hostilities to allow commercial television to again get under way. Television is now growing and expanding and is taking its place among the other regular household facilities which we have come to regard as essential in the modern home.

Television has had a long and interesting history, and while it is not the purpose of this paper to discuss the background or principles of operation, since these have been adequately covered, a few items of background material will be reviewed briefly.

Shortly after some of the first principles of electricity and particularly photo-electricity were known, people conceived methods for achieving the results which we now call television. While it was possible, even before the break of this century, to make crude setups which illustrated the principle, these were not good enough to reproduce moving pictures at a distance. That had to await the coming of the electron tube.

Television made rapid progress in the latter part of the 1920's, but there was virtually an insurmountable obstacle ahead. The methods chosen at that time involved mechanical devices—revolving scanning discs, mirror wheels, etc.—all of which definitely limited how good the picture detail might be made. There were also severe mechanical limitations in the devices themselves. What was needed to carry the art beyond this impass was the conception of some new tools which would give us an electronic method of operation to replace the mechanical

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* Decimal Classification: R583.17.

processes used in television up to that time. Specifically, we needed an electron tube in the receiver, which was capable of producing directly on its optical surface a light image corresponding to the original scene. Zworykin's kinescope or picture tube was the first electron tube to fulfill this specification. Television was quick to take advantage of this new development, but it was further handicapped by the lack of a similar device at the transmitter. This handicap was soon overcome, however, by Zworykin's iconoscope and the Farnsworth dissector tube. With the development of the iconoscope, Zworykin and his associates created the cornerstone of the structure which has become electronic television.

The conception of this electronic method was followed by a period of active reasearch in which the tools were made better and sharper, so as to serve more effectively the purpose for which they were created. There followed, then, consideration of a number of factors related to over-all system planning, such as the practical aspects of building a system using the new tools, the evaluation of the work that had been done by a number of people both in this country and abroad, and the assembly of all of these ideas and tools so as to make the best use of the technical facilities available. It was then necessary to take the resulting system into the field and put it to a practical test. This was done. After the completion of these tests, the technical personnel again reached agreement among themselves as to the fundamental standards, such as the number of scanning lines, method of synchronizing, method of transmitting sound, etc. This standardization was necessary so that the transmissions from all stations would be receivable on any television receiver. These operating standards were modified from time to time as additional operating experience was obtained. All of these changes in the operating standards were in the direction of improving the over-all television system.

At the completion of this work, the standards had to be approved and set up by a Federal agency, the Federal Communiciations Commission. In this work the Commission had the cooperation of all of the technical personnel of the television industry.

Television then moved into its commercial phase, which was abruptly halted by the war. All television developments were then turned toward objectives of a military nature.

During the short period of commercialization, study of the television systems problem indicated that a satisfactory commercial television service would be dependent upon the successful solution of four important problems. These four problems are represented by the four sections of Figure 1.

The first of these problems concerns the television broadcasting transmitter in a given city. In this country, broadcasting is done on a competitive basis. It is desirable, under such a system, to have a number of transmitters in each of the service areas. This means that in a city or metropolitan area such as New York, there should be several television transmitters. The radiation from each television transmitter, however, occupies a space in the radio spectrum, several times as wide as the whole of the region now assigned for standard sound broadcasting, so, if the there are to be several transmitters, a wide expanse of radio frequency channel space is required. This means that it is necessary to use very high radio frequencies. In the years just preceding the war, it was possible to build television transmitters for channels 1, 2, 3, 4, or 5, but in the case of channels 6, 7, 8, 9, 10, etc., the required transmitter tools were not available. This resolved itself into a research problem which would be basic to a transmitter tube and a transmitter which would permit operation with useful power on all of the frequencies which would be assigned for television broadcasting. A power of 5 kilowatts was chosen as a goal for this work. Figure 2 shows the tube resulting from this reasearch program. This tube has been engineered and is available to use in television trans-

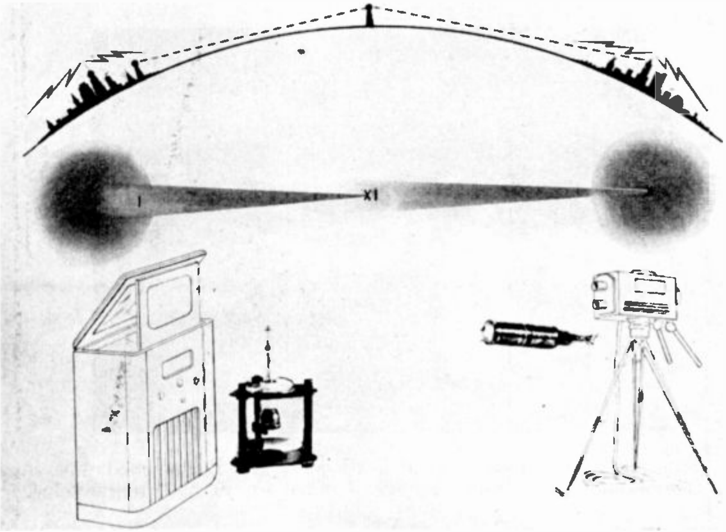

Fig. 1—Illustrations depicting the four principal problems of television: transmission at very high frequencies, television networking, production of larger and brighter pictures in the home and increased camera-tube sensitivity.

mitters which may be designed for any one of the channels which have been assigned by the FCC for commercial broadcasting. In the development of this transmitting tube, there were two conflicting factors: First, due to the high frequency operation, the tube had to be small, and, second, the requirement of high transmitter power meant that the tube must be capable of dissipating large quantities of heat. In order to meet this double requirement, it was necessary to develop new principles of the thermo-dynamics, as well as new principles of electronics.

These tubes made possible the construction of a transmitter at the frequency chosen—288 megacycles—having an output power of 5 kilowatts. This frequency of 288 megacycles is higher than the highest channel assigned for commercial operation (210 - 216 mc.) and therefore gives performance information with a safety factor. This transmitter was installed together with an appropriate omnidirectional antenna atop the Empire State Building in New York City for propagation surveys and reception tests. Figure 3 gives a view of this experimental transmitter, and Figure 4 shows the antenna as installed on top of the Empire State Building. The 288-megacycle antenna is the two-layer turnstile system at the very top. From field tests one may draw the following indications of performance in comparison with a lower frequency channel where experience is, naturally, much more comprehensive:

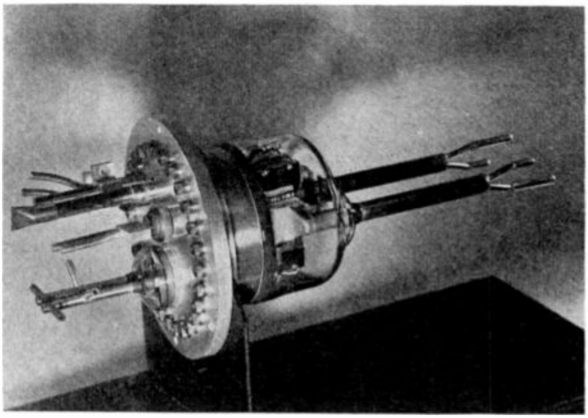

Fig. 2—Experimental model of a 5-kilowatt tube designed specifically for television transmission at frequencies up to 300 megacycles.

(a) In a comparison of 67 and 288 megacycles, it was found that "shadow" effects on the higher frequency are much more severe. This means that signals are reduced more on the higher frequency by buildings, hills, and other obstructions.)

(b) In a comparison of 67 and 288 megacycles where multipath conditions exist, it was found that multipath phenomena were about equally serious.

(c) In a comparison of relative field strengths, it was found that over actual urban regions the field strength on the higher frequency was, in general, less and subject to greater variations.

The second problem pertains to televison networks—the ability to tie together transmitters in several cities so that they may all participate in the same program. A network for television cannot make use of ordinary telephone lines, nor can it make use of the same kind of radio transmitters as are used for world-wide communication. Instead, radio frequencies which are many times higher than even those frequencies assigned to commercial television broadcasting must be used. These very high frequencies have the property that they do not bend around the earth's surface as do the longer waves, but travel in straight paths somewhat analogous to light.

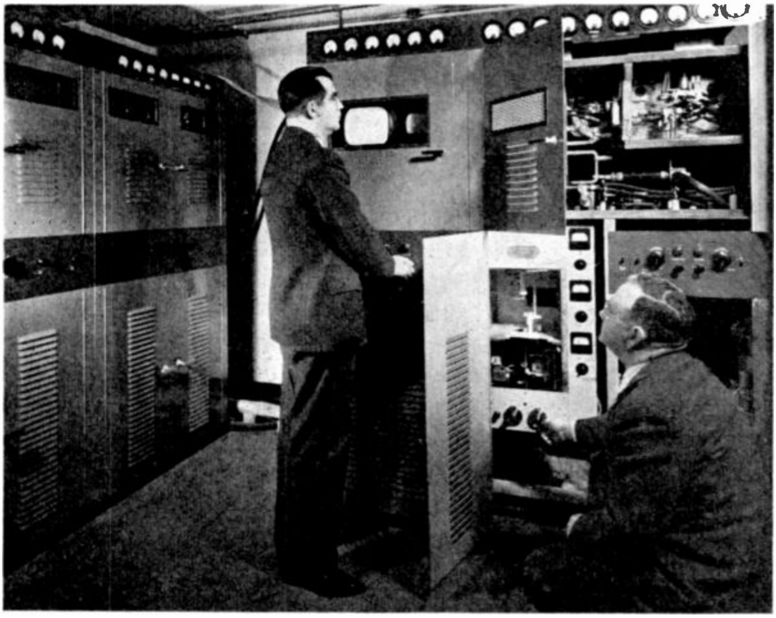

Fig. 3—288-megacycle television transmitter.

The radio relaying problem is also represented in Figure 1. At the first city, a broadcast type of service is indicated, and, in addition, a directive type of radio transmitter is shown, which directs the signal only in the direction of the second city. Due to the curvature of the earth, this directed signal will pass over the second city. In order to redirect and strengthen the signal so that it may travel on, there is placed, at an intermediate point, a radio receiver for the signal originating in the first city, and a repeating transmitter which will direct its signal to the second city. Thus the signal is carried around the curvature of the earth.

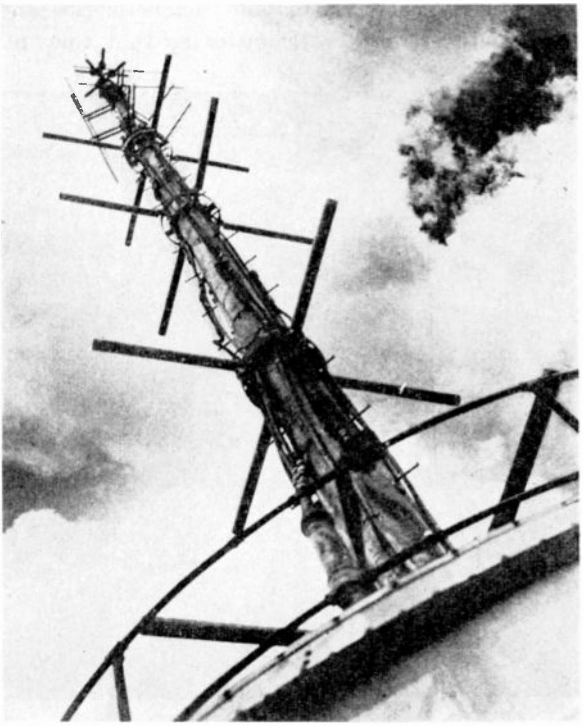

Fig. 4—Empire State Building antenna structure—288-megacycle antenna at the top.

A single repeater is shown. The number of repeaters may be increased almost indefinitely. For example, if the two cities are separated by as much as 300 miles, approximately 10 repeaters might be used to cover the distance.

Figure 5 is a map showing the arrangement of an experimental television relay circuit being installed by the American Telephone and Telegraph Company. The two terminals of the circuit are Boston and New York.

A second method of providing network facilities for television is to

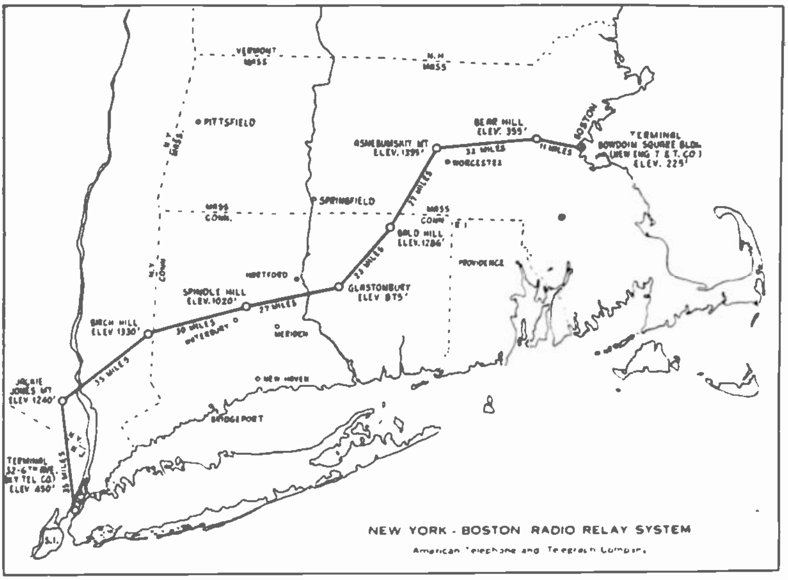

Fig. 5—Map showing arrangement of experimental radio relay circuits suitable for television.

use special coaxial cables. Figure 6 is a map indicating the extent of the coaxial cable installations in the United States. The filled-in lines are cables now installed; the heavy dotted lines are routes for installation contemplated so that possibly by 1950 those heavy dotted lines will be filled in. This map shows a primary network that extends up

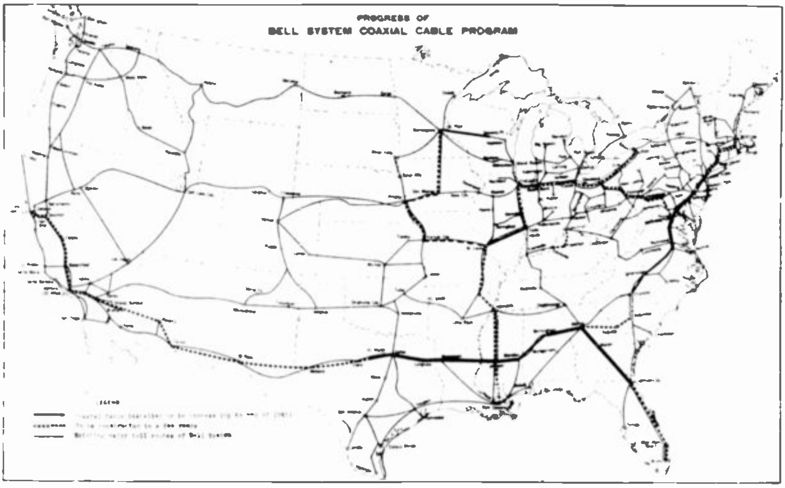

Fig. 6—Coaxial cable routes in the United States, suitable for networking television programs.

and down the east coast, with a branch that makes a loop around the central states. A major extension from the eastern and southern states to the West, joining the large metropolitan area of the west coast is also shown. When these circuits are completed, there will be facilities for a nation-wide network of television. This work is considered to be one of the important steps in the development of a national service. Sound broadcasting depends, to a large extent, on the national program circuits for its effective operation. The same will be even more true for television, because it is a means of distributing over a large audience a very large expense that will have to be borne by the broadcaster and advertiser who produce the television programs.

The third problem was the operating sensitivity of the television camera. A camera using an iconoscope or an orthicon was reasonably satisfactory in the studio, or other locations where the illumination could be increased to the desired level. In moving the television camera to points of interest on the street, in the theatre, or wherever one might wish to go, it was frequently necessary to bring along lighting facilities similar to those used in the making of motion pictures. The problems involved in overcoming this serious handicap to the flexibility of television programming formed the basis for the third problem. The goal for the project was set—a television camera so sensitive that it might be taken wherever people, themselves, might go, and see with satisfaction. This requirement can not be met by either the present day still or motion picture cameras. As the U. S. entered the war, it was thought that the elements of such a camera tube existed. The development of such a tube was of great military importance, and research continued. Because of continued research during the war period, production of such a tube began soon after the war.

This new camera tube, shown in Figure 7, is called an image orthicon, reflecting the names of two earlier tubes. It is considerably smaller in size than earlier tubes. This is important, because it is necessary to form an optical image with a photographic type lens on the photo-sensitive surface in the tube. To cover this small area, it is possible to use the excellent fast photographic lenses that have been developed for regular photography.

Figure 8 is a line diagram showing the elements of this new tube. The tube is operated in a magnetic field produced by the focusing coil. The field of this coil is parallel to the axis of the tube. The tube has a photosensitive surface indicated as the photo-cathode. In earlier camera tubes, the photo-sensitive surface usually had to perform other functions than that of merely emiting photo-electrons. In this tube only the one fundamental function is performed by the surface, thus

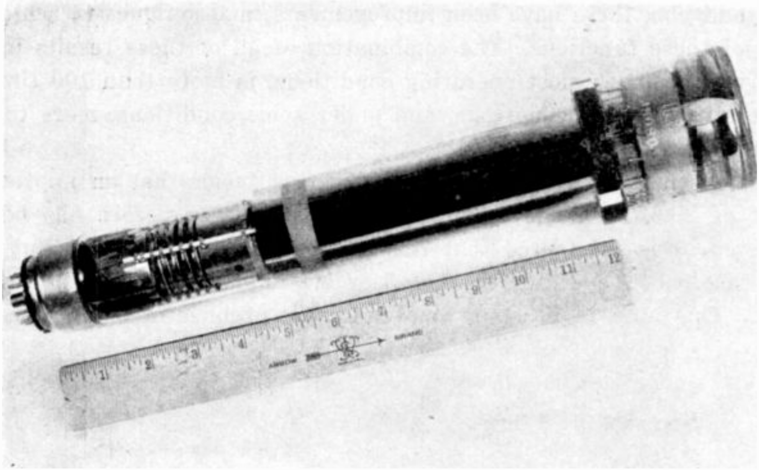

Fig. 7—The image orthicon—a sensitive television pick-up tube.

allowing optimum sensitivity. The photo-electrons are focussed by means of the magnetic field so as to form an electric charge image on the surface of the target screen. This target is an extremely thin plate of glass. The electron beam produced by the gun is caused to scan the back surface of the target screen. The returning electron beam is deprived of electrons in proportion to the requirements for neutralizing the charges produced on the glass target by the bombardment of the target by the photo-electrons. These returning signal electrons are collected and passed through several stages of secondary electron multiplication, which increases the magnitude of the output current by approximately 1000 times. With this method of operation, each part of the tube can be so constructed as to serve only that function for which it is designed, thus allowing operation under optimum conditions.

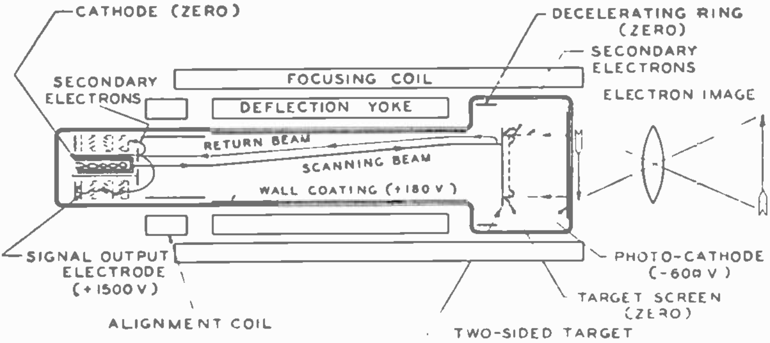

Fig. 8—Line diagram of the image orthicon.

In addition, there have been improvements in techniques to achieve each of these functions. The combination of all of these results in a tube which, under most operating conditions, is more than 100 times as sensitive as the iconoscope, and under some conditions, more than 1000 times as sensitive.

When one normally thinks in engineering terms that an improvement of 2, 3, or 4 times as good as what existed before has been achieved, it is considered that real progress has been made. But in this case one is speaking of 2 or 3 orders of magnitude of improvement. This tube, then, is the answer to that problem of being able to take a television camera wherever one wishes to go, and obtain a picture under practically any light conditions.

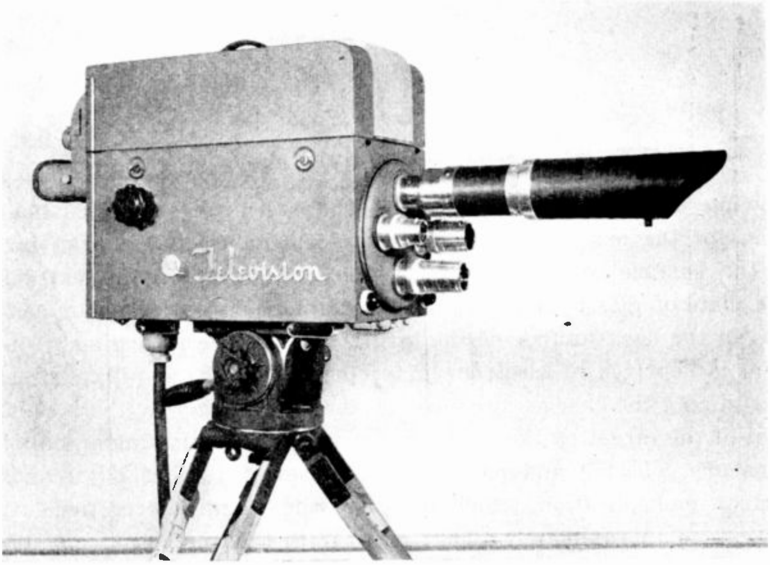

Fig. 9—The image orthicon camera.

The image orthicon is now in regular use and has extended tremendously the horizons for television programs. Figure 9 is a view of one of the cameras now available using the image orthicon.

The fourth problem was to produce larger and brighter pictures for the home receiver. Working in this direction, research has recently made an outstanding advance. This is illustrated in Figure 10. As the phosphor crystals on the face of a kinescope are excited by an electron beam, they emit light in all directions. Some of the light goes out through the tube face to the person viewing the picture. That is the useful part, but some of it is emitted toward the rear and is scattered around in the tube. That part is not useful, and is, in fact, detrimental because some of it is reflected onto the phosphor and reduces the picture contrast.

A method has been developed by which an extremely thin layer of aluminum is placed on the back surface of the phosphor. The film of aluminum is so thin that the electrons readily penetrate through the metal and into the phosphor, but it is also so smooth that it has a mirror-like surface. Thus the light that would ordinarily pass back into the tube now is reflected outwards and is added to the useful light. Because of this, the brightness of the tube is nearly doubled, and because there is no scattering within the tube there is also an increase in contrast.

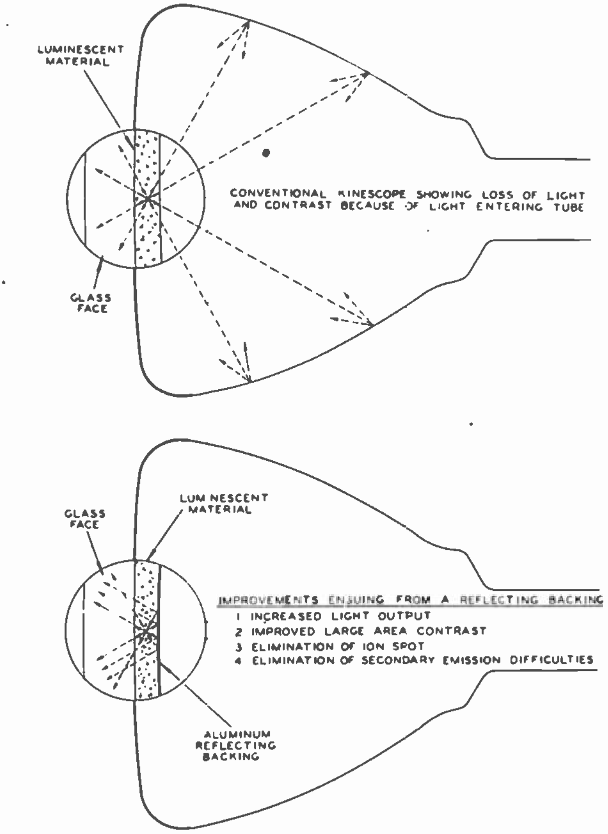

Fig. 10—Line diagram showing kinescope with and without metal-backed screen.

Also, the earlier kinescope, before the war, had an annoying defect.

They showed a yellowish-brown blemish at the center after some use. This was because the heavier slower-moving ions would get into the beam path, strike the phosphor, and cause deterioration. These slow-moving ions can not penetrate the metal film and an aluminized tube is free of this difficulty.

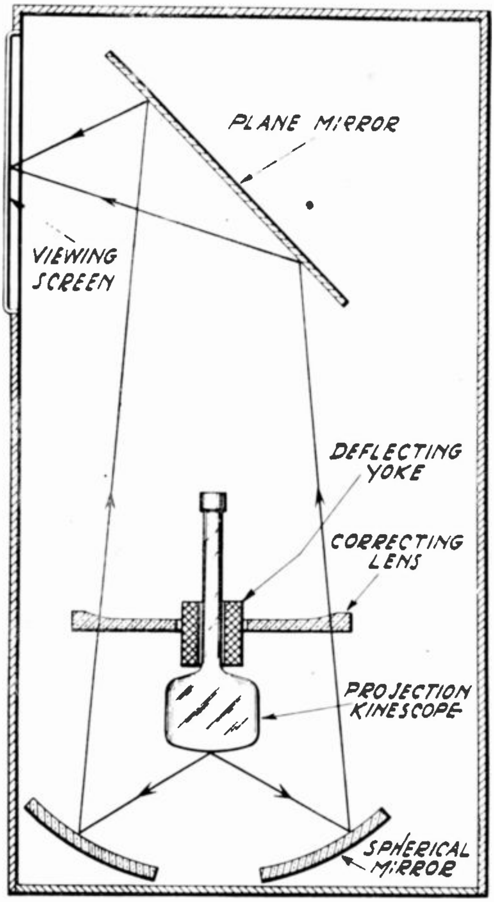

Fig. 11—Cross-sectional sketch of reflective optical system used in television receivers to produce an enlarged image by projection.

The solution to the larger picture, as indicated in Figure 11, involves a cathode-ray tube, and its associated optical system. The tube has a spherical face which is a part of the optical system, and on it is produced an extremely bright picture—too bright to be viewed directly. If we placed in front of this tube a conventional projection lens, we could produce an enlarged view of the image on the screen, but the optical efficiency of the system would be very low. Only a few per cent of the light produced by the tube would reach the screen. To overcome this difficulty, the television engineer has taken recourse to an optical method somewhat related to that used by the astronomer. The particular type of optical system used in television image enlargement focuses an image from the image source on a viewing screen located at a finite distance from the area at which the image is initially produced. This optical system consists of a large spherical mirror, which is capable of collecting much of the light developed by the image-producing tube. The mirror is shown at the bottom of the cabinet in Figure 11. The spherical mirror directs the light rays toward the final viewing screen, but between the mirror and the screen it is necessary to interpose a lens in order to produce a good optical image. This lens is called a correcting lens and is relatively weak, optically. Its surface is of the aspherical type, which makes it extremely difficult to grind from glass and much too expensive for the home type of receiver if produced by normal methods. In order to make such an optical system practical, it has been necessary to learn how to make a negative of the lens surface in metal, and how to use this as a mold to produce the lens from plastic material. With such an optical system, a very practical means now exists for producing large television pictures.

Figure 12 is a view of an experimental type of projection receiver, using the above-described optical system, which may be viewed by large groups. Figure 13 is a view of one of the table model television receivers now available on the market. It has a kinescope, or cathode-ray tube, 10 inches in diameter and produces a picture large enough to be viewed by small groups.

This discussion has indicated how definite solutions have been found to the four major problems that existed as television service was started in 1939. Now cameras are available which remove the limitations of programming; transmitters can be built for all the channels the Federal Communications Commission has allocated; ahead lies the facilities which will give television networks; and larger, brighter pictures can now be seen in the home. With reference to monochrome television, the research worker has passed the problem onward; little more is needed from him. The task now lies with others—to do the work of producing in the factories, selling in the stores, and programming in the stations. All of this is now under way; the research men have, for some time, been concentrating on the problems of color television, believing that the addition of color in a system that is both technologically sound and compatible with present television is the next and orderly step in the growth of television service.

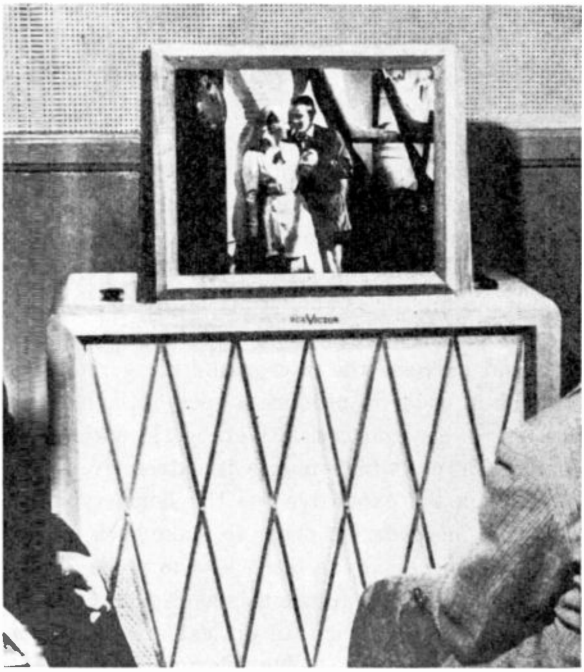

Fig. 12—A developmental model of a projection-type television receiver providing a picture 16 by 22 inches in size.

Television, from a technical point of view, is a reality, and for those who have been so active during the past ten, twenty or more years in the research and development of television, a goal has been

Fig. 13—Table model television receiver with a 10-inch screen.

attained. Now others must appear on the scene to assist in developing a public service. It is a service of great significance, and the horizon should be one of great expanse. There is a steep hill to climb, to get away from the shadows that limit vision and to seek to comprehend the complicated panorama that lies in view.

Television is going to affect men's lives in many ways. It will permit men to transport their eyes to distant points so that they can see things while they are taking place. It will give them new means of entertainment and new means for improving their education. It can assist merchandising practices and correlate manufacturing operations.

Everyone can participate in this effort, and by their acts of omission or commission will be determined the excellence and character of the service entrusted to television.

TELEVISION BROADCASTING—1946*

BY

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Summary—A general review of television broadcasting activities in 1946 is given together with discussions of factors which influenced this activity. The paper concludes with a consideration of the future of television broadcasting.

THE year 1946 could well go down in history as the year in which television broadcasting sped from around that mythical corner where rumor had placed it for the previous ten years. Monochrome television was ready to go, both technically and commercially, in 1941, and received the blessing of the Federal Communications Commission in June of that year. It started to emerge from around the corner when it was stopped in its tracks on December 7, 1941, when the United States became a combatant in World War II. For a four year period, hostilities arrested its commercial development.

Television itself went to war and played an important role in our military developments. It was the television techniques that gave birth to the now fabulous radar, many wonders of which materially aided in winning the war and saving the lives of countless thousands of American citizens.

During the war, television broadcasting was not entirely somnolent, as WNBT in New York City carried on with limited program schedule throughout the war and was used by the New York Police Department and the Office of Civilian Defense to train many thousands of air raid wardens.

With the cessation of hostilities, television again sprang to life, and while new television equipment and receivers were as yet unavailable, existing stations reestablished or increased their program schedules to the limit of available trained man power. During 1946, the Federal Communications Commission, after reviewing the technical standards for television, again approved those standards and urged industry to go full speed ahead. The Federal Communications Commission also made new allocations for television, providing thirteen channels in the ultra-high frequency portion of the spectrum, and these thirteen channels make it possible to assign frequencies to approxi-

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* Decimal Classification: R583.17.

mately four hundred television stations throughout the United States. Thus television broadcasting is assured of adequate channel space to provide spectacular growth within the next several years. During 1946, the Federal Communications Commission also granted construction permits for forty-seven commercial television stations (and one educational television station) scattered through major cities from the East to the West Coast; nineteen more applications are in process with many others anticipated.

During the early part of the year, several existing television stations made the necessary changes in their equipment to enable them to broadcast on the new channels assigned by the FCC. The NBC station in New York, WNBT, installed a new transmitter and erected on the pinnacle of the Empire State Building a new high-gain television transmitting antenna. This equipment has been in operation with an average of twenty program hours per week since May 9, 1946. Many outstanding programs have been transmitted among which was the spectacular Louis-Conn heavyweight championship fight. This particular program was picked up by field equipment, using five cameras at the Yankee Stadium, and transmitted by microwave to Radio City, whence it was distributed by coaxial cable to television stations in Washington, Philadelphia and New York, and by a mountain-top radio relay to the General Electric station in Schenectady. A large audience of government officials seated before projection receivers in the Statler Hotel in Washington also viewed the event. It is estimated that over two hundred thousand persons witnessed this event by television.

February 12, 1946, Lincoln's Birthday Anniversary, marked the opening of the American Telephone and Telegraph Company's coaxial cable for television program transmission from Washington to New York. Television cameras viewed the ceremonies taking place at the Lincoln Memorial and at other spots in the Nation's Capitol. Television audiences in Washington, Philadelphia, New York and Schenectady viewed those ceremonies, while they took place, through the magic of television networking. Thus the practicality of intercity television networking was again demonstrated, emphasizing the importance of intercity television links, and holding forth the promise of nation-wide service in the relatively near future.

The American Telephone and Telegraph Company is now in the process of constructing a transcontinental coaxial cable which will link the East Coast to the West Coast. This cable will probably be ready for television transmission in late 1948, and to the many progressive broadcasters throughout the country, this promise of television network service is a great stimulus. The Bell System is also in the process of constructing a radio-relay system suitable for carrying television programs between New York and Boston, which will be available for such purpose late in the summer of 1947. The local telephone companies have also made available to television a method of transmitting television over telephone wires for short distances to facilitate pickups in urban areas. This service is also augmented by the use of radio relays. It was through the use of such a radio relay that the television audience was able to watch the Army football games as they were played at the Military Academy at West Point.

As part of regular weekly broadcasts, boxing is televised from Madison Square Garden and St. Nicholas Arena in New York City, and through the use of light-weight portable microwave relay equipment, practically any point within twenty miles of Radio City becomes available for television camera coverage; through the use of a double relay, even greater distances can be covered, giving television cameras a greater program flexibility than that which was available before the war. The New York, Schenectady, Philadelphia and Washington television audiences have witnessed many outstanding baseball and football games televised during 1946.

The television frontier has been immensely extended due to the availability in 1946 of new portable pickup equipment, the cameras of which incorporated the wartime-developed highly sensitive image orthicon camera tube which is able to produce pictures of excellent quality at extremely low light levels, exceeding in sensitivity that attributable to the readily available motion picture film. This one great development enables television to present any event, day or night, requiring no more light than that which one encounters in any night spot where the public gathers.

The President of the United States, Harry S. Truman, was televised as he addressed the commencement exercises at Fordham University on May 11, 1946, which episode gives a glimpse into the future of the part which television broadcasting will play in the political affairs of the future. In October, 1946, the Television Broadcasters Association held its Second Conference and Exhibition which attracted over a thousand radio executives and engineers from all parts of the country. All of the main sessions of this conference were televised over a network covering New York, Schenectady, Philadelphia and Washington. At this exhibition, the first of the post-war home television sets of several different manufactures were publicly demonstrated. These sets included table models, console models incorporating standard band broadcasting, FM, and automatic phonographs, and large screen projection receivers. Great interest was displayed in these sets which were just then starting to roll off the assembly lines. It was estimated that some fifteen thousand receivers would be available for public purchase before the end of 1946, and it is further estimated that between two hundred thousand and three hundred thousand sets will have reached the public by the end of 1947; within three years, approximately one million sets should be in the hands of the public.

Television, destined to play a world-wide role in the future, brought to the present four-city television audience many of the sessions of the United Nations as they gathered at their headquarters in Flushing, Long Island. Famous personalities representing member nations of the United Nations appeared daily on the home screens of the television audience.

Again forecasting the important role which television will play in the destinies of national politics, television cameras televised the opening session of the 80th Congress as it convened at the Capitol on January 2, 1947 and January 6, 1947. As President Truman addressed the combined houses of Congress, he was seen by the television audience in the four above-mentioned cities through the eyes of the television cameras.

Nineteen hundred and forty-six has, indeed, marked an important milestone in the growth of television. These important events will fade, however, into insignificance compared to the anticipated growth in 1947. Assembly lines producing home receivers will be in full swing, and factories will be turning out, in quantity, television transmitters, cameras, studio and field equipment, supplying those broadcasters fortunate enough to have received from the Federal Communications Commission their construction permits for television stations.

As always, the research and engineering laboratories of this corporation have pressed their reasearch for new and ever-improved television devices, and there has emerged in 1946 a promise for the future of color television. Several methods of attacking this problem have been demonstrated, but as yet, there remains much work to be done before color television can provide a public service as practical and reliable as the monochrome service which the public now enjoys. The wonders of radio never cease, and television is the greatest marvel to emerge from electronic laboratories. It holds an influence for the good of mankind that no man can entirely comprehend.

TELEVISION TODAY AND ITS PROBLEMS—1946*

BY

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New York, N. Y.

Summary—This paper is a general review of present day television. The outstanding characteristics of television as a public service are discussed. Specific topics covered include studio equipment and operations remote pickups, special studio techniques, programming, network facilities, receivers, and other television development factors.

COMMERCIAL television broadcasting is now an active and growing art in America. Despite its comparative youth, it has attained much of the picture clarity, program flexibility, and entertainment capabilities of motion pictures. It surpasses even that well-developed art in its spontaneity of presentation, its occasional rendering of the unexpected, its instantaneity and immediacy of production and reproduction, and its economical production technique. In due course, it may be expected to take its place side by side with its older sister art.

The following discussion of present-day television and certain of its problems is necessarily a general review. To have listed all details and to have given due credit to the capable engineers and program planners who have contributed to present-day television would have far exceeded the permissible limits of this brief description.

STUDIO EQUIPMENT AND OPERATIONS

As the result of several years of careful planning and testing, studio equipment may be regarded as semi-standardized. It is able to meet all normal operating requirements to a reasonable extent. Most of its future development will be more in the direction of detailed improvement, in all likelihood, than of radical modification (unless, at some future date, basically different methods of television operation should be discovered and found superior to those now available—a somewhat unlikely contingency).

Studio and remote-pickup cameras are now highly sensitive and readily handled. The image orthicon has proven to be a picture tube of unparalleled capabilities. Its sensitivity is a hundredfold greater

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* Decimal Classification: R583.17.

than the original iconoscope and its resolution or detail-recording power has been brought to a satisfactory level. Television cameras may in the future be made even more mobile and free from limitations on, or rapid changes of, their position. The view finders, being electronic, are non-parallactic and truly indicative of the actually transmitted pictures. A full complement of lenses is available which can be used at distances of from tens to hundreds of feet from the subject, and capable of covering anything from a clear wide-angle view to a nearly flawless close-up. Practically any desired "effect", wipe, fade, or the like can now be optically or electronically secured.

At the present time television pictures lack depth of field in some cases. The problem of increasing this depth can be attacked by the use of cooler light sources and more of them, more sensitive camera tubes, or specialized optico-electrical systems of studio illumination inherently capable of delivering an increased range of field.

Studio lighting today is usually adequate in amount, but there is room for future improvement in its dramatic values and in the more widespread and skillful use of modelling lighting of types showing actors and sets to best advantage.

When color television is ultimately commercially achieved, additional lighting problems will arise. The color quality of illumination must be more carefully controlled. Adequate lighting for color pickup and reproduction ranges from eight to fifteen times the amount necessary for black-and-white operation, the exact figure depending in part upon design skill and in part upon the desired fidelity of color reproduction.

Studio control equipment during the last years has become more dependable, accurate, and convenient in manipulation. The task of the director in "editing" the television production in the control room, and as it progresses, has become more effective and less trying. Further, the control of the activities of studio personnel from the director's control desk is now more rapid and complete than it was in the past. There remains for the future the development of methods for simplifying the tasks of the camera men, the microphone-boom operators, and the actors, through coordinated assistance from the control desk.

REMOTE PICKUPS

A new era in the scope and quality of remote pickups was ushered in by the modern highly sensitive television camera. Sports events under normal lighting are now most acceptable elements of the television program. In the days to come, the portability and ease of manipulation of remote equipment will doubtless be even further enhanced.

It is necessary that connections between the cameras, the remote-pickup control equipment, and the receiver which controls the main transmitter shall be compact, light, easy to handle, and yet capable (as they are today) of transmitting the sight and sound signals without noticeable distortion.

SPECIAL STUDIO TECHNIQUES

The optical or electrical insertion of backgrounds has been under study. It appears likely to prove a useful factor in the future in readily enabling more elaborate and yet economic productions. If a length of film or a slide can adequately replace a physical set for background purposes, many studio problems and economic limitations will be worked out in the control room rather than on the studio floor. Picture presentations are increasingly elegant and smooth. Fades, dissolves, and wipes can be optically or electrically contrived. Numerous novel and attractive "effects" have been found feasible—and doubtless many more of these will be evolved in the future.

At present, live orchestras are unavailable because of restrictions imposed by their labor organizations. These restrictions have, however, benefited television in one respect. They have made necessary the development of convenient and economical ways of using orchestral and vocal phonograph records in television, employing an evolved dubbing technique. The home audience may, for example, hear a singer in actuality, accompanied by an orchestral recording of the selection which he is singing. Or the entire sound reproduction in the home may originate on a vocal and instrumental record which is transmitted on the audio channel and which also "cues" the actor in the studio so that, so far as the audience is concerned, he seems to be singing the actual selection in exact time. These methods may be further elaborated in time, even if live orchestras become available, since they have certain inherent technical, operating, and economic advantages.

The use of film transmissions in television is considerable at present. The dependability and quality of such transmissions are now in the main satisfactory. Certain details of operation may be added or improved in the future. For example, it might prove convenient to run certain types of film at any desired speed and yet transmit them at the standard number of frames per second. Experimentation on the remote control of film projectors from the television studio control room will doubtless be studied in the future.

Experimenters have indicated the possibility of recording television programs on 16-millimeter or 35-millimeter film, with the accompanying sound. Such film serves as a permanent record of the program, and provides for its later repetition or for its syndication on a transcription basis. Another application of the recording of a television program on film is in the field of theater presentations. If a television program is received in a theater, photographed on film, developed at high speed, and then projected on the theater screen, it offers one method of large-screen television for theater purposes. The use of television in theaters is a field under development, and one having attractive possibilities.

PROGRAMMING

The program structure of today is relatively simple and is based largely on most economical production for a limited number of operating hours per week. The future still presents numerous program problems. It will become necessary to present a far wider variety of program types, and to learn how to produce thoroughly attractive program material on a large-scale and economical basis. Television on such an expanded scale will require locating and training large groups of skilled personnel capable of meeting television-program needs.

Closely tied into the program problems of the future are certain economic questions which must be studied. It remains to determine the most desirable number of stations in each type and size of community, the sort of programs which should be transmitted, and the measure of sponsor support which will be necessary to provide adequate program service in various sorts of communities. To determine these factors may involve the detailed and continued study of audience reactions to an extent not yet found necessary in radio broadcasting.

NETWORK FACILITIES

As of the end of 1946, over three thousand miles of ccaxial cable and many hundreds of miles of radio-relay systems have been established. By 1950, it is stated that 12,000 miles of coaxial cable capable of carrying television programs will form a national network. No rate structure has as yet been established for the use of such facilities. It will be a complex task, though a promising one, to provide nation-wide high-fidelity black-and-white television syndication on an economic basis. The prospects are hopeful in view of the many conceivable and profitable by-product uses of the wide-channel circuits required for television syndication.

To secure maximum flexibility and greatest usefulness of networks for television program distribution, it is desirable that the engineers develop standards and simplifications in the use of cables, wave guides, and radio-relay systems. Interchangeable operation of standard equipment will thus be promoted. It may be added that, when color television becomes commercial, network operations will present technical and economic problems of increased difficulty.

One television problem, shared with other higher-frequency services, and as yet almost untouched, is that of providing large-area rural television coverage in non-mountainous regions, and controlling the corresponding transmissions from national networks. In view of all that television can offer the farmer, it is greatly to be hoped that a solution will be found for this problem. To some extent such a solution will be "tied in" with detailed studies of wave propagation from stations of various carrier frequencies, powers, and heights above ground

RECEIVERS

Present-day television receivers operate on signals transmitted on thirteen black-and-white channels, each six megacycles wide, and located in groups between 44 and 216 megacycles. Some of the receivers show pictures by direct viewing and others by projection of an intensely brilliant image through a high-efficiency optical system. Other optical methods of enlarging pictures efficiently have been considered. The present-day picture is a 525-line (nominal) image with slightly greater resolution vertically than horizontally. Its brightness is of the order of 20-60 foot-lamberts, which is a marked improvement on earlier pictures and adequate for home reception under normal conditions. The gradation range of the picture has been greatly increased.

Television receivers of today are compact, attractive in appearance, and easily handled by most persons. They may fairly be said to be far in advance, in their stage of technical and operating development, of the radio receivers of the mid-1920's during the period of the vast expansion of radio broadcasting.

Home antennas are mostly of the dipole type, using one or more of such elements with or without passive reflectors. The problem of providing television service in multiple-apartment dwellings has been studied in principle. It involves centralized antenna, amplifier, and distribution systems leading to the various apartments. Technically, such systems seem practicable. Their economics—as well as the landlord reaction to their installation—remain to be explored.

Television still offers a number of questions to the analyst. It is not known what number of available channels would meet all reasonable requirements in each type and size of community. The relative status and feasibility of operation on 44-216 megacycles versus operation on 480-920 megacycles (or even operation on frequencies well above 3000 or even 5000 megacycles) will require more technical work and field experimentation than has yet been devoted to this important question. Then, too, as picture size increases in the home, it may be found that 525-line pictures are still entirely adequate—or, alternatively, it may prove desirable to go to pictures of greater resolution. This last question, however, is far from urgent since it is generally agreed that the 525-line picture is capable of giving continuing entertainment value to the television audience.

TELEVISION-DEVELOPMENT FACTORS

The progress of television will involve some further developments or expansions whereby the television audience will be satisfied at all times that television progress continues.

One element of television which cannot be slighted is high-quality servicing. The installation and maintenance of television receivers can be handled only by highly qualified service men. Such men are being supplied by the television-receiver manufacturers or by organizations sponsored by such manufacturers. This is a wise and helpful procedure.

Satisfaction in television reception depends upon the absence of visible interference with the picture and audible interference with the sound portion of the program. It is well known that certain medical and industrial equipment, such as diathermy apparatus, industrial-heating devices, and some automobile ignition systems may seriously affect picture quality and cause noticeable interference with the sound portion of the program. A combination of regulatory measures and public education may lead to the reduction or elimination of such disturbances.

Those who watched the early development of radio broadcasting will recall the serious interference which resulted from radiating receivers. This possibility has already been found to exist in television. The elimination of such interference will require agreement on engineering standards of good practice for "non-radiating" receivers, and possibly the promulgation of corresponding regulations as to permissible field strength at specific distances from suitably designed television receivers.

The cooperation of all major factors in the television industry is important if the art is to grow normally. The Radio Manufacturers Association provides such cooperation between manufacturing groups, and the Radio Technical Planning Board, through its various Panels, offers an even wider forum to all interested and significant organizations and individuals.

Perhaps the least studied portion of the basis of television today is its psycho-physical background. The effect of various types of image presentation upon the observer merits, and will doubtless receive, careful investigation and analysis in the future. The improvements which would result in the entertainment value of television would probably repay extensive investigations along psycho-physical lines.

As of the end of 1946, color television is under vigorous discussion. Two major types of color television: simultaneous and sequential, have been demonstrated. The considerations favoring each of these have been forcibly urged by their respective proponents. Color television is a more complicated matter than most enthusiasts in that field appreciate at this time. Our knowledge of color reproduction methods, and of their desirability in television, is certainly in its early stages in some respects. Also under active analysis are questions of the economics of color television involving transmitter cost, studio and program operation costs, and receiver costs. It is evident, however, that a healthy development trend exists in color television and that, in the years to come, this art may also find its sphere of application in home entertainment.

Some workers have, from time to time, discussed stereoscopic television, in black-and-white or color. The subject is of limited interest at present but may attract increasing attention in time.

It may also be mentioned that important industrial television applications, as well as uses of television in military and other fields, will increase in number and grow in value.

* * * * *

As of the close of 1946, it has been fundamentally established that television broadcasting is an outstanding contribution to mass communication and that it rests on basically correct technical, esthetic, and long-term commercial grounds. Among its remaining problems are further simplification of equipment and operations, achievement of large-scale and high-quality production and reproduction, expansion of the scope and coverage of television programs, development of coordination between television and other methods of entertainment and instruction, the maintenance of a constructive and helpful attitude on the part of the governmental regulatory authorities, the growth of intra-industry cooperation, and the gradual evolution of television toward even greater technical and artistic capabilities and achievements. Considering its present capabilities and accomplishments, television presents most attractive and hopeful vistas of future achievements and universal acceptance.

SUMMARIES

The following papers are presented in summary form only. The journal in which the full paper appears is indicated in each case.

MEASUREMENT OF THE SLOPE AND DURATION OF TELEVISION SYNCHRONIZING IMPULSES*†

By

R. A. Monfort# and F. J. Somers‡

Summary

Satisfactory operation of television receivers in the field requires that the waveform of the transmitted synchronizing signals be held to narrow tolerances. It is therefore essential that suitable measuring equipment and techniques be available at the transmitter so that synchronizing waveshapes can be accurately and rapidly checked. This paper describes several measurement methods which have been found to be satisfactory under practical operating conditions.

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* Decimal Classification: R583.13 × R200.

† *RCA REVIEW*, January, 1942.

Formerly with the Engineering Department, National Broadcasting, Company, Inc., New York, N. Y.

‡ Engineering Department, National Broadcasting Company, Inc., New York, N. Y.

THE RELATIVE SENSITIVITIES OF TELEVISION PICKUP TUBES, PHOTOGRAPHIC FILM, AND THE HUMAN EYE*†

By

A. Rose‡

Summary

The threshold scene brightness which a picture-reproducing device can record, a measure of its "operating sensitivity", depends not only upon the lens speed and the exposure time, but also upon the amount of detail in the recorded image. A general expression for the "operating sensitivity" of a picture-reproducing device is obtained which includes these factors together with the threshold number of quanta per picture element. This parameter *characterizes the "true sensitivity" of the given device. The "true" and "operating" sensitivities of four types of television pickup tubes, photographic film, the human eye, and an ideal picture-reproducing device are obtained. Eye and film have of the order of one one-hundredth the "true sensitivity" of an ideal picture-reproducing device. Some recent television pickup tubes have of the order of one one-hundred-thousandth the "true sensitivity" of an ideal device.*

To compare "operating sensitivities", the same exposure time and equivalent lens systems are taken for the three devices. A television pickup tube which has a photoelectric response of 10 microamperes per lumen and makes full use f the storage principle can record scenes with no more illumination than that required by some of the "faster" photographic films. The relatively low "operating sensitivity" of film results from the large amount of intrinsic picture detail (a picture element is taken to be a single grain). The human eye has a range of "operating sensitivities" extending from that of film to a value several thousand times higher. This range depends upon the ability of the eye to coarsen the detail of its perceived image as the scene brightness is lowered.

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* Decimal Classification: R583.1.

† *Proc. I. R. E.*, June, 1942.

‡ Research Department, RCA Laboratories Division, Princeton, N. J.

A PORTABLE HIGH-FREQUENCY SQUARE-WAVE OSCILLOGRAPH FOR TELEVISION*†

BY

R. D. KELL,# A. V. BEDFORD# AND H. N. KOZANOWSKI‡

Summary

A portable high-frequency oscillograph for television is described by which a square-wave (100-kilocycle) response may be viewed as a dotted wave and readily recorded as a series of readings. The dots are spaced at 1/30 — (or 1/20 —) microsecond intervals. No electrical connection is required between the oscillograph and the square-wave generator other than that established through the apparatus under test since the synchronous sweep and timing dots are derived from the square-wave response of the apparatus. Circuit diagrams of the square-wave generator and square-wave oscillograph are given.

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* Decimal Classification: R371.5 × R583.

† *Proc. I. R. E.*, October, 1942.

\# Research Department, RCA Laboratories Division, Princeton, N. J.

‡ Engineering Products Department, RCA Victor Division, Camden, N. J.

CATHODE-RAY CONTROL OF TELEVISION LIGHT VALVES*†

By

J. S. Donal, Jr.‡

Summary

When a light valve is employed for the reproduction of television pictures, it is desirable to make use of a cathode-ray beam to control the light valve in order to preserve the all-electronic character of the television system. A number of procedures of cathode-ray control are described, the majority of which are applicable particularly to the control of the suspension light valve.

The general method employed is shown to be the production of an electric field through the light valve by bombarding one side of the valve with electrons of very high velocity, causing the valve areas to be charged in a negative direction toward the limiting potential of the bombarded surface. Removal of the electric field is then accomplished by charging these areas back toward their original potential by the use of electrons of substantially reduced velocity.

The most elementary procedure described is one in which a single beam of electrons of constant velocity is employed, discharge being accomplished by secondary electrons generated by the action of the beam of primary electrons.

The effects of polarization of the light valve, resulting from the comparatively low resistivity of the suspension, are described and explained. It is shown that a suspension of such low resistivity as to be uncontrollable by the other procedures may be made operative when the valve is used in combination with a spatially modulated electron spray and when, in addition, the potential of one wall of the valve is increased and decreased at a moderate frequency.

Of the procedures described, the most effective from the practical standpoint is shown to be one in which the light-valve field is developed by a scanning beam, and in which the field is later removed by rescanning with the same beam at a reduced electron velocity. A photograph is shown of a picture reproduced by the light valve when controlled by this method.

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* Decimal Classification: R583.15.

† *Proc. I. R. E.*, May, 1943.

‡ Research Department, RCA Laboratories Division, Princeton, N. J.

A REFLECTIVE OPTICAL SYSTEM FOR TELEVISION*†

By

E. W. Wilby‡

Summary

The most logical reason for the use of a reflective optical system for projection television is that it is the only known simple means capable of

focusing a large field with high efficiency. The maximum efficiency obtainable from conventional projection lenses lies between 6 and 12 percent, while it is quite feasible to obtain between 20 and 40 percent from reflective optical systems. The term efficiency, as used here, refers to the fraction of the total light emitted by a central point on the tube face that is focused on the image point.

It has long been known that there are very efficient means for focusing very small sources, for example, parabolic reflectors for infinite throw and ellipsoidal mirrors for finite throw. A considerable number of good reflective systems capable of handling large fields have been proposed and many of them have been used. All of these however contain one or more aspherical elements, and were designed for infinite focus for astronomical use. The simplest of these was proposed by B. Schmidt and consists of a spherical mirror and a very weak aspherical correcting lens.

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* Decimal Classification: R583.15 × R138.3.

† RCA Licensee Bulletin LB-630, November, 1944.

‡ Industry Service Laboratory, RCA Laboratories Division, Chicago, Ill.

PROJECTION TELEVISION*†

BY

D. W. EPSTEIN# AND I. G. MALOFF‡

Summary

Projection television, which is simply the projection onto a viewing screen of the picture originating on a cathode-ray tube seems, at present to be the most practical means of producing large television pictures.

The 2 basic problems of projection television are: (1) the problem of providing a cathode-ray tube capable of producing very bright pictures with the necessary resolution and (2) the problem of providing the most efficient optical system so as to utilize the largest possible percentage of the light generated. These problems were very vigorously attacked over a period of years and the progress made toward their solution has been very satisfactory.

Problem (1) has been solved largely by the development of cathode-ray tubes capable of operating at high voltages. Problem (2) has been solved by the development of a reflective optical system about 6 to 7 times more efficient than a good f/2 refractive lens. The reflective optical system consists of a spherical front face mirror and an aspherical correcting lens.

A handicap of this optical system, for use in a home projection receiver, was the high cost of the aspherical lens. This has been overcome by the development of machines for making aspherical molds and by the development of a process for molding aspherical lenses from plastics. RCA reflective optical systems are designed for projection at a fixed throw and require cathode-ray tubes with face curvatures fixed in relation to the curvature of the mirrors in the system. A number of such systems, suitable for project-

ing television pictures with diagonals ranging from 25 in. to 25 ft., have been developed.

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* Decimal Classification: R583.

† *Jour. Soc. Mot. Pic. Eng.*, June, 1945.

Research Department, RCA Laboratories Division, Princeton, N. J.

‡ Home Instrument Department, RCA Victor Division, Camden, N. J.

BAND-PASS BRIDGED-T NETWORK FOR TELEVISION INTERMEDIATE-FREQUENCY AMPLIFIERS*†

BY

G. C. SZIKLAI‡ AND A. C. SCHROEDER‡

Summary

Bridged-T networks offer great economy in television intermediate-frequency amplifiers for sharp attenuation of the associated and adjacent sound channels.

A simple design method was obtained by the use of the equivalent lattice. By the same method, general formulas were obtained for the phase, attenuation, and delay characteristics. Two designs are given to illustrate the convenience of the method.

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* Decimal Classification: R583.5.

† *Proc. I. R. E.*, October, 1945.

‡ Research Department, RCA Laboratories Division, Princeton, N. J.

INPUT IMPEDANCE OF SEVERAL RECEIVING-TYPE PENTODES AT FM AND TELEVISION FREQUENCIES*†

BY

F. MURAL‡

Summary

The input impedance of vacuum tubes is an important circuit design consideration at high frequencies. This report includes information on the input impedance of a number of currently available r-f pentodes.

Measurements are given on the variation of input resistance with frequency, and on the variation of input resistance and capacitance with plate *current. Measurements are also given for the compensation of input resistance and capacitance variation with plate current by means of unby-passed cathode resistance.*

The frequency range of measurement was chosen roughly to cover the frequency modulation and television transmission assignments as well as the recommended intermediate frequencies of receivers for these services.

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* Decimal Classification: R583.6.

† RCA Licensee Bulletin LB-661, March, 1946.

‡ Industry Service Laboratory, RCA Laboratories, New York, N. Y.

TELEVISION HIGH-VOLTAGE R-F SUPPLIES*†

BY

R. S. MAUTNER# AND O. H. SCHADE‡

Summary

Because of the contemplated large scale production of television receivers, considerable interest is being shown in various methods for obtaining a source of high voltage simply and economically.

The r-f oscillator type high voltage supply described in this bulletin offers advantages in economy and space requirements, and provides satisfactory regulation over the normal operating range of cathode-ray tube beam currents. Furthermore, its safety factor against dangerous shocks is considerably greater than that obtained with conventional sixty-cycle supplies. The principles of operation and design of these supplies have been previously described (Schade, O. H., "Radio-Frequency Operated High-Voltage Supplies for Cathode-Ray Tubes", Proc. I.R.E., pp. 158-163, April 1943). These are here reviewed and considered in greater detail. Constructional features of four typical units are shown and their performance is illustrated by curves indicating the magnitudes of current and voltage obtained under typical operating conditions. Sample calculations for the specific cases of a 75-watt 90-kilovolt supply and a 10-watt 30-kilovolt supply are included to illustrate the progressive steps in designing and calculating the circuit elements and operating conditions for a specified performance. During the time that these supplies have been in operation they have given stable trouble-free performance and have required a minimum of attention. However, careful shielding and filtering are required to minimize undesired radiation. Corona problems especially require careful consideration to prevent ionization during periods of high humidity.

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* Decimal Classification: R583.5 × R366.

† RCA Licensee Bulletin LB-675. August, 1946.

\# Industry Service Laboratory, RCA Laboratories Division, New York, N. Y.

‡ Tube Department, RCA Victor Division, Harrison, N. J.

APPENDIX

TELEVISION

A Bibliography of Technical Papers

by RCA Authors

1929 — 1946

This listing includes some 275 technical papers on TELEVISION and closely related subjects, selected from those written by RCA Authors and published during the period 1929-1946.

Papers are listed chronologically except in cases of multiple publication. Papers which have appeared in more than one journal are listed once, with additional publication data appended.

Abbreviations used in listing the various journals are given on the following page.

ABBREVIATIONS

An. Amer. Acad. Polit. Soc. Sci.	ANNALS OF THE AMERICAN ACADEMY OF POLITICAL AND SOCIAL SCIENCES
Broadcast News	BROADCAST NEWS
Broad. Eng. Jour.	BROADCAST ENGINEERS JOURNAL
Communications	COMMUNICATIONS
Elec. Eng.	ELECTRICAL ENGINEERING (TRANSACTION A.I.E.E.)
Electronics	ELECTRONICS
Electronic Ind.	ELECTRONIC INDUSTRIES
FM and Tele.	FM AND TELEVISION
FM Business	FM BUSINESS
Inter. Project	INTERNATIONAL PROJECTIONIST
Jour. Appl. Phys.	JOURNAL OF APPLIED PHYSICS
Jour. Frank. Inst.	JOURNAL OF THE FRANKLIN INSTITUTE
Jour. Opt. Soc. Amer.	JOURNAL OF THE OPTICAL SOCIETY OF AMERICA
Jour. Soc. Mot. Pic. Eng.	JOURNAL OF THE SOCIETY OF MOTION PICTURE ENGINEERS
Jour. Tele. Soc.	JOURNAL OF THE TELEVISION SOCIETY
Phys. Rev.	PHYSICAL REVIEW
Proc. I.R.E.	PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS
QST	QST (A.R.R.L.)
Radio and Tele.	RADIO AND TELEVISION
Radio Craft	RADIO CRAFT
Radio Eng.	RADIO ENGINEERING
Radio News	RADIO NEWS
Radio Tech. Digest	RADIO TECHNICAL DIGEST
RCA Rad. Serv. News	RCA RADIO SERVICE NEWS
RCA REVIEW	RCA REVIEW
RMA Eng.	RMA ENGINEER
Short Wave and Tele.	SHORT WAVE AND TELEVISION
TBA Annual	ANNUAL OF THE TELEVISION BROADCASTERS ASSOCIATION
Tele. News	TELEVISION NEWS
Televiser	TELEVISER
Television	TELEVISION

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