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VISION OF FUTURE DEVELOPMENTS

By

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President, Radio Corporation of America

Excerpted from a recent book by Gleason L. Archer, President, Suffolk University*

IT IS impossible to forecast with any degree of precision the events which will follow even during the next few years, to say nothing of the developments which will materialize during the coming decade or so. However, the signs of the times indicate that we are moving rapidly towards perfecting and adopting instantaneous methods of communication which should enable transmission of intelligence of unlimited volume, and over any desired distance, at much higher speeds than have heretofore been possible. These new agencies of communication will serve the eye as well as the ear and thus add enormously to the individual's power to interpret and comprehend the intelligence transmitted.

The fields in which I believe we may reasonably expect to see important future developments are, (a) Telephony, (b) Telegraphy, (c) Radio and Television, (d) The Entertainment Field and (e) Application of the Discoveries of Radio Scientists to Other Fields.

(a) TELEPHONY

The telephone system of communication as it exists today has been based upon the use of a number of conductors which carry electrical impulses. Great ingenuity has been exercised in making it possible to transmit over a minimum number of such conductors, several telephone conversations, telegraph messages, printer telegraph services and facsimile service. Demands of the future will require new methods for such services. The greater future of telephony lies, I believe, in developments which will make it possible to carry all these and many other services over or through a new kind of cable called coaxial cable. This consists of a hollow tube or pipe of relatively small dimensions, with a single conductor along its central axis.

* *Big Business and Radio*, American Historical Company, Inc., New York, 1939.

Much has already been accomplished by scientists to develop such a cable, but the tremendous possibilities which lie in this and kindred fields are now becoming apparent. That scientists will continue to perfect cables of this kind I have no doubt. Think of the amazing prospect of being able to send over a single cable at the same instant of time, telephone messages, telegraph messages, facsimile service, printer telegraph service and most remarkable of all, television service! To be sure, such a cable is expensive—at present very expensive—yet if we can feed into it all of these types of communications and then provide suitable devices for the recapture at a distance of each, then we may forget its cost. Probably more than a hundred such channels are even now possible, and undoubtedly this number will be greatly increased.

I look forward with confidence to the time when the cost of long distance telephony will be so greatly reduced that a talk between New York and San Francisco will be no more expensive than between New York and Boston. It is even conceivable that the time may come when people will pick up a telephone and speak with a person in any part of the United States for a nickel or a dime.

I have spoken of this new type of cable as a land-line system. While it is possible that ocean cables also may utilize this same medium, yet in this field the cost may be almost prohibitive. This is due to the fact that in ocean space there are no customers—and only by large customer demand may the cost of such a cable system be warranted.

In international communications I believe that radio has great advantages over any kind of cables now on the technical horizon, since radio furnishes direct telephone or telegraph communication with any desired nation. A cable terminates at the shore-end and this requires additional land wires or cable circuits to complete the task. Radio is now used exclusively for trans-oceanic telephone communications.

(b) Telegraphy

Obviously such a development of telephone communications as is predicted above will require the telegraph systems to keep pace or else they will lose out technically, which means finally losing out financially. It would seem that a final chapter is also being written to the Morse Code. While there is always a residue from yesterday that remains useful today and even tomorrow, yet the residue of usefulness of the Morse Code is rapidly diminishing. The world is moving on and speed is its watchword. The ingenuity of man has devised a system of instantaneous communication known as facsimile that so far outmodes the electric telegraph that the latter must eventually take its niche in the corridors of history.

Let me give you a crude illustration of the difference between the electric telegraph system and this latest miracle—facsimile. Suppose one wishes to transport a crate of freight from New York to San Francisco. What would you think of a process which would do the job in the following manner? It first broke up the freight into small fragments, then enclosed it in the crate, then hauled it to the train, then carried it across the country, at which point the crate would be taken from the train, the fragments reassembled and the product put together for delivery to the addressee in the form it originally left the factory. One would think that a pretty cumbersome method! Well, that is the present method of telegraphy by the Morse Code. A man writes a telegram and the messenger boy calls for it. It is taken to the telegraph office. There the paragraphs are broken down into sentences; the sentences into words; the words into letters; the letters into dots and dashes. In the form of dots and dashes it is transmitted by wire from New York to San Francisco. At that point the receiving operator reverses the process. He puts the dots and dashes into letters; the letters into words! the words into sentences; the sentences into paragraphs and then a messenger boy delivers the message to the addressee.

However much of sentiment we may have for the Morse Code or the Continental Code—and I confess to that sentiment myself, having been a wireless operator in my youth—we cannot permit sentiment to delay human progress. Obviously the ideal system of communication would be to take the message in its original form, hold it up to something or other, throw a lever, or press a button, and have an exact facsimile of it made instantaneously at the other end of the circuit. To be sure, facsimile has not yet reached this stage of development, but when it comes—and it will come—we will have a new method of telegraph communication.

Facsimile has already been in use over the Atlantic Ocean by radio for a number of years. The pictures in the newspapers one sees today coming from abroad are carried by radio. Pictures are also carried within the United States over telegraph wires by the telephone and telegraph companies. But the methods employed are still subject to much further improvement. The instantaneous system that I refer to will not only replace the dot and dash system of Morse telegraphy but it will also make the present teletype system which still deals piecemeal with letters, words and sentences, a less important method in the future. Here again we shall see human progress on the march. Every step of that march is necessary yet steps become but 'footprints on the sands of time'.

Radio will play an important part in this march of human progress. The discoveries of radio scientists and engineers are now beginning to produce methods and systems which will utilize wave-lengths measured in feet and in inches rather than in hundreds, or thousands of feet. These very short waves will be used in radio circuits for interconnecting cities and for short distance communications. This will be accomplished by a system of automatic or unattended relay stations spaced along the route and so arranged that impulses introduced at one point of the radio circuit will be transmitted automatically through each relay point intermediate to its destination. These new wave-lengths will make it possible to use relay stations of small power to carry simultaneously many channels of communication. Such radio relay circuits will open the way for widespread distribution of television, for flashing facsimile messages between cities, and for many other services of communication.

When finally developed, the new system of facsimile communication should be cheaper than present-day methods. It should be able effectively to compete with voice communication and perhaps offer competition to the mail-bag. It is conceivable that a letter from New York to San Francisco can one day be carried either over coaxial cables, or through the air, at no greater cost than is now involved in sending it by air mail. And it will surely get there more quickly.

A newspaper in the home, by radio facsimile, has already been hinted at, and this too, will be advanced through the further development of new communication channels in the air.

(c) Radio and Television

It must be borne in mind that the principal difference between radio and other means of communication is in the medium. The telegraph and the telephone use wires or cables while radio uses the air, but there is much similarity in the terminal equipment used by each. What has been said about the possible new applications to wires and cables equally applies to radio. Here, too, a last chapter is now being written and a new one beginning.

The reason for this statement is that the new types of services we are discussing take up room in the air, and present-day wave lengths are insufficient to provide the necessary room for the new services. Despite the phrase 'As wide as all outdoors', there is already traffic congestion in the ether. What the scientist must do is not only to learn how to increase traffic upon present lanes but also to carve out new lanes in the ether. This he began to do with what were once called 'short waves' of 100 meters or so. Today, they are already regarded as 'long waves'. The scientist of the air did not stop with 100 meters. He has been exploring the air and has found useful lanes in the wave lengths substantially below 100 meters. Thus present-day international services by radio are conducted on wave lengths varying between 14 and 50 meters.

The new service of television is being commenced on a wavelength of about 6.5 meters. Television requires transmitting channels about 600 times wider than those used in existing sound broadcasting services. For this reason, ultra short waves are used as they provide space for a television service which otherwise could not be accomplished.

The scientists of the world are now only beginning to probe the mysteries of that vast new region of the radio frequency spectrum called ultra short waves. These are waves from about 10 meters in length to waves of less than 1 centimeter (1/100 meter) in length. That portion of this new spectrum between 10 meters and 1 meter provides 9 times as much room as is now occupied by all existing radio services regardless of the wavelengths they use; between 1 meter and 1 decimeter (1/10 meter), 90 times as much space is provided and between 1 decimeter and 1 centimeter this space is 900 times greater. Thus between wave lengths of 10 meters and 1 centimeter we find nearly 1,000 times more space in the frequency spectrum than is now occupied by all of the radio services in existence today. The scientists of the world are merely at the threshold of new possibilities in radio. They are just beginning to learn how to use effectively the many wavelengths of this vast domain of the spectrum.

Ultra short waves open entirely new fields for radio communication. They can provide services especially useful to police and fire departments for the protection of life and property. They will extend the uses of radio in aircraft communication and navigation; in the prevention of collisions in the air and sea transport when obstructions are not visible; in guiding aircraft to land safely, and so on.

In mobile communication, the use of ultra short waves should make it possible for vehicles on land, in the air and on the water, to establish and maintain contacts with anyone, anywhere. Automobiles may be equipped with simple transmitting, as well as receiving gear, enabling the occupants to talk, as well as to listen, while the cars are in motion. It is also possible to extend the uses of radio so that individuals may carry 'pocket radios' and be able to receive messages and programs wherever they may be. Portable radio transmitting devices of small size, using very little power and operating on centimeter or millimeter waves, may enable individuals to establish instant communication with the main wire or radio systems of the country and thus to obtain a 'connection' with any person in any part of the world. All these are in addition to the services of commercial telegraphy, telephony, high speed facsimile, broadcasting and television, already discussed.

(d) Entertainment

Let us turn for a moment to what has been accomplished in the amusement field. The addition of sight to sound was certainly the beginning of a new chapter of progress. It may reasonably be expected that when television has reached a substantially commercial stage, radio listeners will not be satisfied unless they can also be viewers. Viewers will not be satisfied unless they can also be listeners. This is the case with talking movies at the present time. People once were satisfied to watch a silent actor on the screen, but when the laboratory gave him an electric tongue he no longer was interesting only to look at. The audience also wanted to listen to what the actor had to say. The same may be expected ultimately with sight and sound by radio.

In discussing the field of entertainment and the effects wrought upon that field through the creation of these new instrumentalities of mass communication, it is interesting to observe that here we find the problem to be 'production', whereas in practically all other fields of industry the problem is 'consumption'. Let a really good picture be made in a studio and distributed to the theatres of the country. People in great numbers come to the theatres to see that picture. The cry is 'Let us have more such pictures'. A company may manufacture the best radio set, or the best electric refrigerator, or the best electric vacuum cleaner, but it still has to sell the idea and induce the public to purchase the article. Creative advertising must blaze the trail. Partial payments must be permitted. Ticklers for the trade must be resorted to. Not so with real art! From the humblest to the richest they are willing to deposit their 25¢ or $1.00 at the cashier's window at any theatre that runs a real picture or gives a real show.

What has happened in the entertainment field is that the agencies of distribution have outstripped the agencies of production. It may take six months to a year and a million to two millions of dollars to produce a first-class moving picture, yet it can be shown in the leading theatres of the United States in less than one week. If given on a television network, when such becomes available, it can be shown in 30,000,000 homes in the United States within one hour. 'What comes next?' will be the cry. These agencies of distribution through the new wire mediums and through the air can distribute art instantaneously. The whole nation can consume art instantaneously. But art cannot be created or produced instantaneously. There is the bottle neck of the future. To say this, however, is no to infer that the problem is unsolvable. Indeed it is solvable; but new solutions will have to be created.

Here opportunity beckons to the young and rising generations. It will require all the creative brains the world can supply to feed the growing demand of the masses, as well as the classes, for more and better art. To meet this need it will require the same degree of imagination and invention on the part of the creative writer, artist and performer that is required from the scientist and engineer in the laboratory. To recognize the need for such material may in itself prove to be the first element of invention. And so the theatre of the future may not only show motion pictures as now, but it may be fed from a central point with the greatest variety show that man has thus far produced. Hundreds or thousands of theatres can make that show available simultaneously to audiences.

Methods may also be found whereby the show given to the theatre by television will be received only in theatres. The television intended for the home may be a different show, transmitted and received with different kinds of apparatus and on different wavelengths. Not only may the method of transmission and reception be different but the type of show may also be different. The person in the home may be part of the audience looking in upon the city in which he lives; upon the nation, and perhaps upon the world. People always wish to gratify their primal instinct to see what is going on—to listen and to behold. To be able to sit in one's home, at ease, and by means of television to see and to hear events of the world at the very moment those events are taking place will gratify a human desire.

On the other hand there is the gregarious instinct of humanity to consider—the desire to mingle with the crowd. This, too, is a primal instinct. Where the crowd goes, where ladies wear pretty dresses so that others may see them and use perfume to enhance their charms with the opposite sex—there men also will go and the ancient cycle of life will continue. New amusements will take the place of the old but the concert, the stage and the screen will continue to satisfy that gregarious instinct of humanity. In all these things the wonders of science that I have described and a thousand more yet undreamed of will minister to the life of society in which we live and breathe and have our being.

(e) Applications of the Discoveries of Radio Scientists to Other Fields

As this new chapter in the future development of radio unfolds, it will reveal abundance of contributions to other arts, to sciences and to industries.

The principles of electron optics, for example, which find application in television, will mark advances of great scientific significance in other fields. When applied in electronic microscopy, they will make it possible to develop magnification from 10 to 20 times greater than can now be obtained by any known method. The electron microscope will then become an instrument that will permit observation and study of the most minute cell structures in a detail heretofore impossible. It will provide new eyes to search and to behold the unseen. Through this medium it promises to add greatly to our store of knowledge of biology.

Electron microscopy will make important contributions also in many other fields. An increase of 10 to 20 times in magnification when applied to the science of metallurgy will open up new avenues for the study of alloys. It will greatly increase our knowledge in the field of crystallography. It will facilitate obtaining solutions to problems of industry in the production of both organic and inorganic fibrous material and in other fields of industrial chemistry.

Think also what it will mean in the science of astronomy! Electron optics may go far in reducing the barriers between man and his knowledge of the planets imposed by limitations of even the most highly perfected optical methods of today. Through electron optics and other instrumentalities applied in the art of communication, the planets of the heavens may be brought thousands of light years nearer to the earth for observation and study. Thus may be revealed to mankind in greater measure the secrets of the universe.

In the field of medicine, the discoveries of radio scientists are rich in promise of instrumentalities for facilitating diagnosis, and for combating and curing diseases.

When applied to geophysics, these discoveries will aid man in charting underground geological formations. Mines and oil deposits hitherto unknown will yield to this new X-ray of science. Thus man may find and draw from the earth those things which he seeks in the advancement of our social and economic well being.

The devices originated by radio scientists and engineers should find widespread application in many other industries. Photoelectric cells, vacuum tubes and electronic devices of all kinds will be utilized for operating machinery; for calipering; for color analysis; for control and timing devices; for grading and sorting and for numerous other purposes. Radio frequency current may be used in preserving food by destroying microscopic agencies of food decay. It will free grains from noxious germs and substances. It will make war on insects that prey upon plant life.

One of the significant developments in the industrial application of radio frequency current is seen in the metal producing and refining industry. The older processes were largely guesswork, dependent upon human judgment and uncertain methods in the control of temperature and harmful contaminations. Now, however, the exact methods of control of these essentials provided by radio frequency currents in the electric furnace make possible the production of new alloys and metals of a new order of purity which will enable new industries to be founded.

These new devices will contribute also to the perfection of existing products, and will serve as tools for scientists and engineers to extend their explorations and resolve the unknown of today into the known of tomorrow.

Science recognizes no impassable barriers in its quest for truth. The radio scientist and the technical specialist in the art forever press on to new horizons. They will not limit their researches to the atmosphere in which we live. They will explore the heavens above and the depths of the earth beneath. The limits of human imagination will be the only possible circumscribing frontier of scientific research.

But we cannot place upon the scientist alone the responsibility of translating the products of the human imagination into tangible products and services that are of every-day usefulness to mankind. The scientist is only the first link in the chain. Statesmen and administrators in government are guiding forces; as are also leaders in such fields as education and labor. And the scientist must be supported by financial resources, by the imagination and courage of management, and by man-power intelligently directed—in a word, by industry.

The inventor, as we must appreciate, is a great originator of change in our complex civilization. But great inventions must prove themselves—must run the gauntlet, so to speak, before they emerge as real forces in society. There is natural resistance to change until there is definite evidence of a need for the new idea.

The statesman in government is the guardian of social progress. Research and invention furnish the raw materials of progress, while industry translates these materials into social and economic benefits. As guardian of our social progress, the statesman in government is vested with the power and duty to regulate, but he can also stimulate progress. He can accomplish this by encouraging invention to develop the arts and sciences and by providing for industry freedom of opportunity for effectively translating inventions into social and economic benefits.

Progress is the life-blood of industry. Nevertheless, those who risk their financial resources must be convinced of the soundness of the invention which promotes change. They must envision the nature and extent of the change and the preparedness of the public to accept the innovation; and be given opportunity for reward reasonably commensurate with the risks involved.

To emerge as constructive social forces, great inventions require imagination, skill, courage, faith and finance. Science, industry and government, in cooperation, can follow the road that leads to real progress and provides benefits for all society.

FLUCTUATIONS IN SPACE-CHARGE-LIMITED CURRENTS AT MODERATELY HIGH FREQUENCIES

By

B. J. Thompson, D. O. North and W. A. Harris

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PART I — GENERAL SURVEY

By B. J. Thompson

Summary—*This paper is presented in five parts of which this is the first, serving as an introduction to the others which present detailed analyses of fluctuations in space-charge-limited currents in diodes and negative-grid triodes, in multi-collector tubes, and in tubes with appreciable collision ionization, and of the fluctuations in vacuum-tube amplifiers under operating conditions. The purpose of this part is to present the theoretical and historical backgrounds for the other parts.*

Previous work led to a recognition of the fact that fluctuations in space-charge-limited currents are less than in temperature-limited currents. Attempts were made to explain the actual magnitude of such residual fluctuations in terms of thermal agitations in the internal resistance of the tube at cathode temperature. Agreement between such explanation and experimental results was not observed, the observed mean-square value of fluctuations being only about six-tenths of the predicted value.

It is clear that the thermodynamic reasoning on which the thermal agitations argument is based cannot properly be applied to a case such as a vacuum tube in which the drift velocities of the electrons greatly exceed the thermal velocities. A microscopic examination of the mechanism of space-charge reduction of shot effect is required, therefore.

The number of emitted electrons having velocities greater than any fixed value fluctuates at random. If the retarding potential (virtual cathode) established near the cathode by the space charge were fixed in magnitude, the current passing to the anode would show fluctuations identical with those observed in temperature-limited currents of the same value. Since this is not the case, it is apparent that the retarding potential fluctuates with fluctuations in emission in such a way as to oppose the changes in anode current.

The method of analysis is to determine the effect on the potential minimum of a small pulse of electrons of each velocity group from zero to infinite velocity, and from this effect the change in anode current. The average rate of emission of electrons of each velocity group is known for a given cathode temperature. The mean-square fluctuations in number emitted is proportional to the average number. From this and from the effect of electrons of different velocities on the anode current, the total current fluctuations may be determined by integration.

It has been observed that fluctuations in positive-grid tubes exceed those in triodes. This difference must be due to fluctuations in division of current between positive grid and anode. Such fluctuations in division can be accounted for by the fact that, for each pulse of excess emission confined to a small area of the cathode, there must be a compensating reduction of anode current extending over a considerable area. The initial excess pulse will flow chiefly to the grid or to the anode, while the compensating pulse will divide between the two.

Fluctuations in anode current caused by collision ionization have been observed, but are of little importance under practical conditions. Ballantine's analysis is incomplete because of difficulties arising from the study of the effect of a single ion. It is more convenient to treat the ion flow as a continuous stream as is done in the case of electron currents.

The application of the methods of approach outlined in this part will be given in the later parts of this paper.

A—Introduction

FEW technical problems have been the subject of so much discussion over such a long period without general agreement as that of the random fluctuations in plate current of a space-charge-limited vacuum tube. Until recently this lack of agreement embraced experimental fact as well as theory. In this situation, it was natural that the theories should be nearly as numerous and as dissimilar as the authors thereof. Now, however, it appears that there is something approaching general agreement on the observed facts. The time seems ripe, therefore, for agreement on theory. It is the hope of the present writer and his collaborators to present in this series of papers a reasonably complete theoretical analysis of fluctuations in vacuum-tube amplifiers under operating conditions, and experimental confirmation thereof.

The work on which these papers are based extends over a period of more than ten years. While we have found ourselves in disagreement with many of the papers which have appeared on this subject, we have delayed presenting our work in published form until repeated experimental verification and continued analytical refinement, together with the acceptance of our conclusions by others, have convinced us of the soundness of our results. In the course of preparation of these papers, we have received further assurance from the publication by distinguished German workers of results closely similar in part to ours.

The extensive literature on the subject of current fluctuations is, to some extent, an indication of its importance. Vacuum-tube amplifiers enable us to perceive extremely minute electrical impulses, the minuteness being limited only by the magnitude of the inevitable extraneous background of "noise". This application of amplifiers is of major importance and "noise" is the only serious technical limitation at present.

The desired signal which is introduced into an amplifier is not, in itself, free from noise. The amplifier, however, introduces additional noise which often is greatly in excess of the inherent noise of the signal. It is recognized that both the circuits and the vacuum tubes of the amplifier contribute to the amplifier noise. In many circumstances, circuit noise predominates; in other cases—notably of late, television amplifiers—tube noise is the limiting factor.

The extraneous background noise produced by a vacuum tube may be divided into three general classifications:

1. Disturbances caused by wall charges, vibration of electrodes, faulty contacts, etc. These are largely confined to the lower frequencies and may be reduced without theoretical limit by more careful tube design and construction.

2. Flicker effect, caused by random changes in surface conditions of the emitter. This disturbance is confined chiefly to frequencies below 500 cycles per second, and may be reduced by choice and treatment of the emitter.

3. Fluctuations in space-charge-limited currents associated with the random emission of electrons. These disturbances are distributed uniformly over the whole frequency spectrum from zero to frequencies such that the electron transit time is comparable with the period. They are of a fundamental nature.

Because the first two effects are largely confined to the lowest frequencies and because they may be controlled by well-known means, this part and subsequent parts of the paper will deal solely with the third class of fluctuations.

In an attempt to present our work in the form most convenient to possible readers, whose interests may vary from the purely theoretical to the intensely practical, we shall divide the material into several parts, of which this is Part I. These parts will, in effect, be independent papers which may be read separately, though, of course, each part will use without repetition of proof the pertinent conclusions of other parts. These parts are as follows:

Part I. General Survey, by B. J. Thompson.

Part II. Fluctuations in Diodes and Triodes, by D. O. North.

Part III. Fluctuations in Multi-Collector Tubes, by D. O. North.

Part IV. Fluctuations Caused by Collision Ionization, by B. J. Thompson and D. O. North.

Part V. Fluctuations in Vacuum-Tube Amplifiers, by W. A. Harris.

The titles of Parts II, III, and IV list the major sources of noise which will be treated separately.

It is the purpose of this general survey to present the historical and theoretical backgrounds for the individual analyses. Those who are concerned solely with the application of the results will be interested chiefly in Part V, since the rest of this part and Parts II, III, and IV describe how the results are obtained.

B—Fluctuations in Space-Charge-Limited Diodes and Triodes

1. Historical Survey

Historically, the first publication concerning current fluctuations was the classic paper by Schottky[1] in 1918 in which he predicted that there would be fluctuations in the plate current of a temperature-limited diode, the mean-square value of which is proportional to the plate current, the electron charge, and the frequency interval over which the measurement is made. Subsequently this prediction was verified experimentally.

In 1928, Nyquist[2] and Johnson[3] published their important papers showing that there are fluctuations in voltage across the terminals of an open-circuited resistor, the mean-square value of which is proportional to the resistance, the absolute temperature, and the frequency interval over which the measurement is made.

It was recognized that these two types of fluctuations set a minimum limit to the magnitude of signal which may be perceived by means of a vacuum-tube amplifier. As early as 1925, however, it was recognized that the shot-effect formula of Schottky does not apply directly to space-charge-limited currents, the careful measurements of Hull and Williams[4] having shown a reduction to as low as 16 per cent of the root-mean-square temperature-limited shot effect. The question then arose as to the lowest magnitude of fluctuations to be expected under operating conditions.

One view, attractive in its simplicity, was that in the presence of "complete space charge" the fluctuations would be zero. The fact that they were not found to be zero was taken to be an indication of the lack of complete space charge.

The first attempt at a complete analysis of the problem was that of Llewellyn[5] in 1930. The fluctuations were divided into two component parts: that resulting from the lack of complete space charge (measured by the rate of change of plate current with emission current) and appearing as a fraction of the temperature-limited shot effect, and that produced by thermal agitations in the plate resistance of the tube, demonstrated by thermodynamic considerations to be effectively at the cathode temperature.

This analysis was widely accepted, but came to be recognized as suffering from the grave defect that the predicted values of shot effect or of thermal agitations taken alone often considerably exceeded the total measured fluctuations. The published measurements of Pearson[6] and of F. C. Williams[7] and unpublished work of others established this fact clearly and led to an attempt to obtain agreement between theory and measurement by dropping the shot-effect component and making the quite reasonable substitution of cathode resistance for plate resistance in computing the thermal agitations of a triode.[8] This still left an excess of "theoretical" over measured fluctuations.

F. C. Williams[7] appears to have been the first to point out in a published paper the fallacy of viewing the anode current fluctuations as thermal agitations in the internal anode resistance. It seems clear that thermodynamic reasoning cannot be applied to a situation where the drift velocities of the electrons are greatly in excess of their thermal velocities.

In November, 1936, Dr. North and I presented at the Rochester Fall Meeting of the I. R. E. a brief summary of our theoretical and experimental work which will form a considerable part of the present contribution.[9] As relates to diodes and triodes, it was shown by a microscopic study of the actual mechanism of reduction of shot effect by space charge that the magnitude of mean-square current fluctuations should be approximately six-tenths of the thermal agitations in a resistor equal in magnitude to the cathode resistance at the cathode temperature. This result was stated to be in agreement with experimental observations.

At about the same time Schottky published the first[10] of a projected series of papers;[11] this constituted the first published attempt to analyze the actual mechanism of space-charge reduction of shot effect. Unfortunately, the different effects of different velocity classes of electrons were ignored, leading to an erroneous result.

A subsequent paper by Schottky and Spenke[12] has corrected the errors and omissions of the earlier papers. So far as relates to fluctuations in space-charge-limited diodes and triodes, these writers are now in substantial agreement with the results presented by us in 1936 and now published in detail.

Rothe and Engbert[13] have published a large number of measurements on a variety of tubes which establish more firmly the experimental facts.

2. Nature of the Problem

The analysis of fluctuations in temperature-limited currents (to be called true shot effect) is based on the assumptions that the emission of any one electron is an independent event determined solely by chance and uninfluenced by the emission of other electrons at that or any other time, and that all emitted electrons instantaneously pass to the anode. Under such conditions, it is shown that the number of electrons emitted in successive equal time intervals varies according to the laws of probability and that this variation in number constitutes a spectrum of alternating-current components having constant amplitude independent of frequency. A tuned circuit through which the anode current flows will develop a voltage the mean-square value of which is proportional to the plate current, the square of the resonant impedance, and the band width over which the circuit responds. To give a numerical example, a circuit of 50,000 ohms impedance having an effective band pass of 10,000 cycles per second through which a temperature-limited current of one milliampere flows will develop a voltage of 89 microvolts across its terminals, independent of the resonant frequency. This prediction has been accurately verified.

The assumption of complete randomness of electron emission applies equally well to the space-charge-limited case, of course. It is obvious that the second assumption of instantaneous passage of all electrons to the anode is not justified, however. The mechanism of space-charge limitation of anode current is such that part of the emitted electrons are turned back to the cathode. As the total emission is arbitrarily increased—by increasing cathode temperature, for example—a larger fraction of the electrons is turned back. This results from the mutual repulsion between electrons, the presence of one electron tending to turn back following electrons. It is to be expected, then, that when an instantaneous excess of electrons over the average number happens to be emitted, a greater fraction of the total number of electrons will be returned to the cathode. It follows that the anode current fluctuations will be reduced by the action of the space charge to a value less than the true shot effect. We shall call this value the reduced shot effect.

It seems clear that it will be necessary to examine minutely the mechanism of space-charge limitation of anode current in order to determine the effect of space charge on shot effect.

Before proceeding with a discussion of the problem, it may be well to pause here and clear up several points which have tended to becloud the technical issues in the past, both by leading to experimental error and by confusing theoretical discussion.

It has been erroneously stated that, since shot effect is produced by the arrival of individual discrete electrons at the anode, nothing may be expected to reduce its magnitude for a given current. Shot effect is not a matter of the discreteness of the electron, directly. A current of 1.0 milliampere corresponds to a flow of nearly 10^{16} electrons per second. If they were to arrive at the anode at a uniform rate, there would be no frequency component below 10^{16} cycles per second.

It has been erroneously stated that shot-effect currents are the same in the presence of space charge as with temperature limitation of emission and that the reduction in shot-effect voltage results only from the shunting effect of the lower anode resistance in the case of space-charge limitation.

On the other hand, experimental results have frequently been published in which the varying shunting effect of the anode resistance was ignored.

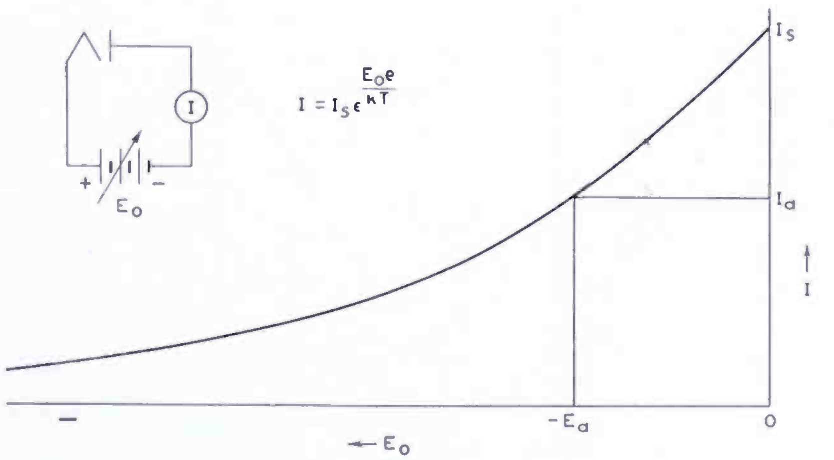

Fig. 1

Shot effect is essentially the varying rate of flow of electrons between cathode and anode brought about by the random rate of emission of electrons. The study or discussion of current flow in a vacuum tube is expedited by assuming no external impedance in any electrode circuit. We shall, therefore, make this assumption, unless otherwise stated. The fluctuation voltage in the anode circuit for any fluctuation current may be obtained in the same manner as for varying currents of other origin: by multiplying the alternating component of current by the effective anode impedance, including the internal anode resistance in parallel with the external circuit. The internal anode resistance has no unique influence on shot effect. Of course, measurements of shot effect require an impedance in the circuit. The experimental results presented in these papers represent either the current which would flow in the electrode circuit without impedance, or the voltage developed across a stated impedance, including the internal electrode resistance.

3. General Theoretical Considerations

In a temperature-limited diode, there is an accelerating field at the cathode and all emitted electrons pass over to the anode. According to the well-known formula for true shot effect, the mean-square fluctuation current $\overline{i^2}$ produced by an anode current I is $\overline{i^2} = 2eI \Delta f$ where e is the electron charge and Δf is the frequency band over which the measurement is made. *If the total temperature-limited current I is divided in any arbitrary fixed manner, the mean-square fluctuation current of any part of the current is in proportion to that part, thus corresponding to true shot effect.* This is stated axiomatically, and will be called the principle of fixed division. It applies to divided

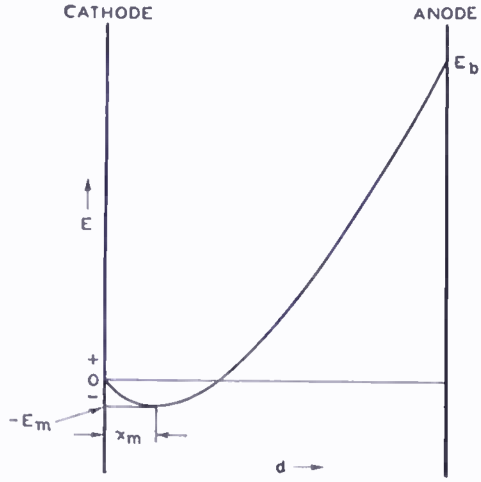

Fig. 2

cathodes, divided anodes, selected velocity groups, or any other fixed division.

Let us consider the velocity distribution of the emitted electrons, shown in Figure 1. This is really a plot of the current which would flow to a negative electrode closely surrounding the cathode, the potential of the electrode being E_o and the current flowing to it, I. As the electrode is made more negative, fewer electrons have sufficient velocity of emission to reach the electrode against the retarding field. If the potential be set at $-E_a$, a current I_a will flow to the electrode, electrons having lower velocity being turned back. Since we are fixing the division of the total current by the retarding potential $-E_a$, there will be true shot-effect fluctuations in current I_a.

The potential distribution in the space between cathode and anode in a parallel-plane diode with space-charge-limited current is represented in Figure 2. Near the cathode there is a potential minimum of magnitude E_m (negative) at a distance x_m from the cathode. Between this potential minimum and the cathode, all electrons having an initial velocity of emission E_o normal to the cathode less than E_m (expressed in electron volts, the actual velocity being given by $v = \left(\frac{2E_o e}{m}\right)^{\frac{1}{2}}$ where v is the velocity, e is the electron charge, and m is the mass of the electron) are turned back to the cathode. All electrons with greater initial velocity pass to the anode. The values of x_m, E_m, and plate current I_b, may be determined for given cathode temperature, electron emission, electrode spacing, and anode potential from well-known analyses.[14]

From our discussion of the fluctuations in current flowing to a retarding electrode, we must conclude that, *if E_m remains constant, the fluctuations in plate current of the space-charge-limited diode must be*

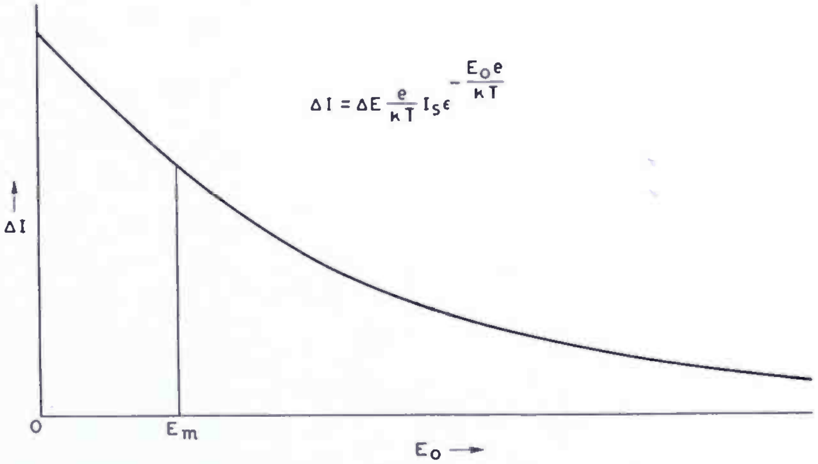

Fig. 3

equal to the true shot-effect value. The study of the effect of space charge on shot effect is then the study of the behavior of the potential minimum with fluctuations in emission.

Qualitative considerations show that E_m does not remain fixed, however. Its value is determined at any instant by the instantaneous space-charge distribution between cathode and anode. When the number of electrons having velocities of emission in excess of the mean value of E_m is momentarily in excess of the average, the number of electrons passing across the cathode-anode space increases and, as a result of their space-charge effect, E_m becomes more negative, thus turning back a certain number of electrons which would otherwise have passed to the anode. When the number of high-velocity electrons is less than the average, E_m becomes less negative than the mean value and some of the lower-velocity electrons are permitted to pass over to the anode.

To determine the magnitude of the reduced shot effect it is necessary to analyze the quantitative relations between instantaneous velocity distribution and instantaneous value of potential minimum.

Let us consider again the velocity distribution of emitted electrons. Figure 3 shows the current corresponding to the average number of electrons in each velocity group $(E_o + \Delta E)$, where velocity of emission E_o is expressed in volts as before. The total area under the curve (on to infinity) is equal to the total average emission from the cathode. The area under the curve to the right of E_m (the potential minimum) is equal to the average anode current. The area to the left of E_m represents the current turned back to the cathode by the potential minimum. This curve represents average conditions, only. From our previously stated principle of fixed division, we know that the current corresponding to *each* velocity group of electrons fluctuates at random. If the velocity increment is made small enough, all electrons in a given

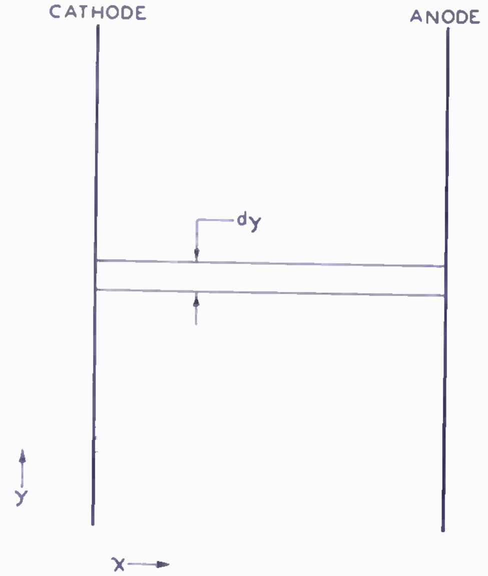

Fig. 4

velocity group have the same space-charge effect on the potential distribution and, hence, on the potential E_m. Knowing as we do the fluctuations in number of electrons in each group and being able to calculate the effect of each group on the potential E_m, we are in a position to determine the over-all net fluctuations.

To simplify the physical viewpoint in this analysis, let us consider a vacuum tube in cross-section, Figure 4. We may say that, since the fluctuations in current are small compared to the average current, the fluctuations in E_m are small compared to the average value of E_m. We may then apply the principle of superposition. Let us say that the emission is constant outside of the differential strip *dy*. Within *dy* it fluctuates at random. E_m is, of course, constant except for infinitesimal changes produced by fluctuations in emission within *dy*. An excess number of electrons of any velocity group emitted within *dy* will pass to the anode if their initial velocity E_o is greater than E_m or will be returned to the cathode if it is less than E_m. Whichever course they take, their effect on E_m (an infinitesimal effect) may be calculated. Suppose a group of electrons having a velocity greater than E_m be emitted. These electrons pass on to the anode and, on their way, add to the space charge and depress the potential E_m over a considerable area. This lowering of E_m turns back some of the lower velocity electrons in the region adjacent to dy, so that, in effect, there is a compensating current flowing in the opposite direction which reduces the effect of the original group of electrons.

If the velocity of this group of electrons is less than E_m, none of the electrons reach the anode but E_m is depressed by the additional space charge, with the result that there is a reduction in anode current.

For each velocity group of electrons in dy which reaches the anode there is, then, a compensating current in the opposite direction the magnitude of which depends on the velocity of the electrons. The net current pulse due to this group of electrons is the difference between the current flow corresponding to the electrons and the compensating current.

For each velocity group of electrons which fails to reach the anode, there is a negative pulse of anode current, the magnitude of which depends on the velocity of the electrons.

By integrating in the proper way the effect of each velocity group of electrons in dy on the resulting anode current we may find the net fluctuations produced by the random emission of all the current in dy. Integrating over the entire cathode area will then give us the total fluctuations in current.

Simple considerations show that it is really unnecessary to integrate over the whole cathode surface, however. The effect of a given velocity group of electrons emitted in one place is the same as that of a similar group distributed over the entire cathode surface, so long as the group is infinitesimal in size. We may, then, for analytical purposes equally well consider the velocity groups distributed over the entire cathode surface. The actual mathematical process to be followed will then be the dividing of the total cathode emission into an infinite number of velocity groups, the multiplying of the mean-square fluctuation current corresponding to each velocity group by a factor indicating the relative effect of electrons of each velocity on the anode current, and the integrating of the product for the entire range of velocities. The result is the total mean-square fluctuation in anode current.

In the above discussion it is implicitly assumed that a fluctuation involving a given velocity group of electrons persists long enough for a new equilibrium to be established. This simplifies the analysis, but renders it inapplicable to frequencies such that the transit time of an electron is comparable with the period—frequencies of several hundred megacycles per second in usual tubes.

The application of the diode analysis to triodes is obvious. We see that the addition of a negative grid between cathode and anode has little effect on the fluctuation current for a given anode current, but chiefly serves to alter the internal anode resistance of the tube.

It has been customary to view reduced shot effect and thermal agitation as identical, or at least closely related. The above discussion had laid the background for an analysis without mention of thermal agitations nor has any mechanism for establishing thermal equilibrium between the electron stream and the cathode been found or assumed. Does this mean that the thermal-agitations viewpoint is considered false?

The answer is yes to the extent that that viewpoint is used as a means of predicting the reduced shot effect without detailed microscopic analysis. Reduced shot effect is essentially shot effect in the presence of space charge and should be approached from the shot-effect and space-charge viewpoints. The thermal-agitations viewpoint is useful only in the interpretation and application of the results.

Part II continues the above analysis and presents an experimental verification of the results.

C—Fluctuations in Positive-Grid Tubes.

1. Historical Introduction.

Comparatively little attention has been accorded to the question of current fluctuations in screen-grid tubes, it being generally implicitly assumed that, where the cathode current divides between two electrodes such as the screen and the anode, the fluctuations would divide in the same manner. (Here I ignore the question of thermal agitations in the plate resistance, for so far as I know this concept has never been seriously applied to screen-grid tubes.) It was fairly generally recognized, however, that triodes were "quieter" than tetrodes and pentodes, though a very great deal of confusion was introduced by a general failure to take into proper consideration the greater shunting effect of the lower plate resistance of triodes.

About 1933, Mr. Stuart Ballantine very courteously told me of his experimental results which showed greater fluctuations in the anode current of screen-grid tetrodes than in that of triodes. He was able to account for this by the fluctuations in emission of secondary electrons from the screen grid. Measurements which I undertook as a result of this with suppressor-grid pentodes seemed to show that this explanation, while undoubtedly based on an existing phenomenon, was incomplete, since no important difference in fluctuations appeared to exist between tetrodes and pentodes although it was known that the suppressor grid effectively prevented secondary emission.

In 1935, Messrs. S. W. Seeley and W. S. Barden called to the attention of Dr. North and me their ingenious measurements[15] with suppressor-grid pentodes in which a resistance in the cathode circuit (see Figure 5) should "degenerate" practically all cathode-current fluctuations, and, therefore, should eliminate the component of anode-current fluctuations resulting therefrom. Nevertheless, they observed

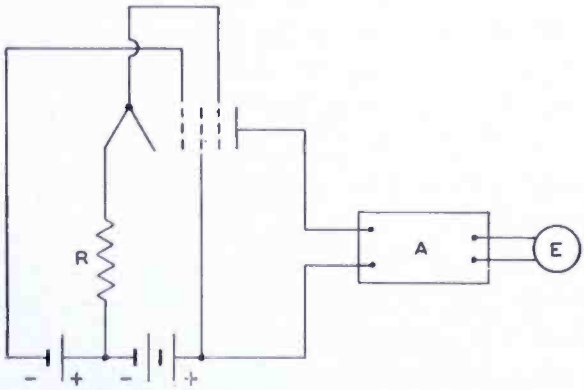

Fig. 5

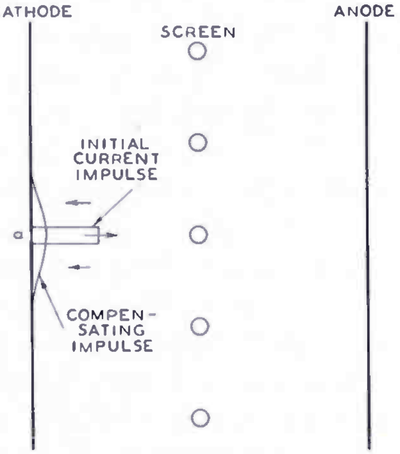

Fig. 6

large fluctuations in the anode current. These could not be accounted for by thermal agitations in the cathode resistor. The conclusion was, therefore, that there was a fluctuation in division of current between screen and anode. That this was so they verified by direct measurement.

On the explanation of the Seeley and Barden effect is based the analysis presented in Part III by North. This analysis was made immediately after the discussion with Messrs. Seeley and Barden. It was presented in brief at the Rochester Fall Meeting of I. R. E. in November, 1936.[9]

In December 1937, Rothe and Engbert[13] published a qualitative theory which explains the fluctuations in division of current between screen and anode as resulting from the chance determination of the destination of the electrons which either strike close to edges of the screen wires or pass on to the anode after narrowly missing the wire. The measurements which they report are in agreement with those reported by us.

2. Theoretical Considerations

The explanation of the fluctuation in division of the current between screen and anode which was offered (and later developed in detail by Dr. North) is as follows:

Consider a vacuum tube as shown in cross section in Figure 6. Suppose that an excess number of electrons of high velocity is emitted over some short period of time at point (a). These electrons will pass through the region of potential minimum and will pass on to the positive screen-grid wire directly before them (the negative control grid is omitted for simplicity). On their way, they increase the space charge of the space through which they pass to greater than the normal value. This increase in space charge depresses the potential of the potential minimum, thus bringing about a reduction in current below the normal value in the area surrounding the closely confined initial impulse in the manner discussed under (C). This reduction in current is in effect a compensating current flowing in the opposite direction to the initial impulse. The difference between the two currents is the net fluctuation which would exist in a diode.

In the case of the positive-grid tube, however, the initial impulse just considered went entirely to the screen grid. The compensating current, being distributed over a considerable area, divides between screen and anode almost as the average current divides. If the screen-grid current is a small fraction of the anode current, the screen-current fluctuations will be almost uncompensated and it would, therefore, be expected that the magnitude of these fluctuations would be given by the formula for true shot effect. If the net cathode-current fluctuations are small, the plate-current fluctuations must be approximately the same as the screen fluctuations.

Of course, this explanation ignores the effect of the cathode resistor in the Seeley and Barden experiment. The degenerative fluctuation voltage fed back to the grid through the action of this resistor merely makes more certain that the compensating current is distributed over a large cathode area.

Rothe and Engbert[13] have assumed that the arrival of a certain class of electrons at the screen is entirely fortuitous. That is, electrons emitted directly under the edge of the screen wire have the same chance of reaching the screen wire as of reaching the anode. This leads to the conclusion that some fraction of the current in the screen should fluctuate at random.

This last analysis should lead to correct results for extremely minute grid wires, such that the motions of electrons in directions parallel to the cathode as the result of random velocities of emission would be large compared to the grid wire diameter. It appears, however, that it would be difficult to apply quantitatively in the practical case. The analysis presented by North should give quantitatively correct results in the case of infinitesimal as well as of finite collectors.

In the event that an intermediate collector of current intercepts a large continuous area, the conclusions arrived at by North would be invalid. This is because the fluctuation in distribution of current could occur only at the boundaries of the large area.

D—Fluctuations Caused by Collision Ionization

No vacuum tube is perfectly evacuated. The residual gas molecules will occasionally be struck by an electron on its way from cathode to anode. If the velocity of the electron is sufficient, the molecule may lose an electron, thus becoming a positive ion. The effect of the positive ion is well known to be that of increasing the electron current which may flow across the space. Inasmuch as the formation of positive ions is a random event, the current produced by them must fluctuate at random, and thus be a source of noise.

The existence of such an effect has long been recognized. Ballantine[16] in 1933 analyzed the effect of a single ion and from that arrived at the nature of the qualitative relationship between ion current and plate-current fluctuations. He found experimentally that this relationship was correct. It was not possible, however, to determine analytically the magnitude of fluctuations to be expected.

In Part IV of this paper, I present an analysis which was carried out in 1933. This treats the flow of positive ions toward the cathode as the movement of a positively charged continuous stream through the negatively charged electron stream, in the same manner as is conventionally done in treatments of electron-space-charge effects. This is a great simplification as compared with Ballantine's analysis of the effects of a single ion. It is justified because it may readily be shown that the average distance between ions is extremely small as compared with electrode spacings, even under conditions most unfavorable to the assumption.

E—Other Sources of Fluctuations

We have already considered the fluctuations caused by random emission of electrons from the cathode, by inconstant division of current between positive electrodes, and by varying production of positive ions. There are other sources of fluctuations.

It is well known that positive ions may be emitted from a hot cathode or may be formed at its surface and that these ions cause fluctuations in cathode current.[17] Experimentally, we have detected the effects of such ions only in special cases. Under normal conditions this source of fluctuations need not be considered.

Flicker effect, wall charges, microphonics, etc., have been mentioned and dismissed from our present consideration.

Secondary emission from the screen grid plays a small part in determining the fluctuations in current of screen-grid tetrodes. The prevailing use of suppressor-grid pentodes makes it unnecessary to concern ourselves with this.

Grid current produced by gas ionization will develop a fluctuation voltage in an impedance in the grid circuit according to the shot-effect formula. With usual values of grid current this fluctuation voltage may be ignored, though it may readily be taken account of if necessary.

The analyses which have been made do not apply at the highest frequencies. Further, at ultra-high frequencies there are other sources of noise. These matters we hope to discuss in another paper.

F—Conclusion

This introductory part of our paper has, by its nature, been devoid of results. The succeeding parts, of which the next is Part II, Fluctuations in Diodes and Triodes, by D. O. North, will continue the analyses by more rigorous methods and will present the results.

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TELEVISION RECEPTION IN AN AIRPLANE

By

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ON THE occasion of the twentieth anniversary, October 17, 1939, of the formation of the Radio Corporation of America, a demonstration of television reception in an airplane was given for the press in cooperation with United Airlines. The principal feature of the demonstration was the reception of television images in the plane while flying at about 21,000 feet over Washington, D. C., an airline distance of about 200 miles from the transmitter on top of the Empire State Building in New York City.

As part of this demonstration a two-way conversation was carried on between the television studio and the airplane. Mr. David Sarnoff, President of the Radio Corporation of America, and Mr. W. A. Patterson, President of the United Airlines, conversed with several members of the group in the plane, who were watching them on the television receiver. This conversation was broadcast over the Blue Network of the National Broadcasting Company and both the studio picture and the conversation were broadcast over television station W2XBS.

A further interesting feature of the demonstration occurred when a landing of the plane at North Beach Airport was televised by the NBC Telemobile Units, relayed to Empire State, and observed on the screen of the receiver inside the plane. Thus the people in the plane saw themselves land as viewed from the ground.

The receiver used for this demonstration was a standard television broadcast receiver, Model TRK-12, which is equipped with a 12-inch kinescope. It was operated on a 50-cycle power supply provided by a rotary converter driven by storage batteries. Close speed regulation of the rotary converter was not necessary, since the normal variations in power supply frequency with respect to the vertical scanning frequency had no effect upon the operation of the receiver. A radio-frequency amplifier was added to the receiver to increase its sensitivity.

The receiving antenna was a half-wave dipole mounted crosswise under the belly of the ship as shown in Figure 1. A transmission line about 15 feet long connected it to the receiver, which was mounted in the forward end of the passenger compartment as shown in Figure 2.

The airplane was a Douglas DC3 21-passenger United "Mainliner." Two seats were removed to make room for the receiver and power equipment.

During the test flights preparatory to the demonstration and

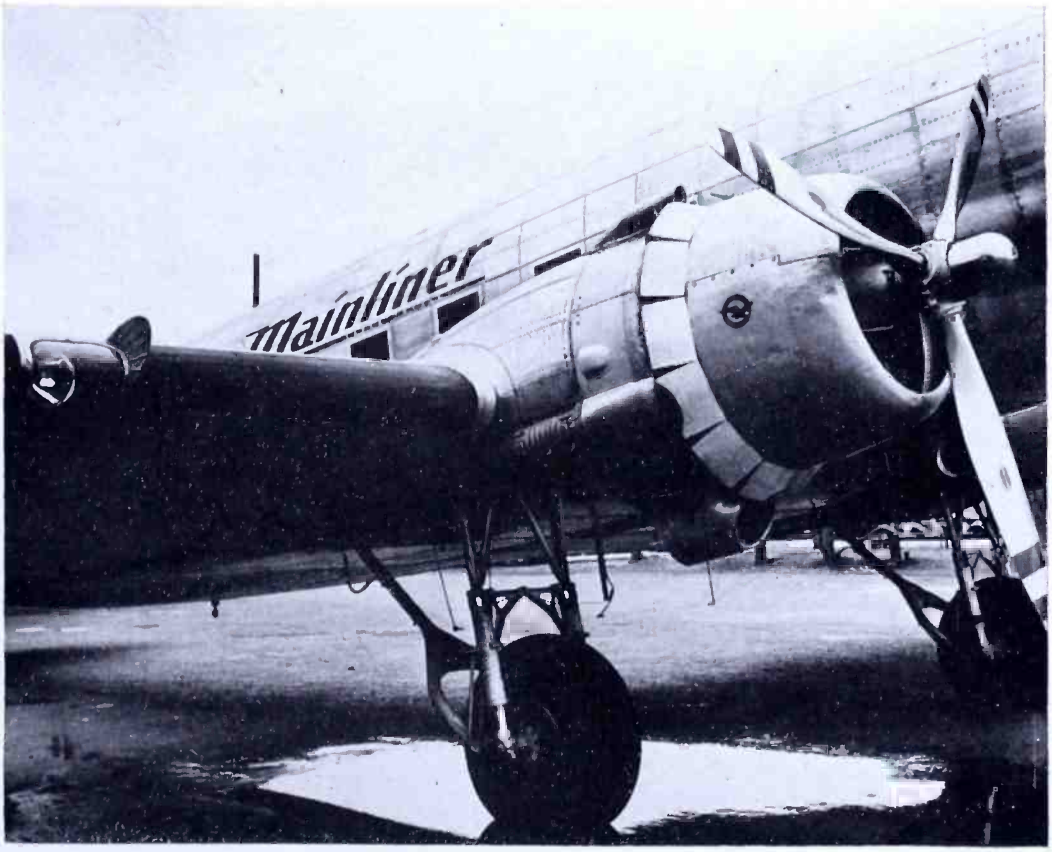

Fig. 1—Illustrating the half-wave dipole antenna mounted crosswise under the ship.

during the demonstration itself several observations of a technical nature were made. Flights were made on four different days. The first flight, on October 2, was a short hop around New York. As to be expected in an airplane, reflections from ground and buildings produced multiple images that were observed when the plane was close to the transmitter. There were innumerable multiple images, both positive and negative, spread almost completely across the picture, some moving rapidly and others nearly stationary. All of these multiple images decreased in intensity as distance increased from the transmitter and they disappeared completely, leaving a clean picture, at about 25 miles from the transmitter. For the most part the multiple images were much lower in intensity than the main signal, but at infrequent intervals they would rise in amplitude to equal, or even exceed, that of the main image.

On the other test flights the plane was flown from New York to Washington and back. In general the picture was very good so long as the plane was within line of sight from the top of the Empire State Building.

Fig. 2—The television receiver installed in the forward part of the passenger compartment.

During the second flight, on October 4, there was a heavy overcast about 300 or 400 feet thick at 4500 feet. Below the overcast it was hazy, and above, it was entirely clear. On this day no signal was received directly over Washington at 17,500 feet, which is approximately line of sight for this 200-mile path. The signal was picked up about 20 miles northeast of Washington at this elevation, and was satisfactory most of the way back to New York.

On the third test flight, October 16, the weather was clear, with no clouds at all. The signal was received all the way to Washington so long as the plane was in line of sight. In fact the signal was reasonably good at 16,000 feet over Washington and improved with elevation up to 21,600 feet, the highest altitude attained.

On the day of the demonstration, October 17, there were scattered clouds. The signal on this day appeared to be maximum over Washington at about 18,000 feet, was somewhat less at 21,000 feet, and disappeared at 15,000 or 16,000 feet.

Sufficient data were not obtained to determine why the signal distribution varied with elevation on the different days. Exact field strength measurements were not made. Information on the weather has been included to indicate a difference in conditions during the several days of tests. Some interference in the picture was experienced from electrical equipment and motor ignition systems in the airplane. This was corrected by adjustment of the ground and shield connections. Other interference, also created in the airplane and probably due to intermittent bonding connections or vibrating control wires actuated by motor vibration, was not entirely eliminated. Severe diathermy interference was experienced at times. No difficulty was experienced with the receiving equipment from vibration of the plane or from operation at high elevations. The plane carried oxygen equipment for use of the passengers and crew at the unusual elevations where most of the tests were made.

These tests were carried out as a joint project of RCA and United Airlines. The writer wishes to acknowledge the cooperation given by Mr. Stangby of United Airlines, Mr. W. A. R. Brown of NBC, the UAL staff at the Newark Airport, and the NBC television staff.

RCA TELEVISION FIELD PICKUP EQUIPMENT

By

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ON December 1, 1939, the RCA Manufacturing Company demonstrated its recently developed television field pickup apparatus in the first showing which has been made of the equipment outside of the Camden laboratories. The apparatus was set up at Washington, D. C., and demonstrated to the Federal Communications Commission to illustrate the progress which has been made in providing field equipment which approaches the compactness of apparatus used for present-day remote sound broadcasting.

From observations made in New York and in London, it has been apparent that television programs showing sports, news events, parades, etc., have met with great public interest. Unfortunately, however, the apparatus which has been required to broadcast such programs by television, has been costly, complicated, and inconvenient to move about from one location to another. The first attempt which was made in this country to provide apparatus which could be used regularly for field transmissions resulted in the RCA-NBC mobile units which are now in service in New York. The camera and video equipment is located in one of these units; the apparatus employed is standard rack and panel equipment and includes a full-sized electronic synchronizing generator. The truck contains two camera channels and while the cameras may be taken out of the truck and extended for a distance of approximately 500 feet, the remainder of the apparatus is permanently mounted in the van. The second of the mobile units contains the radio picture transmitter and the video terminal equipment. The associated transmitter apparatus includes the power supply units and a water-cooling system. The radio transmitter is provided with an antenna mounted on the roof of the transmitter truck. This antenna may be raised or lowered. The radio transmitter operates on a frequency of (approximately) 159 megacycles. These mobile units have proved exceedingly useful in picking up interesting and current programs taking place in the New York area and have been employed for the transmission of boxing matches, wrestling, baseball and football games, events at the New York World's Fair, parades, etc. They have contributed greatly to the interesting program material which has been transmitted over the NBC television station W2XBS. Unfortunately, however, since the vans are large and heavy and since the extension of the cameras from the trucks is limited, it has been impossible to transmit some programs because conditions were such that the trucks could not be taken to the scene of action.

During the early summer of this year the RCA Manufacturing Company began the development of a new type of television pickup apparatus which would be considerably more flexible in application

Fig. 1—FCC members see new RCA television field equipment, (left to right) James Lawrence Fly, Chairman of the FCC, handling the television camera; Commissioners Thad H. Brown, Norman S. Case, and T. A. M. Craven. The complete unit is not only more efficient than, but also approximately one-tenth the weight and one-sixth the cost of the only other unit ever designed in this country for the same purpose.

and which would incorporate the improvements which had been found desirable through experience on the NBC mobile units. This apparatus was to be constructed in such form that it could be mounted in light trucks or could be taken out of the trucks and carried to a location when required. The equipment was to be suitable for use with one, two or three cameras and was to be provided with another auxiliary input as, for example, the output of a television receiver which might be used for picking up another television program. In addition, it was considered desirable to employ a lower powered radio transmitter for relaying the signals back to the main station, this transmitter being portable so that it might be taken to the top of a building and used with a directive antenna system in order to extend its range. It was also considered desirable to utilize a higher frequency so that a relatively small directional antenna system could be employed and also so that the use of channels which might be needed at some later date for television broadcasting service to homes could be avoided. Consequently, the carrier-frequency range of the transmitter was specified as being from 288 to 346 megacycles.

The first apparatus was to be designed for use with either regular-sized, or with 4″ Iconoscopes. However, provision was to be made in the

Fig. 2—Illustrating a single camera video equipment setup.

design of the equipment so that it might be modified for use with "Orthicon" apparatus at a later date.

Consequently two sets of equipment were manufactured. The first employs standard Iconoscope cameras and was delivered to the Don Lee Broadcasting System for use in Los Angeles. This apparatus employs two Iconoscope cameras although one of these will be replaced by an "Orthicon" camera chain and a second "Orthicon" chain will be added at a later date to provide for a total of three cameras. The second equipment has been constructed for the National Broadcasting Company in New York and employs two of the small-sized Iconoscope cameras with provision for adding a third camera chain at a later date. The small Iconoscope cameras were specially constructed and are designed to provide for maximum portability and minimum obstruction of view. It was the equipment built for NBC that was used in Washington for the television demonstration to the Federal Communications Commission.

At the FCC demonstration, one of the television cameras was set up at the corner of Pennslyvania Avenue and Thirteenth Street at ground level and the other was located on the top of the Post Office Building where it could be used to present a panorama of the city of Washington. The video equipment was located near the first camera in the open and the transmitter, which operated at 288 megacycles,

Fig. 3—Watching the results of the demonstration as shown on an RCA receiver in the office of Commissioner T. A. M. Craven are, left to right, Ralph Beal, RCA Director of Research; James Lawrence Fly, chairman of the FCC, and Dr. Charles B. Jolliffe, head of the RCA Frequency Bureau.

was set up on the ground nearby. The master receiver was located in the office of Mr. A. D. Ring, Assistant Chief Engineer of the Commission. This installation simulated the receiver installation which would normally be made at the main transmitter installation of a television broadcasting station. Viewing receivers were connected by cable to this master receiver, two of these being located in the office of Commissioner T. A. M. Craven and one additional receiver in Mr. Ring's office. These additional receivers might be compared to home receivers of a television broadcasting system, and were RCA

TRK-12 instruments. A standard RCA broadcast field amplifier and broadcast pack transmitter were employed for the sound channels and a separate sound receiver was also located in Mr. Ring's office. The power to operate television apparatus was obtained from a 110-volt, 60-cycle source in the Post Office Building. The camera equipment was set up and tested prior to the demonstration which required approximately thirty minutes' time to have it in proper operating condition. It naturally required additional time to set up and operate the

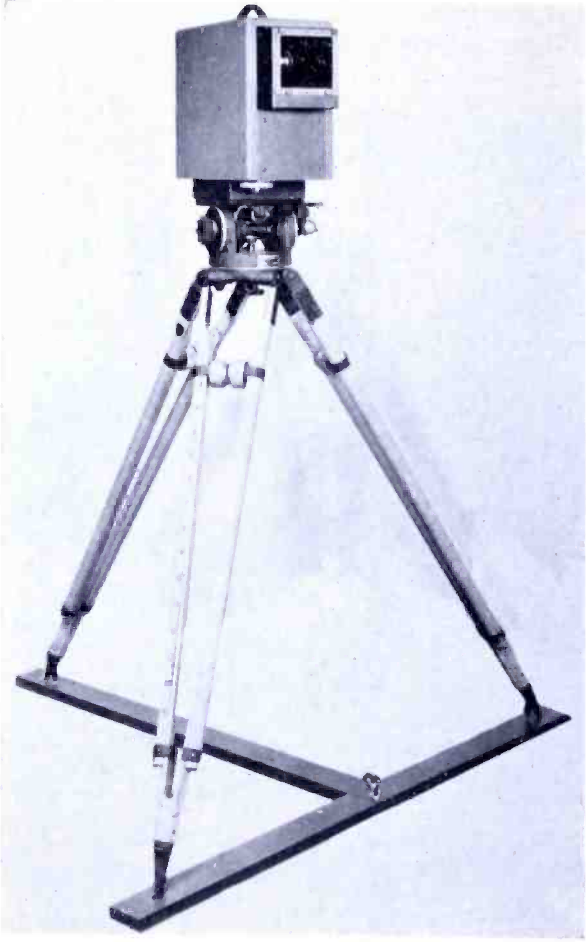

Fig. 4 — The portable television camera erected on its portable tripod mounting.

transmitter together with its directive antenna. The receiver installation was a more lengthy job, but since this receiver would be permanently located at the main studio in a regularly established service, the time required for this installation may be overlooked.

The television picture antenna employed a horizontal dipole with a reflector consisting of a number of elements located in the shape of a wedge, and placed about five feet above the ground. The receiving antenna was a directive array similar in general design to the one employed on the top of the RCA tower in the New York World's Fair. The receiving and transmitting antennas were approximately one city block apart. They were not, however, line-of-sight and it was found that best results were obtained by directing the transmitting antenna toward another building approximately two blocks away and permitting the reflected signal to strike the receiving antenna. No difficulty was experienced with multiple reflections in the received picture.

At the time of the Commission demonstration, the Secretary, Mr. Slowie, was interviewed before the camera and the television camera was used to show a picture of the apparatus being employed for the demonstration. A switch was then made to the roof camera and scenes of the city were shown. The ground camera was then used to show

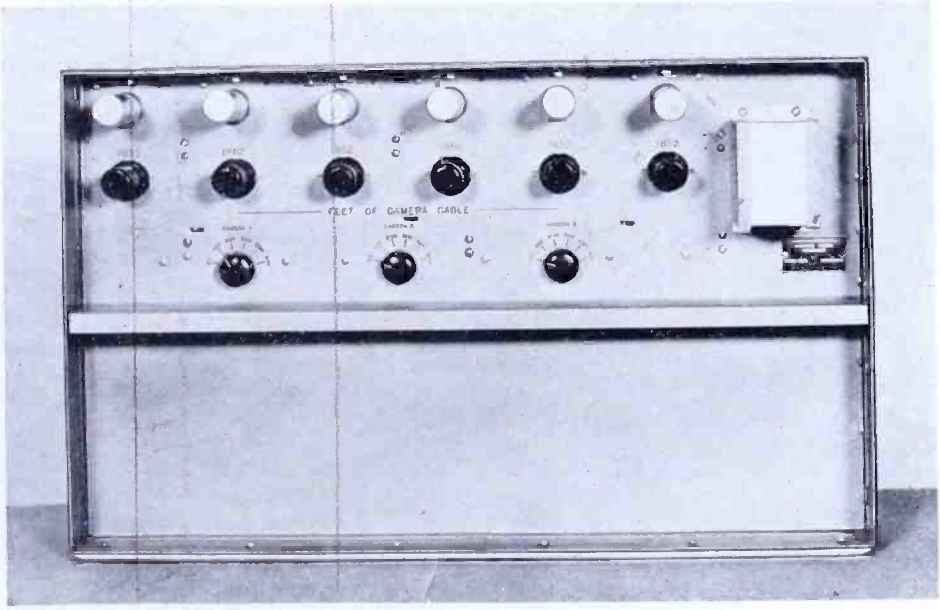

Fig. 5—Side view of delay unit, with cover removed. Controls are calibrated directly in feet of camera cable.

street scenes and the Commissioners were then invited to inspect the apparatus at the street level. Dr. C. B. Jolliffe and Mr. R. R. Beal conducted the showing in Commissioner Craven's office. The demonstration was directed by Mr. Harry Sadenwater, Television Project Engineer. The equipment was kept operating for several hours afterward in order to permit members of the Commission's personnel to see the demonstration. The field equipment consists of the following units:

Television Camera

The cameras employed are either of the standard Iconoscope type or the small Iconoscope type, the latter-type camera having been designed to be as small and compact as possible. Focusing is accomplished by means of a Selsyn motor remotely controlled by the operator who views the picture in a monitor on the camera control unit. The camera employs a single amplifier stage and the additional video amplifiers and deflection amplifiers are located in the camera auxiliary unit.

This unit is approximately suitcase size and connects with the camera by an eight-foot cable. It contains its own a-c power supply. It is connected with the camera control unit and may be separated by

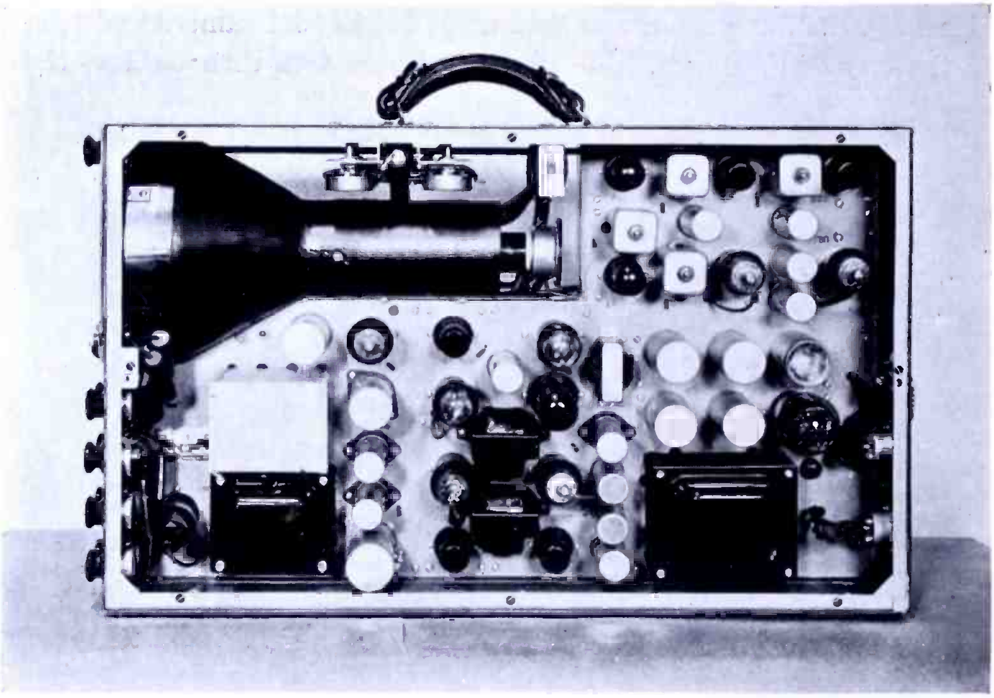

Fig. 6—Side view of camera control unit with cover removed. The center web mounting makes all parts unusually accessible.

a distance of approximately 500 feet. The camera control unit contains shading, gain, and contrast controls, and a 7-inch Kinescope for picture monitoring as well as a 2-inch oscilloscope for purposes of setting levels. If only one camera is employed, the output of the line amplifier contained in this unit may be fed directly into the radio transmitter. However, if more than one camera is used, the output is fed into the master control unit.

This unit is employed for switching between cameras and is provided with four push-button controls. These are employed for selecting any one of three cameras, or an auxiliary video input position. This unit also contains a 7-inch picture monitor and a 2-inch oscilloscope for viewing the signal which is fed to the transmitter. The picture which is seen on the monitor of this unit changes when cameras are switched whereas the picture seen on the individual camera control units remains fixed for each individual camera. The master control unit also contains a line amplifier for feeding the signal to the transmitter.

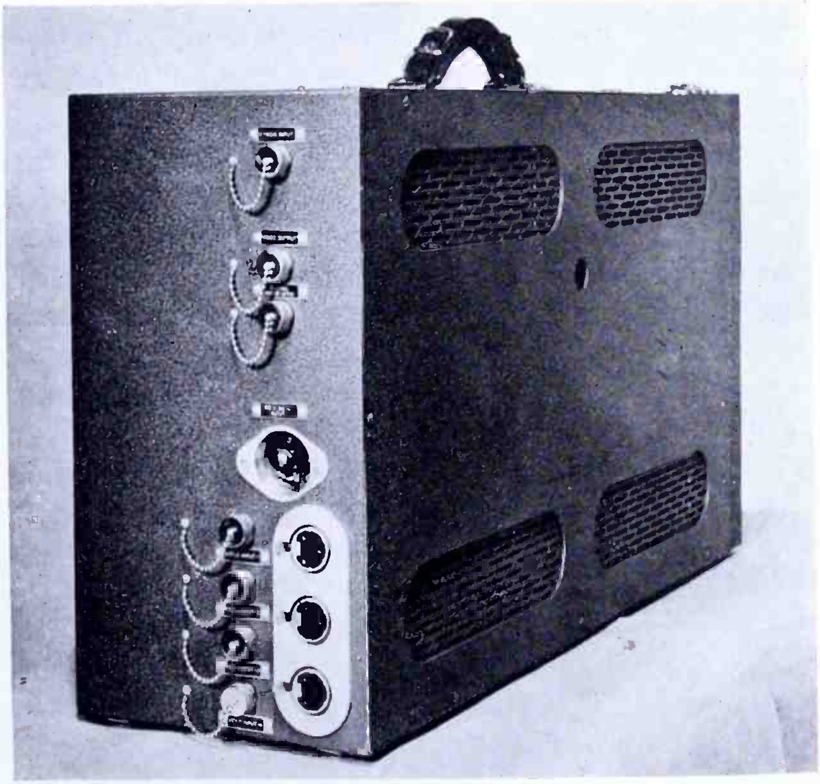

Fig. 7—Rear view of master control unit, showing cable connections and typical case construction.

Synchronizing Generator

The synchronizing generator is comprised of three units. The first contains the pulse generator, which employs a rotating disc driven by a synchronous motor. A capacity pickup is used to produce the impulses. The shaping unit contains the vacuum tubes required to shape the impulses into a signal form of the character recommended by the RMA for standardization. The third unit, the delay unit, provides compensation for the various lengths of camera cables which may be employed. This latter unit is not required if all camera cables are the same length or if only one camera is used. A power-supply unit enables all of the cases to be connected into one box and a common power-supply lead to be run to a convenient point. All of the apparatus operates from a 110-volt, single-phase supply.

The radio transmitter has a peak output power of approximately 25 watts. It is crystal-controlled and has a frequency-response characteristic which is substantially flat up to nearly 7 megacycles. It is arranged to provide for d-c transmission and is equipped with a small 2-inch monitoring oscilloscope for checking the output. The station-type receiver is a superheterodyne circuit fixed-tuned to frequency and a-c operated. It is designed for rack mounting and may be used in conjunction with a standard picture-monitor unit. Interconnecting cables are supplied between all of the camera units and the transmitter.

The apparatus has been developed by the Research and Transmitter Engineering groups of the RCA Manufacturing Company of Camden, N. J. Several of the ideas which were originally incorporated in apparatus developed by the RCA Radiotron Division have been employed in this commercial apparatus. During the design and construction of the apparatus, representatives of the National Broadcasting Company were present and assisted from their knowledge of the operating problems which have been encountered with the mobile units.

It is believed that apparatus of this type will assist greatly in the production of television programs, and that since the cost of the equipment is but a small fraction of the cost of the former apparatus, it will be of material aid to television stations in the establishment of service.

SIMPLIFIED TELEVISION I-F SYSTEMS

By

Garrard Mountjoy

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Summary—Two examples of television picture i-f amplifiers are described in this article, one for receivers requiring a relatively narrow band, and the other for wide band receivers. These show the simplicity which results from the choice of mutual inductance coupling and capacitance tuning, and demonstrate that this simplification may be secured together with good performance. Several practical types of transformers and trap circuits are discussed, and choice among these may be made on the basis of design and performance requirements of a particular amplifier. Consideration of the principles and illustrative examples discussed will enable the engineer to design a television i-f amplifier to meet desired requirements and to predict the performance thereof with a useful degree of accuracy.

Mutual Inductance Coupling

MUTUAL inductance coupling, where the windings are in proximity to each other, greatly simplifies amplifier design and construction. It not only makes unnecessary any physical coupling elements, but also permits winding all inductances per stage on a common form. The d-c isolation of grid and plate circuits is inherent so that no filter components are required for this purpose.

Capacitance tuning is indicated with mutual inductance coupling in order that alignment and adjustment may be accomplished without disturbance of coupling magnitude. In order to maintain good stage gain the capacitors should have low minimum capacitance and be set, for the average case, at a mean value which will just permit the compensation of expected production variations. For stability of adjustment, air-dielectric trimmer capacitors have been found desirable. Such capacitors are available with a range of 2 to 8 $\mu\mu$f, and if the circuits are designed for a setting of 5 $\mu\mu$f, adjustment range is provided to compensate for usual manufacturing tolerances.

In high-gain amplifiers, it may be necessary to shield one stage, usually the last transformer, to reduce tendency to regeneration. If the transformers are mounted below the chassis and single-ended tubes and chassis bottom plate used, direct i-f signal pickup may be reduced to a satisfactory low level. Shielding the other transformers for electrical reasons is not necessary.

Self-Tuned Primary

If the resonant circuit components are held to sufficiently close limits alignment adjustment means may be eliminated on at least one circuit of the stage. Flat frequency response may still be secured through the compensating effect of adjustment on other circuits, the effect of small variations in a reactance being to shift the entire response curve to one side or other of the normal frequency.

The use of trimmer capacitors involves some small loss in gain and the circuit to be self-tuned should be the one with least value of

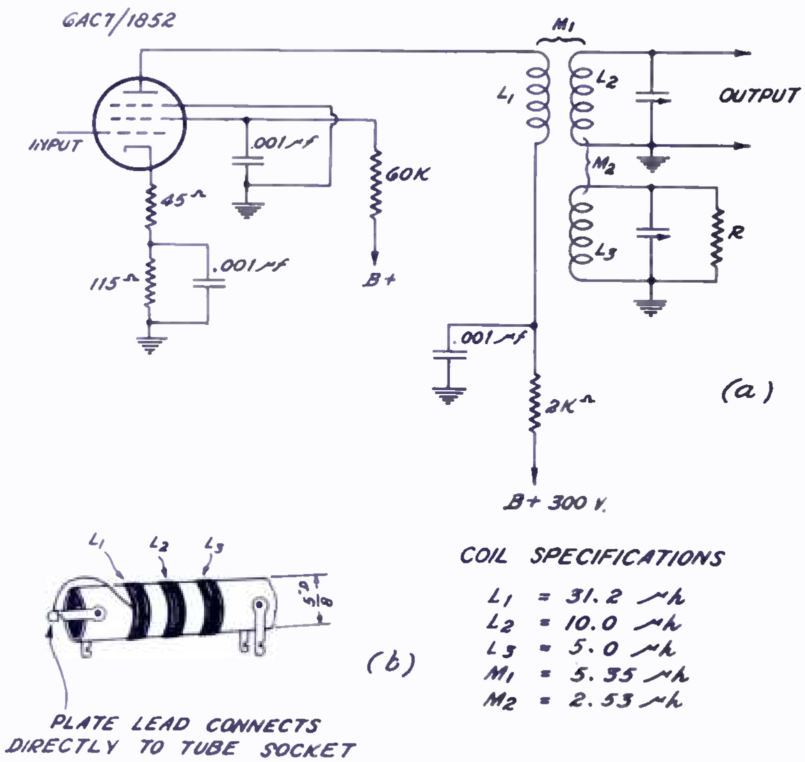

Fig. 1

fixed capacitance, as the trimmer would have most effect in reducing gain on that circuit. The primary circuit is therefore logically the one to be self tuned when tubes such as the 6AC7/1852 or 6AB7/1853 are used, because the plate capacitance is approximately half the grid capacitance.

Measurements on a somewhat limited number of samples of 6AC7/1852 type tubes have indicated that the output capacitance varies from approximately 4.9 $\mu\mu$f to 5.4 $\mu\mu$f, or a variation of $\pm$0.25 $\mu\mu$f about a center value of 5.15 $\mu\mu$f. It would appear from these measurements that there is a possibility of omitting the plate circuit trimmer if manufacturing tolerances on circuit components are held to close tolerances and careful attention is paid to layout and wiring. A production of tubes may show a wider variation. The receiver manufacturer may, if he considers it necessary, test and use only close tolerance tubes, placing those rejected for capacitance in the adjusted stages and other portions of the receiver.

Inductor values may easily be held to adequately low tolerances, and the method of transformer mounting be made uniform in stray capacitance. Soldering the plate lug of the transformer to the plate prong of the tube socket, will eliminate a lead and attending variations in capacitance. The chief source of variation will under these conditions be tube capacitance. If the total circuit capacitance in the primary is say 8 $\mu\mu$f and a variation of ± 0.25 $\mu\mu$f capacitance occurs, the shift in the center frequency of the stage, after the adjustments of other circuits have produced flat response, will be ∓ 0.15 Mc. If the stage is made inherently 0.15 Mc broader on each side of center frequency, the variations may occur without loss in overall fidelity.

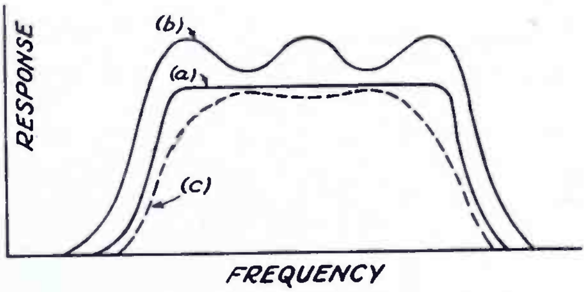

Fig. 2

The stage or stages which produce the carrier side slope may be completely adjustable to provide uniformity in tuning shapes.

3 vs. 2 Resonant Circuit Transformers

Three-resonant-circuit transformers have inherently more gain than the two-circuit type, but are more complex. However, the use of capacitance tuning and proximity coupling simplifies the three-circuit type to a point where its construction and adjustment are relatively easy. The use of a self-tuned primary further reduces the adjustment difficulties of this type. Transformer coupling tolerances must be held somewhat closer than in the case of the two-circuit type. One variety of three-circuit transformer and its application is shown in Figure 1. The stage gain is 26 db for 4 Mc band width and a degenerative cathode resistor of 45 ohms. The stage gain for two resonant circuits of similar construction would be approximately 23 db, and the response would not be as flat out to the edge frequencies. This extra gain is very desirable, particularly in the multi-stage amplifiers of sensitive receivers. The construction is, however, somewhat more complex, and this fact should be weighed against the advantages of gain. Three-circuit transformer design may be assisted by an inspection of Figure 2 which illustrates the difficulties commonly encountered. M_1, Figure 1(a) largely dictates the band width, and M_2 and R dictate flatness of response. Curve (b), Figure 2, is a case of too small an M_2. Curve (a) is a case of correct value for M_2. Curve (c) is a case of too large an M_2; this is approaching the results of two tuned circuits. Curve (a) is then determined by the increase of M_2 to a point where the three peaks *just* disappear. R is changed to produce best average flatness for each case in turn.

If, after determining curve (a), the band width is not satisfactory, M_1 is readjusted in a direction indicated and (a) redetermined. Unfortunately for the experimenter, damping also has some effect on band width.

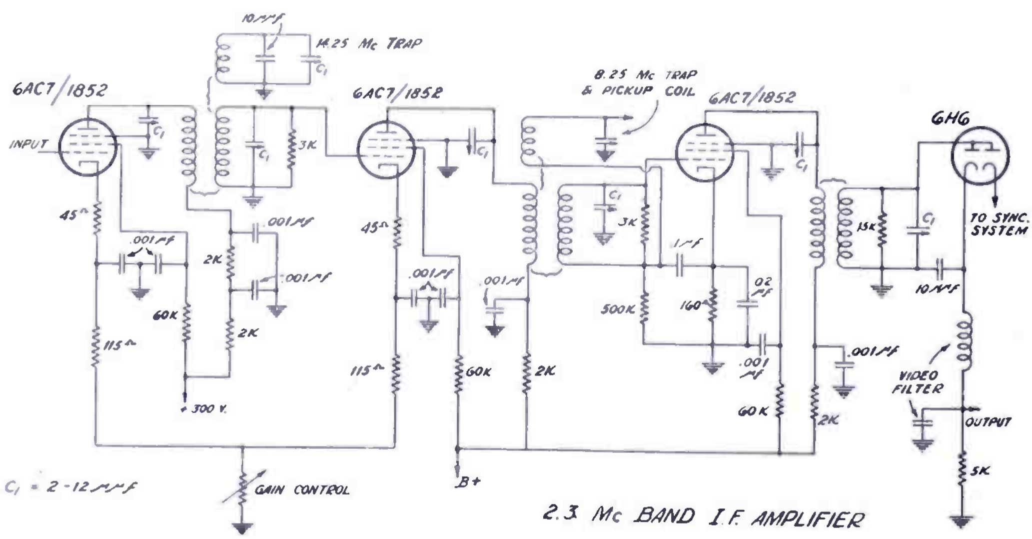

Fig. 3

Sound Carrier Attenuation

Sufficient attenuation must be provided for the accompanying sound channel so that it will not cause interference in the picture. When the desired signal is on the second, fifth or seventh television channel and there is a transmitter in the same locality on the first, fourth, or sixth channel respectively, attenuation must be provided for the adjacent sound frequency. When the picture i-f has the usual value of 12.75 Mc for the picture carrier, the attenuation frequencies become 8.25 Mc and 14.25 Mc.

8.25-Mc Traps

8.25 Mc traps serve the double function of reducing interference in the picture from the simultaneous sound broadcast carrier and of providing a point of pickup for the i-f amplifier.

The problem of adequate attenuation of 8.25 Mc in wide-band amplifiers is complex, since the traps are prone to have an effect on the edge of the desired frequency band. In general, a multiplicity of traps, each with limited 8.25 Mc reducing propensity will have less effect on, say, 8.75 Mc (edge of 4 Mc band) than one trap of overall equivalent 8.25 Mc rejection. Likewise, all traps should have approximately identical 8.25 Mc rejection to provide least attenuation of the desired band.

A proximity-coupled trap is quite effective and simple in construction, since it may be wound on the same form as the stage-tuning inductors. It is also a very excellent means of coupling to the sound i-f amplifier.

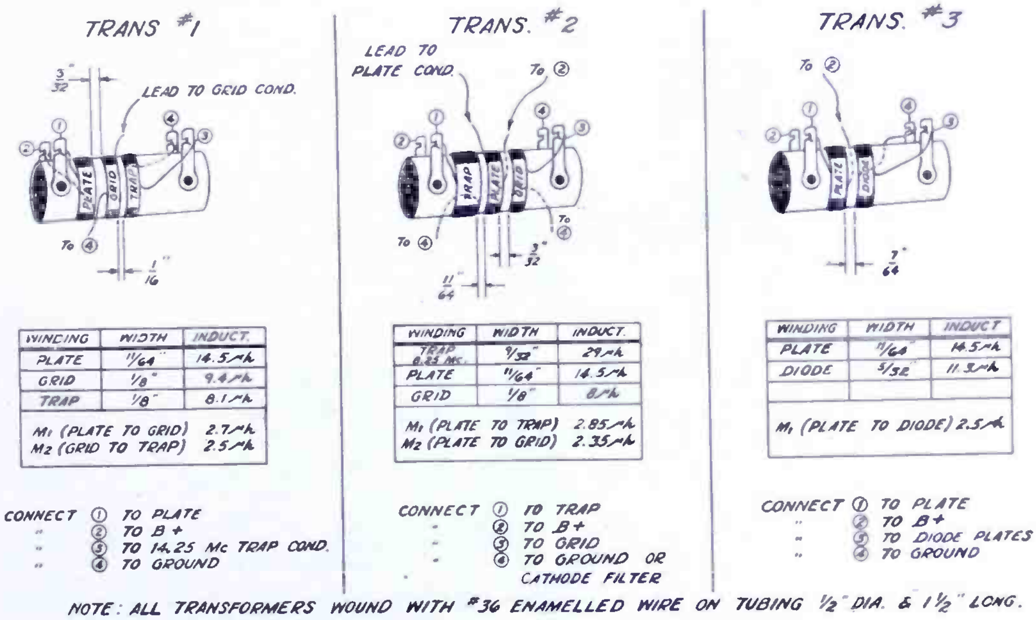

Fig. 4

Another type trap which may be equally effective is the use of cathode degeneration produced by a tuned-cathode circuit. One physical form of this type which has given good results, may be made by shunting a 0.005 μf capacitor with an inductor of two or three turns of flexible multi-strand wire wound on a form fastened to the capacitor body. The trap may be pre-adjusted and cement applied to fix the adjusted spacing between turns. Then the trap may be assembled in the receiver without the necessity of readjustment.

Cathode degeneration is most effective at normal bias on the tube to which it is applied. The trap ceases to function near plate-current cut-off which is quite detrimental, since the sound and picture-signal strengths are comparable. Hence, such an 8.25-Mc trap must be applied to a tube which is not gain controlled. Usually the last i-f amplifier (diode driver) is not on control, and in such cases the trap may be located in that stage.

The necessary minimum-response ratio of the flat portion of the picture transmission to the 8.25-Mc sound has not been fixed by commercial experience. In general, a ratio of 40 db is the smallest ratio to give satisfaction in experimental tests, and some receivers have required as much as 60 db for complete freedom from sound interference.

In narrow-band amplifiers, such as the 2.3-Mc type described below, the picture i-f stages are very selective to 8.25 Mc and traps are not always essential. Since some sound-pickup source is necessary it was found in this amplifier that a trap coupled to the plate winding of the second transformer gave an overall gain of 45 db for the sound i-f carrier. Coupling to the grid winding is not recommended because of the increased rejection of the extra cascaded circuit to the (in this case) desired 8.25-Mc signal.

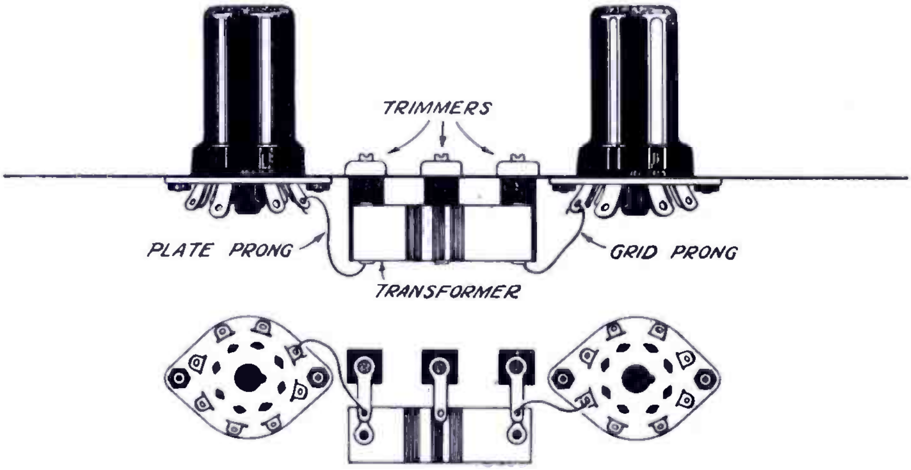

Fig. 5

14.25-Mc Traps

The sound carrier in the channel below the desired channel will heterodyne to produce a 14.25-Mc i-f. Since this carrier may under certain conditions be many times the intensity of the desired picture signal a greater degree of selectivity is necessary than is needed at 8.25 Mc. Again, no conclusive minimum figures have been determined, but 60 db is a selectivity ratio which many engineers are adopting as a first assumption.

Traps for this interference should be placed in the first part of the amplifier to reduce possibility of crosstalk. The same rules discussed for 8.25-Mc traps apply here except in the use of cathode degenerative traps. These may be used for tubes subjected to control bias since the traps are most needed when the desired signal is low, i.e., the traps functioning.

Some appreciable ratio should be maintained without cathode traps to take care of the situation of reception near two powerful signals on adjacent channels. In this case, the cathode traps would not function, but some selectivity would still be required.

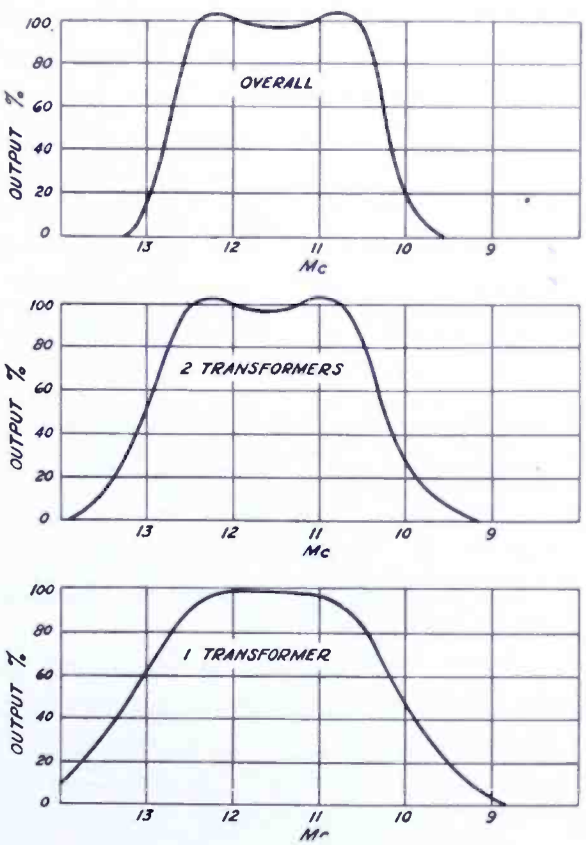

Fig. 6

Proximity traps are effective, and pre-set traps may be inserted in grid leads, etc. as a still further variety. The pre-set trap may take the physical form of the trap used for degeneration, i.e., a few turns of wire shunting a large fixed capacitor.

2.3-Mc Wide Amplifier

The 2.3-Mc amplifier shown schematically in Figure 3 has two stages of i.f. and a stage which may be operated as a converter by application of an oscillator voltage and use of suitable bias on the first tube. Measurements taken were with the first tube operating as an i-f amplifier. Gain at r.f. would be about 10 db less than measured i-f gain under proper conditions of converter operation.

The transformers used are all double tuned. The first one has a 14.25-Mc proximity-coupled trap; the second has an 8.25-Mc pickup coil coupled to the plate winding. The third is without a trap.

These coils are shown in constructional detail in Figure 4. The method of stage assembly is shown in Figure 5. Leads from plate and grid inductors are held as short as possible and the general compactness of the stage reduces any tendency for regeneration. One coil shield is necessary, and the third transformer is covered with a 2″ x 2″ square can.

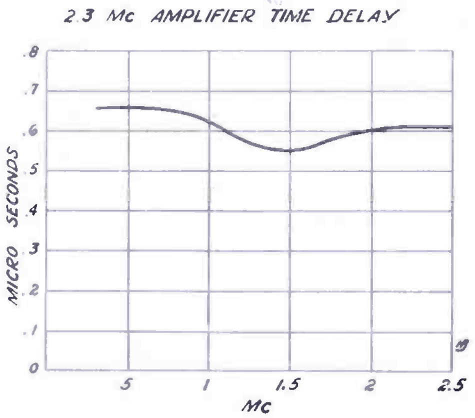

Fig. 7

Transmission shapes taken with inputs at the first, second, and third grids are shown in Figure 6.

Time delay is shown in Figure 7. A maximum variation of 0.1 microsecond is indicated. The amplifier was adjusted for flat amplitude characteristics. No subsequent adjustment was made to correct for variation in time delay.

Other performance characteristics are:

Gain (i.f.)	76 db
Expected gain (r.f.)	66 db
Individual stage gains (i.f.)	
(a) first tube	25 db
(b) second tube	25.3 db
(c) third tube	25.7 db
Attenuation at 8.25 Mc	58 db
Effect of trap alone	6.5 db
Attenuation at 14.25 Mc	47 db
Effect of trap alone	21 db

A second 14.25-Mc trap may be used if considered necessary.

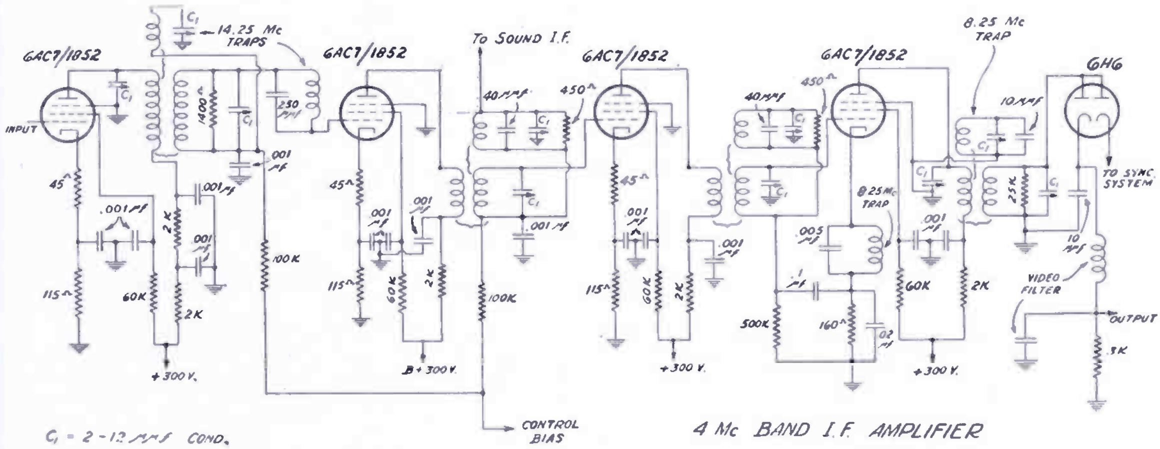

Fig. 8

The 8.25-Mc trap is not essential in rejecting the sound from the picture as the normal selectivity of the tuned circuits is adequate. It serves the purpose of "take off" point for the sound i-f amplifier. A gain of 45 db for sound is thus obtained in the picture amplifier, which considering the first tube as a converter would give a sound carrier gain of about 35 db. One i-f stage in the sound i-f amplifier would, therefore, produce a sensitivity comparable to picture sensitivity.

The selectivity at the 8.25-Mc transmission is 200 kc band width for 6 db down. All other frequencies are attenuated more than 6 db.

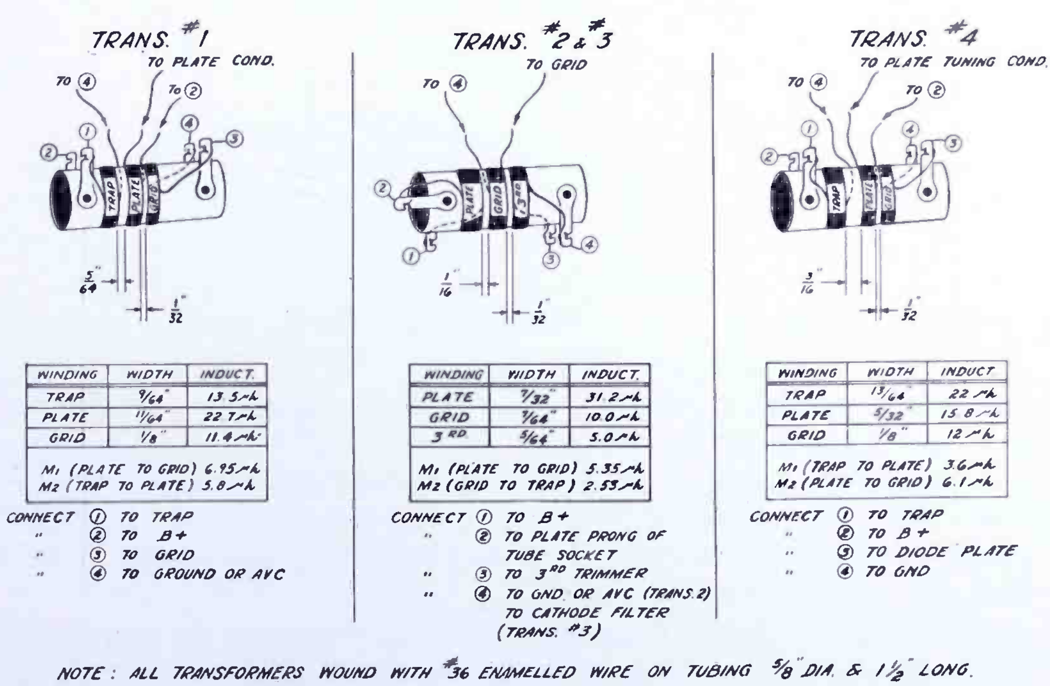

Fig. 9

Self-tuned primaries will produce considerably more gain (about 2.5 db per stage) but small commercial variation in primary tuned frequency should be assured before their use is recommended. Such variations are more detrimental to narrow-band amplifiers than to wide. Likewise, 3-resonant-circuit transformers will add gain (and some complexity) to the amplifier.

4-Mc Wide Amplifier

The 4-Mc wide amplifier shown schematically in Figure 8 has 3 i-f stages and a converter stage which was, as in the case of the

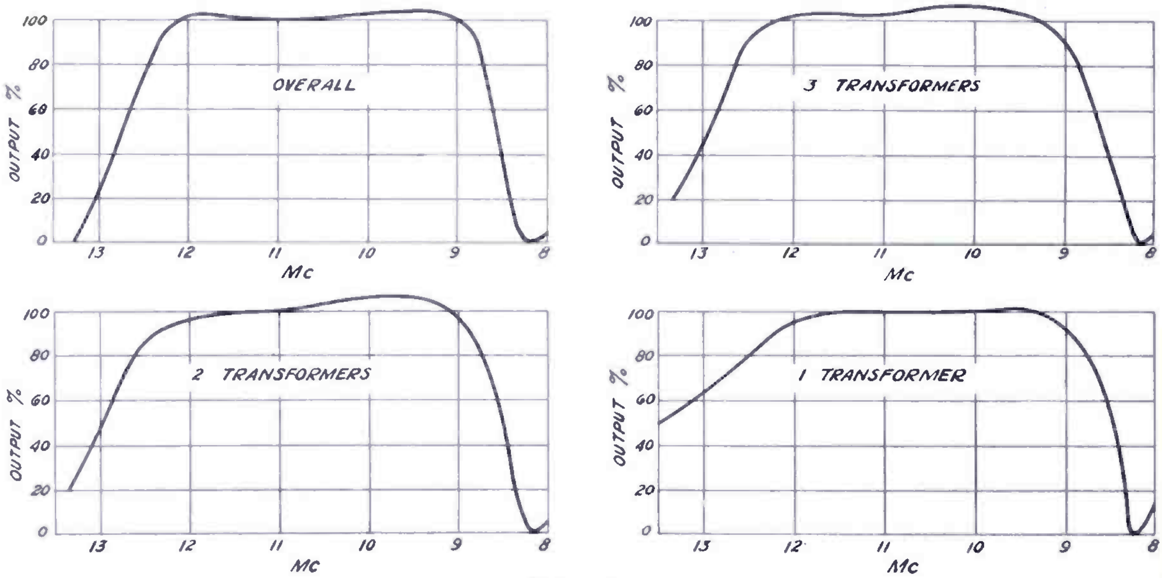

Fig. 10

2.3-Mc amplifier, operated as an i-f amplifier. Gains with r-f inputs and the first tube biased and properly supplied with oscillator voltage should be approximately 10 db less than the i-f gains.

Two traps are used to reject 14.25 Mc. Both are in the first circuits to prevent overload of succeeding grids by a strong adjacent-channel sound signal. One trap is proximity coupled and the other is in series with the second grid lead. It consists of a 250-$\mu\mu$f capacitor shunted by a small tuning inductor. This type may be pre-set before wiring into the main chassis.

The second and third transformers are of the three-resonant-circuit type. Self-resonant primaries are used and the adjustments are thus reduced to two per transformer. Construction data on all transformers and traps used is given in Figure 9.

Two 8.25 Mc traps are placed near the output of the amplifier. These provide 40 db attenuation, which in several amplifiers built from these specifications gave freedom from sound-in-picture interference.

It is generally desired, and certainly safer, to have about 50 db attenuation at 8.25 Mc. A third trap coupled to the primary of the second transformer will provide the additional 10 db attenuation. This trap may be used as the take off for the sound i-f amplifier instead of the tertiary winding of the second transformer as indicated in Figure 8. The gain at this first tertiary for 8.25 Mc is 28 db. One sound i-f stage following this circuit will provide an overall sound receiver sensitivity in the order of 100 μv. The use of the extra trap will materially improve this sensitivity.

The transmission shapes depicting the selectivity from succeeding grids are shown in Figure 10. Other performance characteristics are:

Gain (i.f.)	88 db
Expected gain (r.f.)	78 db
Individual stage gains (i.f.)	
(a) first tube	21.4 db
(b) second tube	25.7 db
(c) third tube	25.7 db
(d) fourth tube	15.2 db
Attenuation at 8.25 Mc	40 db
Attenuation at 14.25 Mc	73 db
Attenuation of grid trap alone	24 db
Attenuation of proximity trap alone	22 db

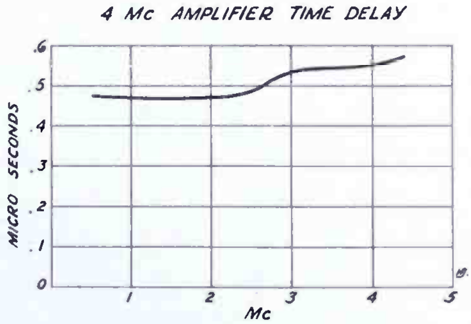

Fig. 11

Overall time delay characteristics are shown in Figure 11. A maximum variation of 0.09 microsecond is indicated. The amplifier was aligned for flat amplitude characteristics and then tested for time delay, no subsequent adjustment being made to correct for time delay.

RECENT ADVANCES IN BARIUM GETTER TECHNIQUE

By

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Summary—Barium metal of high purity and without undesirable by-products can be obtained by chemically reducing barium berylliate. Barium berylliate, the barium salt of the hypothetical dibasic beryllic acid, has been discovered only recently. It is stable in air, but can be easily reduced by tantalum or similar reducing agents at about 1300°C with approximately 60 per cent efficiency. The resulting barium metal is of silvery appearance and is very active in the clean-up of gas. The preparation of barium berylliate and methods for testing it are briefly outlined. The design, manufacture, and use of a simple, inexpensive, but highly efficient getter for the controlled production of barium metal from barium berylliate are described.

1. Introduction

IN a previous paper[1] a simple reaction-type getter (termed "batalum getter") was described. It consists of a core of tantalum, preferably in form of a wire or coil, coated with the carbonates of strontium and barium. After the carbonates are decomposed in vacuum by heating the core electrically to 1000°C approximately, free alkaline-earth metal is produced when the temperature is raised to about 1300°C. While this form of batalum getter has been used with satisfactory results in the manufacture of millions of tubes, it has been replaced by a new form employing a coating of barium berylliate which gives a yield of pure useful barium at a lower cost.

To digress for a moment, let us consider the preferred coating for a batalum getter. If it were possible to use it, the coating would be pure barium oxide since no decomposition of the coating would be required. It is well known, however, that barium oxide is unstable under atmospheric conditions and combines rapidly with the moisture and carbon dioxide of the air. Attempts to "dead burn" barium or strontium oxide have failed even at temperatures of the electric arc between iron electrodes.[2] Many barium compounds like barium zirco-

[1] "Batalum", a Barium Getter for Metal Tubes, by E. A. Lederer and D. H. Wamsley, RCA Review, July, 1937 (U. S. Patent No. 2,130,190).

[2] This investigation was carried out by Mr. D. H. Wamsley, RCA Mfg. Co., Inc., Harrison, N. J.

nate, barium titanate, barium tungstate, barium tantalate and others have been tested for production of barium by reduction with tantalum and zirconium, but if barium was evolved at all the yield was so small as to be worthless for the desired purpose.

2. Barium Berylliate

It occurred to the writer that barium oxide, being strongly alkaline, should form a chemical compound with beryllium oxide which is known to have amphoteric properties. The resultant compound should be the barium salt of the very weak hypothetical dibasic beryllic acid, reducible by tantalum, zirconium, or their equivalents. Tests were made and it was found that the compound can be produced by dissolving beryllium oxide in molten barium hydroxide at about 600°C. The reaction was barely discernible and the mass solidified. The resulting compound was termed barium berylliate. Insofar as the writer is aware, it is a new compound[3,4] and has not been described in the literature prior to its discovery by him. It is stable in dry air, dissolves slowly in concentrated nitric acid, and hydrolyzes slowly in contact with water, forming barium and beryllium hydroxides. Barium berylliate is reduced by tantalum metal in vacuum at about 1300°C. Owing to its high vapor pressure at this temperature, the barium metal evaporates from the zone of reaction and deposits as a mirror on the cool walls of the envelope. Titanium, zirconium, columbium, aluminum, tungsten, molybdenum, and the rare-earth metals also reduce barium berylliate. It is interesting that the beryllium oxide is not perceptibly reduced by the tantalum below 2000°C, but this reaction proceeds rapidly as the temperature approaches the melting point of the tantalum.

Evidently beryllium oxide and barium berylliate form solid solutions and some of the properties of the resulting product are a function of the beryllium oxide content. If barium hydroxide and beryllium oxide react in the proportion to satisfy the equation

$$Ba(OH)_2 + BeO = BaBeO_2 + H_2O$$

the product melts at about 1450°C. It has a density of 4.5 and when

[3] Its preparation and use are described in U. S. patents No. 2,173,258 and No. 2,173,259.

[4] The writer is indebted to Dr. J. M. Cage of the General Electric Company, Schenectady, N. Y., who identified barium berylliate as a compound by means of X-ray spectrograms. The writer also wishes to thank Dr. M. Benjamin of the British General Electric Company in Wembly, England, and Dr. H. Bienfait of the Philips Company in Holland for X-ray spectrograms and other valuable observations.

cast into a chilled mold solidifies to an opaque mass. When pulverized it is easily attacked by moisture. If, however, the amount of beryllium oxide is increased beyond the ratio given in the formula, the melting point and the stability toward moisture are increased.

A convenient method of producing barium berylliate consists in mixing barium carbonate and beryllium oxide in the desired proportions and in firing the mixture in hydrogen at about 1065°C. In order to increase the density and stability of the product, refiring in air at 1250° to 1275°C is advisable. The same method is applicable to the preparation of calcium or strontium berylliate. The tests reported in this paper were made with barium berylliate prepared from barium carbonate and with a ratio of $BaO:BeO = 1:2.4$.

To determine its stability in air, about one gram of freshly prepared barium berylliate was evenly spread on a watch glass and exposed to air in the laboratory. The sample was protected from dust and was weighed at intervals as shown in Table 1.

Table 1

Weight of Sample in Grams	Average Room Temperature	Average Relative Humidity	Exposure Time Hours	Weight Change %
0.9134	85°F	58%	0	
0.9142	82	54	24	+ 0.087
0.9141	82	46	48	— 0.001
0.9142	83	49	72	+ 0.001
0.9146	85	65	96	+ 0.004
0.9150	82	71	168	+ 0.004
0.9145	After two hours in desiccator			— 0.004

As seen from the table, the largest increase in weight occurred in the first 24 hours, and is probably due to physical adsorption of moisture. This assumption is supported by the decrease in weight observed at 48 hours, when the relative humidity dropped below 50 per cent, and after insertion of the sample in the desiccator.

It was next in order to determine the gas content and barium yield of barium berylliate to find out if it would be suitable for use as a practical getter.

3. Determination of Gas Content and of Barium Yield

In a good getter the initial gas content should be low, the occluded gas should be released far below the flashing temperature, no gas should be evolved during the evolution of the barium metal, and the ratio of usable amount of barium metal (yield) to actual amount of barium in the coating should be high. For the study of gas content and barium yield, it is convenient to coat a tantalum wire with a suspension of barium beryllate in a nitrocellulose binder and to mount a section of such coated filament between lead-in wires in a glass bulb. The bulb is sealed onto a vacuum system together with a pressure gauge. The coated filament can be heated to the desired temperature by an electric current. The envelope is degassed in the conventional manner by baking in an oven. When the filament is heated slowly, the peak of gas evolution occurs at about 900°C. The gas evolved is made up of that occluded in the porous coating, contained in the tantalum

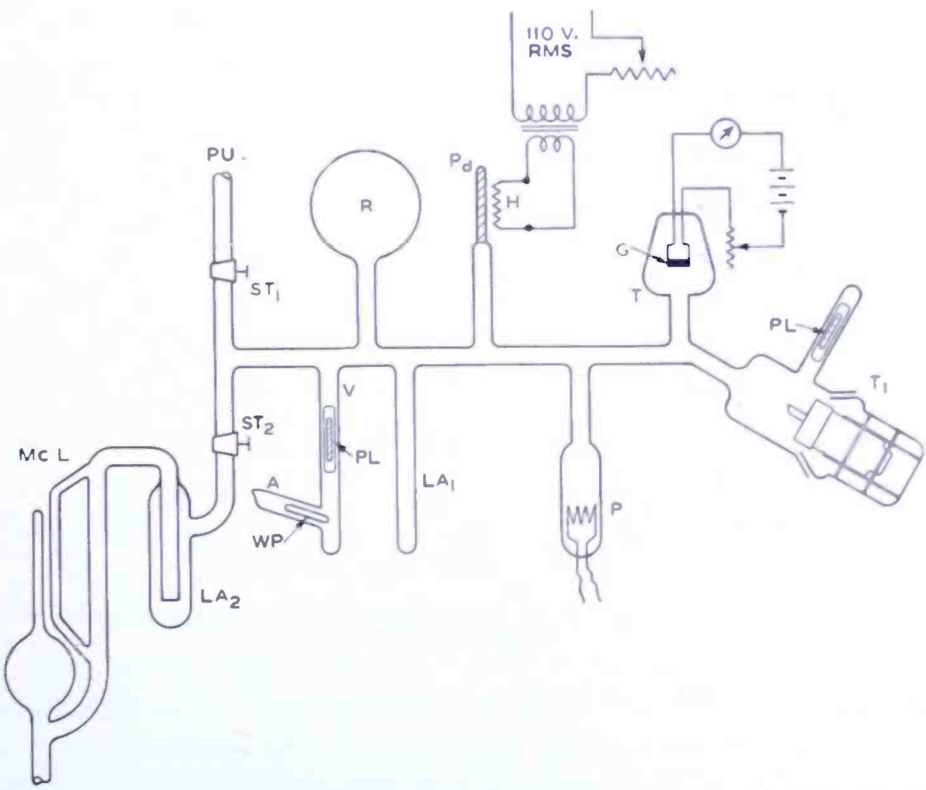

Fig. 1—Vacuum apparatus for measuring gas content and alkaline-earth metal yield.

core, and resulting from the decomposition of the residue from the nitrocellulose binder. In this manner many coated filaments and getter assemblies were studied. The ratio of weight of coating to weight of tantalum usually ranged from 0.5 to 0.3. The average gas content was quite uniform and varied between 2 and 6 liter-microns ($l\mu$) per milligram of coating, or about 4 per cent of that of a barium-strontium carbonate coating of equal weight. This fact is important since the getter when flashed is capable of cleaning up its own occluded gas. The low gas content of the coating has the further advantage that the reducing material is not attacked prematurely by gas.

Increasing the temperature of the assembly to about 1300°C, one observes the evolution of barium metal which manifests itself by a mirror-like deposit on the cool glass walls of the vessel. The barium deposit is almost as white and silvery as a magnesium deposit. The color of the deposit is believed to be an indication of its purity and the absence of gas during the flash.[5]

Besides the purity of the barium metal, the next important question is that of barium yield. Measurements were carried out using the water vapor-hydrogen pressure method first proposed by T. P. Beredenikowa.[6] This method which makes use of the reaction

$$Ba + 2\ H_2O = Ba(OH)_2 + H_2$$

permits measurement of free alkaline-earth metal by means of the

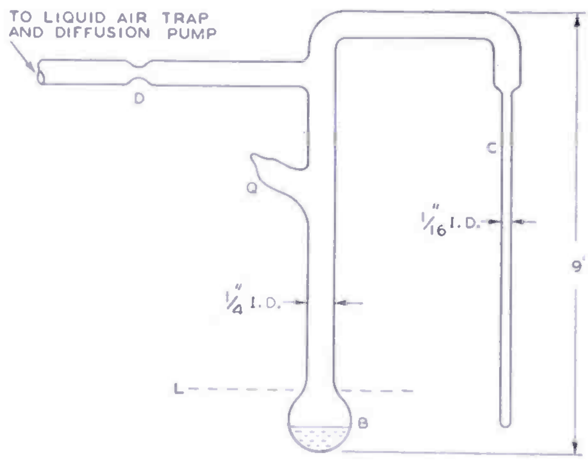

Fig. 2—Diagram showing arrangement of apparatus for making "water pills."

pressure of the liberated hydrogen in a known volume. Its accuracy far surpasses[7] measurements by titration or by gravimetric means. The apparatus modified for convenient and rapid gas and yield determinations is shown schematically in Figure 1 and may deserve description.

T is the tube containing the barium source, the yield of which we wish to determine. *P* is a Pirani gauge, *R* is a reservoir (to adjust volume of system to pressure range of McLeod gauge), LA_1 and LA_2 are liquid-air traps, *McL* is a McLeod gauge, *PU* leads to a mercury diffusion pump, and *V* is the device for introducing the water vapor

[5] Pure barium metal produced in this manner has been used by Dr. M. Benjamin (*Phil. Mag.* 26, 1049-62, 1938) in a study of the migration of barium on nickel and tungsten.

[6] T. P. Beredenikowa, *Phys. Z. Sowjet.* 2, 77, 1932

[7] The writer used the method in 1936 to determine the amount of free barium on vapor-process cathodes and found it well in agreement with the monomolecular layer as suggested by theory.

into the system. Using device *V* reduces the number of stop cocks on the vacuum system and has been so useful that a short description is justified. It consists of a vertical tube *V* with an appendix *A* slanting downward into the lower end of tube *V*. Tube *V* contains a glass-enclosed iron plunger, *PL*, and *A* contains a "water pill", *WP*.

Water pills are made using an apparatus constructed with heavy-wall Nonex glass, as shown in Figure 2. This apparatus is connected through a liquid air trap to a mercury diffusion pump. With *Q* open, approximately 1 cc. of double distilled, boiling water is poured into *B*, then *Q* is tipped off, and connection to the pump is established. Soon the water in *B* becomes so cold that "boiling" stops. *B* is now immersed in liquid air up to level *L* and while the vacuum is maintained, the

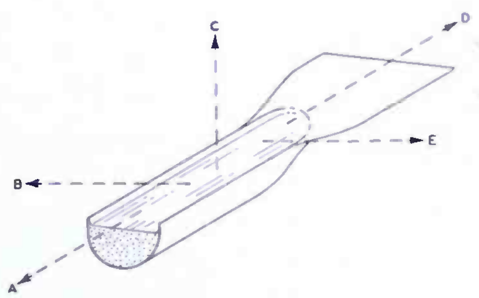

Fig. 3—Portion of getter unit formed in shape of a trough from tantalum ribbon and filled with barium berylliate. Arrows indicate direction of barium-metal evaporation.

entire system is torched out for about 10 minutes. After cooling to room temperature, *B* is warmed up and about ⅓ of the amount of water is allowed to evaporate. The assembly is now sealed off at *D*. Holding *B* in the left hand, thereby warming up the water, *C* is now immersed in liquid air. After about 30 seconds enough water has condensed in *C*. The amount of water is checked by defrosting. When the height of the water column in *C* is approximately ¾″, a water pill of about 1″ length is sealed off and the process is repeated for the next water pill.

The gas-content and barium-yield determination is carried out as follows: In Figure 1, the tube *T*, containing a getter or filament (*G*) of known coating weight, is baked in the conventional manner while the part of the McLeod gauge as shown is baked in a separate oven and the remainder of the system is torched out well. The gas content determination is done with ST_1 closed and by slowly degassing the filament up to a maximum temperature of about 900°C. Knowing the volume of the entire system and the pressure as read either with the Pirani gauge or preferably with the McLeod gauge, it is easy to compute the gas content in litermicrons per milligram of coating weight.

Next, the system is re-exhausted and after attaining a "sticking vacuum" with the McLeod gauge, ST_2 is closed and the barium is flashed in tube T. With ST_1 and ST_2 closed, the water pill is broken by means of the magnetic plunger. After the reaction, the system contains hydrogen and an excess of water vapor. Immersing LA_1 in liquid air soon removes the excess of water vapor and after cooling LA_2 in liquid air, ST_2 can be opened; the pressure is then read on the McLeod gauge and is recorded after constancy is attained. A test for hydrogen only consists of removing the gas from the system by heating the palladium tube Pd with heater H. After constancy is attained, the residual pressure is recorded: the difference in pressure readings is due to pure hydrogen as produced by the reaction between barium and water vapor. It is then possible to compute the amount of barium which produced the measured amount of hydrogen.

In order to test for the amount of free barium getter in a finished metal tube, a container T_1 with a ground-glass stopper is used. The base is removed from the metal tube which is then washed with methanol and inserted in a holder located in the container T_1. After the container is exhausted at room temperature for a few hours, the glass tip of the metal tube is broken off with a plunger and the measurement is carried out as described above. Some typical gas-content and barium-yield measurements are recorded in Table 2.

Table 2

Sample	Coating Weight *mg*	Total Gas Content $l\mu$	H_2 Pressure After Reaction μ	Residue After Diffusing Out H_2 μ	H_2 $l\mu$	Barium Yield %
A	1.26	—	61.5	1.77	59.73	58.0
B	1.38	—	71.3	1.58	71.3	62.3
C	4.1	9.05	75.0	0.81	208.0	57.0

The average yield obtained with barium berylliate coated on tantalum is approximately 60 per cent of the barium metal contained in the coating. Compared with a barium-strontium carbonate coating as described previously,[1] the increase in yield is about 50 per cent. Since barium berylliate is a better and more efficient coating and its density

[1] Loc. Cit.

is higher than that of the presintered double carbonates, it is particularly suited as a coating for the batalum-ribbon type of getter which is described in the next section.

4. Use in Commercial Tubes

If a tantalum ribbon is formed into a trough with semicircular cross-section as shown in Figure 3, and if the trough is filled with barium berylliate, a self-shielding, barium-getter source is obtained. The tantalum carrier, the middle section of which is formed to a trough or channel, has flat ends to facilitate welding to leads. It can

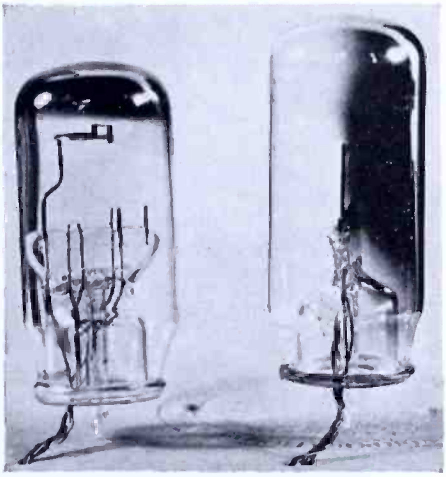

Fig. 4—Showing barium-metal deposit as affected by self-shielding property of ribbon getter for different positions of the trough.

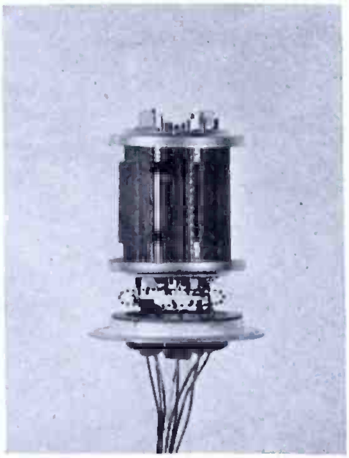

Fig. 5—Mount of metal tube with getter which can be distinguished within dotted area.

be heated electrically. The energy required for flashing the barium metal is only a few watts. The barium metal evaporating from the coating in the trough is confined to the hemispherical region identified by *A B C D E* above the trough. The tantalum trough acts as a shield to prevent the deposition of any barium metal below the trough. This self-shielding property is further shown by the photograph in Figure 4. The getter of the tube on the right was flashed by heating it conductively through external leads while the getter in the tube on the left was flashed by heating it inductively. The directional effect can be judged from the extent of the barium deposit in each case.

Since the barium deposit can be directed and controlled with this form of getter, leakage resulting from getter deposits between the tube electrodes can be eliminated. The residue in the channel, after the flash is completed, adheres well to the tantalum and does not cause loose particles in the tube.

Figure 5 shows the ribbon getter and its position in a metal tube. For this particular application, the getter is formed from a 0.040" x 0.001" tantalum ribbon, and holds approximately 2.5 mg of barium beryllíate. The total weight of the assembly is approximately 7.5 mg and its length is ¾".

The ribbon getter is welded between the shell pin lead (ground) and the shell. The getter can be heated conveniently on the exhaust machine, or it can be flashed on the aging rack after the tube is

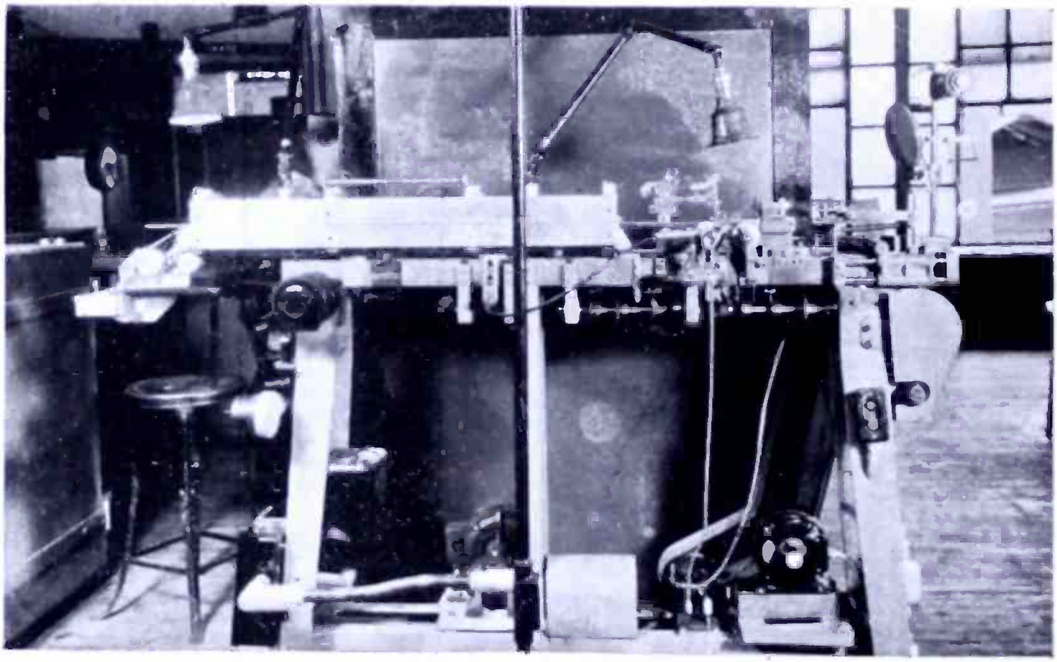

Fig. 6—Automatic "batalum getter machine."

based. Because of the simplicity of this getter, forming, filling of the channels, and cutting can be accomplished automatically on a machine with a minimum loss of material. Figure 6 shows an automatic "batalum getter machine" which is capable of producing 2500 getter units per hour.

Batalum ribbon getter has been used for over a year in the manufacture of millions of metal tubes with entirely satisfactory results. In a modified form, adapted for flashing by induction heating, it is also used for other high-vacuum devices.

The writer gratefully acknowledges helpful advice and assistance received from Dr. G. R. Shaw, D. H. Wamsley, F. Roth, and his other colleagues.

I-F SELECTIVITY IN RECEIVERS FOR COMMERCIAL RADIO SERVICES

By

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Summary—A composite band-pass filter, containing both "M-derived" and "tuned-transformer" sections, is described. Frequency characteristics of commercial-communications type radio receiving equipment employing such i-f filters are given.

SELECTIVITY characteristics of i-f units, employed in the RCA diversity receiving equipment of previous years, have been published in a former paper.[1] More recent design and development work on the problem has resulted in a considerable improvement in

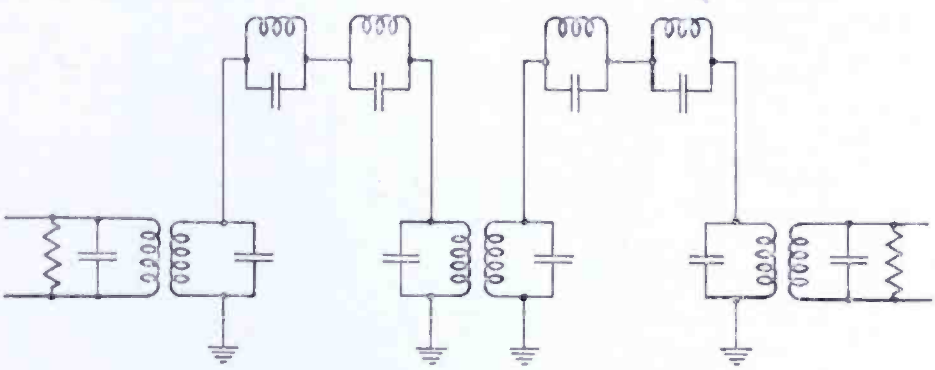

Fig. 1—Designed filter circuit.

the selectivity and flexibility obtainable in a given volume of space. It is the purpose of this present article to describe the chief contribution to this improvement, and the overall i-f selectivity actually obtained.

Past practice has been to use a number of tuned-transformer-coupled stages in cascade to provide the desired selectivity in the last i-f system. As the number of such stages—utilizing a single, tuned transformer each—is increased, the relative improvement per stage is not very great. While the use of five such stages is reasonable, from general considerations, the use of ten would hardly be justified. Furthermore, the use of so great a number of tuned transformers would provide far more attenuation than is required at frequencies remote

[1] Recent Developments in Diversity Receiving Equipment. RCA Review for July 1937. J. B. Moore.

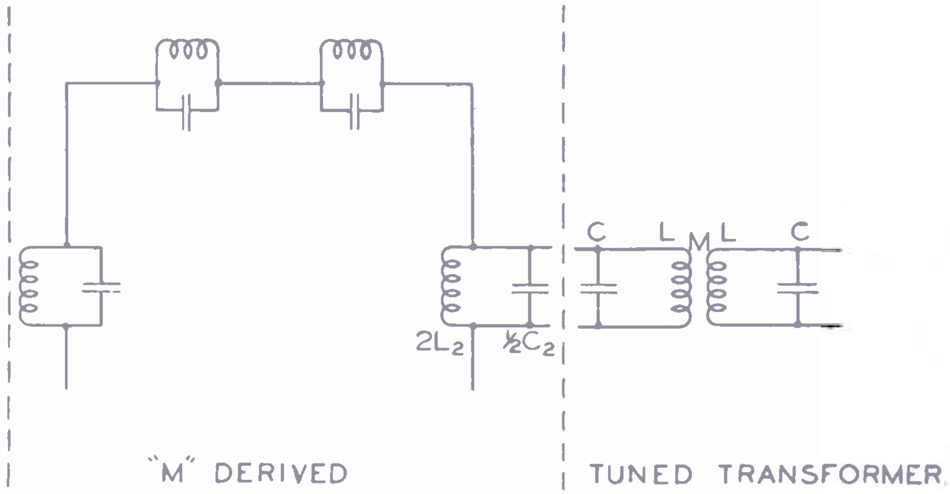

Fig. 2—Combined "M" derived and tuned transformer sections.

from the pass band. The problem is, then, to design a system which will give the maximum steepness of cut-off, and the desired attenuation at all frequencies outside the pass band, with a minimum requirement of circuit elements and physical volume.

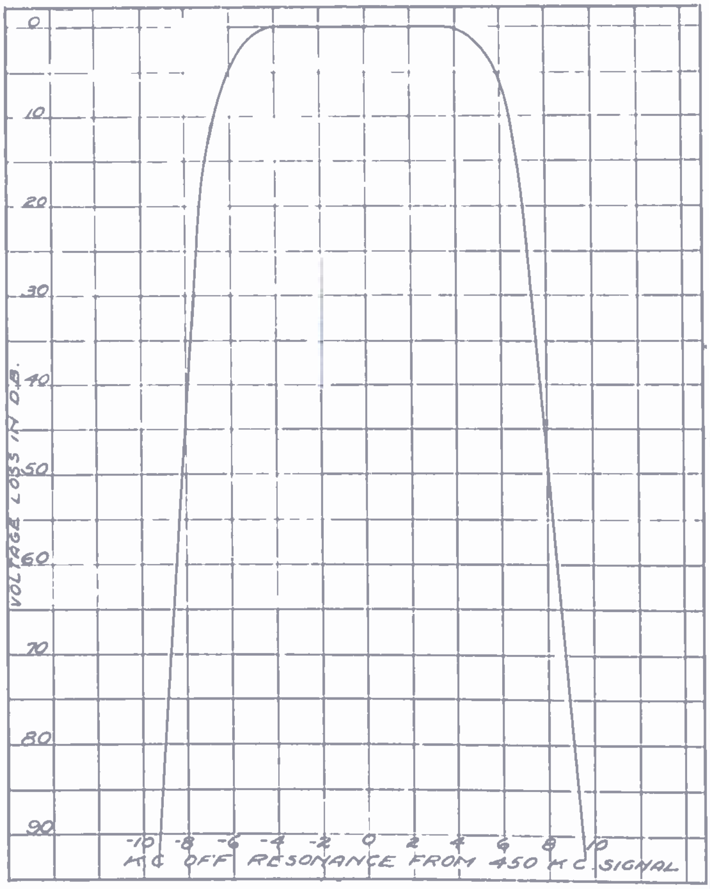

Fig. 3—Typical overall frequency characteristic 10-kc band width.

The problem is obviously best solved by the use of some type of composite band-pass filter. The filter circuit chosen for this application is shown in Figure 1. This is built up by combining two "M-derived" six-element sections with three "tuned-transformer" sections. Figure 2 shows the manner in which these two types of section are combined, in order to reduce the total number of circuit elements (coils and condensers) required.

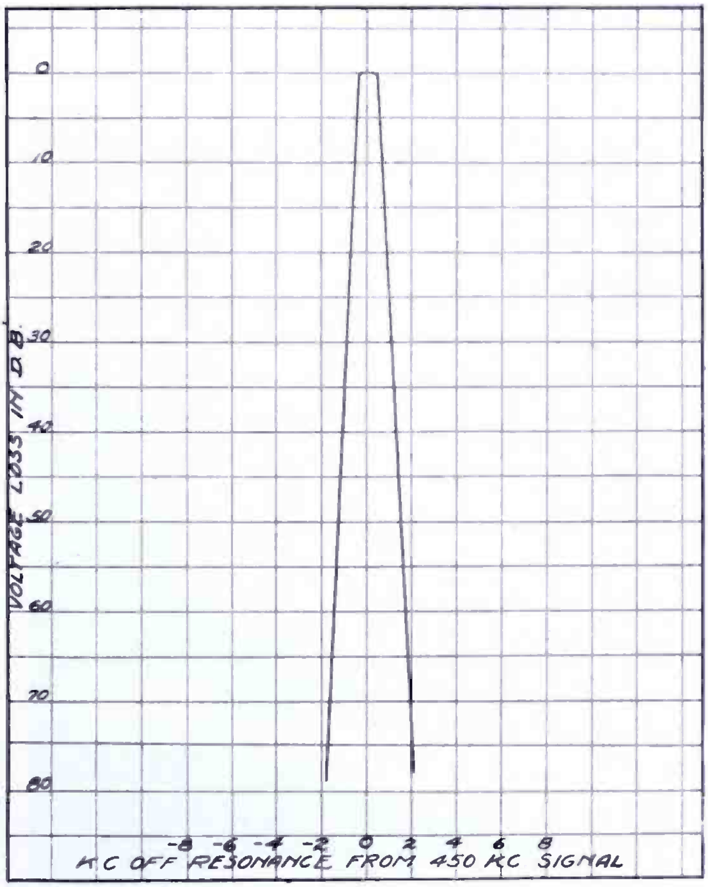

Fig. 4—Overall frequency characteristic 1-kc band width.

Design data for the six-element sections can be found in the literature on the subject. Using standard nomenclature, the design equations for the tuned transformers are:—

$$L = \frac{(f_2 - f_1)R}{4\pi f_1^2}\left[1 + \left(\frac{f_1}{f_2}\right)^2\right] \quad \text{Henrys}$$

$$C = \frac{1}{2\pi(f_2 - f_1)R} \quad \text{Farads}$$

$$M = \frac{(f_2 - f_1) R}{4\pi f_1^2} \left[1 - \left(\frac{f_1}{f_2} \right)^2 \right] \qquad \text{Henrys}$$

$$k = \frac{1 - \left(\dfrac{f_1}{f_2} \right)^2}{1 + \left(\dfrac{f_1}{f_2} \right)^2} \qquad \text{Coefficient of coupling}$$

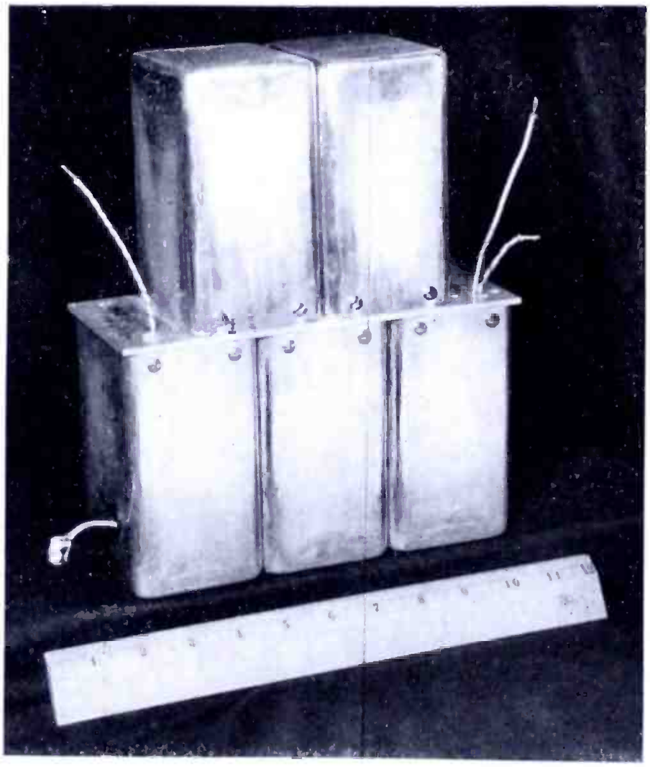

Fig. 5—The completed filter unit.

In the final mechanical assembly, each pair of coupled tuned circuits of Figure 1 is housed in a separate shielding can, rectangular in shape; and the two tuned circuits in each series leg are similarly housed in a single shielding can. The total volume of the five cans, for a filter having a mid-band frequency of 50 kc, is approximately 243 cubic inches. A photograph of a completed filter unit is shown in Figure 5.

Filter band widths in use at the present time are 1 kc, 2 kc, 4 kc, 6 kc, and 10 kc. Typical overall frequency characteristics of the entire i-f unit—including selectivity, the 450-kc first i-f system, the 50-kc band-pass filter in question, and the 50-kc i-f amplifier—are given in

Figures 3 and 4. The band-pass filter is what provides the steep cut-off, and the major part of the selectivity. Only in the case of the 10-kc band width does the selectivity of the first i-f system and of the 50-kc amplifier contribute appreciably to the overall selectivity. Comparison of these characteristics with ones of previous equipment show a decided improvement in steepness of cut-off; and retention of ample protection at frequencies remote from the pass band. Attainment and maintenance of such frequency characteristics requires that coils and condensers have low losses, and that they do not change appreciably with either atmospheric conditions or time. Low coil losses are realized by the use of "universal" wound coils of litz wire, provided with suitable cores of magnetite. Fixed capacitances are a mica dielectric type which have a low temperature coefficient of capacitance and which are not affected by high humidity of the ambient air.

It is believed that this design of filter provides the maximum efficiency of space, for given performance, that can be obtained by the use of commercially producible coils and condensers now generally available. The use of quartz crystals or resistance-compensated coils (requiring associated vacuum tubes and circuits) will give considerably steeper cut-off than that obtainable from the filter described. It is believed, however, that the composite filter described provides performance satisfactory for the great majority of present-day applications; and that it provides such performance in a space, and at a cost, that are entirely justifiable.

SUPERHETERODYNE CONVERTER SYSTEM CONSIDERATIONS IN TELEVISION RECEIVERS*

By

E. W. Herold

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Summary—There are three methods of operation of tubes for use in the converter stage of superheterodyne receivers. In the first, signal and oscillator voltages are impressed on the same electrode of the tube, while in the other two, signal and oscillator voltages are placed on different electrodes. The latter two methods are differentiated by whether the oscillator electrode precedes or follows the signal electrode along the direction of electron flow. By the use of four fictitious tubes, a triode and a pentode using the first method of operation, a hexode for outer-grid oscillator injection, and a hexode for inner-grid oscillator injection, the three methods are compared.

For all converters or mixers, the conversion transconductance is

$$g_c = \frac{1}{2\pi} \int_0^{2\pi} g_m \sin \omega t \, d\,(\omega t)$$

where g_m is the signal-electrode to output-electrode transconductance and ω is the angular frequency of the oscillator. The mean-squared plate noise current is

$$\overline{i_{i\text{-}f}^2} = \frac{1}{2\pi} \int_0^{2\pi} \overline{i_{pn}^2} \, d\,(\omega t)$$

where $\overline{i_{pn}^2}$ is the mean-squared plate noise current of the same tube under amplifier conditions. By using the formulas for $\overline{i_{pn}^2}$ developed by Thompson and North and applicable to the first three of the fictitious tubes and an empirical relation for the noise of the fourth, quantitative relations for the conversion transconductance and equivalent grid noise resistance are developed.

By assuming a similar cathode and first grid structure, the triode, pentode, and outer-grid-oscillator-injection hexode may be compared quantitatively for use in a television converter stage. It is found that the triode with oscillator and signal voltages on the control grid has greatest gain and lowest noise while the pentode is only slightly poorer in this respect. However, the triode has the disadvantage that feedback must be taken into account. The hexode with oscillator on an outer grid is at a disadvantage with respect to both gain and noise and does not appear to be suitable as a first tube in a high-sensitivity television receiver. The hexode with oscillator on an inner grid, although it can be compared only qualitatively, is judged to be not much superior in gain or noise to the other hexode. It is advantageous chiefly in the ease with which a local oscillator may be combined in the same mount.

In tubes such as the triode and pentode in which oscillator and signal are applied to the same grid, the use of automatic bias such as obtained by

* This paper was presented at the N. Y. Convention of the Institute of Radio Engineers on Sept. 23, 1939 under the title, "Superheterodyne First Detector Considerations in Television Receivers".

a high-resistance gridleak minimizes variations of gain with oscillator voltage. Practical operating data at television frequencies are given for some commercially available tubes including two pentagrids designed for broadcast use. The latter are very poor in conversion transconductance and noise as compared with the high-transconductance type 6AC7/1852 pentode. The equivalent noise resistance of the type 6AC7/1852 used in the converter or first-detector stage is about as low as that of the other high-transconductance type, the 6AB7/1853, used as amplifier. No marked improvement in signal-to-noise ratio can be expected, therefore, by adding a type 6AB7/1853 r-f stage ahead of an 6AC7/1852 first detector in a television receiver.

INTRODUCTION

DURING recent years the communication field has seen considerable improvement in the design and performance of frequency converters for superheterodyne receivers. In many ways, however, these improvements have only followed those of amplifier tubes. The reason for this is that, basically, the principle of frequency conversion has remained unchanged: by taking an amplifier and varying its amplification periodically at the oscillator frequency we obtain the converter. Thus the better the amplifier, the better the converter, and this has been the story throughout the years. Unfortunately, the converter always appears to be behind the amplifier, because, as is indicated by simple theory, the gain of a tube used as a converter cannot exceed about one-third the gain of the same tube used as amplifier. As regards fluctuation noise, some converters are at a still greater disadvantage over amplifier tubes because they include a large number of current-drawing electrodes.

With the introduction of commercial television, many tube problems arose and, among them, not the least important was that of the converter, or first detector stage. Although an r-f stage may be used in some designs, many receivers omit this stage. It is with the latter type of receiver that the converter problem is most acute. Fluctuation noise, for example, sets a limit to the sensitivity of a receiver and, when the noise is expressed in terms of an equivalent single-frequency voltage, is very high in the wide-band picture channel. It will be seen, however, that high-transconductance amplifier tubes used in the converter stage provide a reasonably satisfactory solution to the noise problem.

The converter stage of a superheterodyne receiver comprises two parts: the modulator which produces the intermediate frequency, and the local oscillator. In a television receiver, the oscillator must operate stably and with little frequency shift up to 120 Mc. Such qualities are most readily obtained when a separate tube is used for the oscillator. This paper will be mainly concerned, therefore, with the modulator section of the converter stage, the oscillator section being treated only when it bears directly on the modulator problem.

General Characteristics of Converter Operation

The process of frequency conversion is essentially one of low-percentage modulation of the local oscillator frequency by the signal frequency. One of the sidebands of the modulation process is utilized as the intermediate frequency. Such low-percentage modulation may be carried out in any one of three ways. In the first, signal and oscillator are impressed on the same electrode of the tube and modulation occurs by virtue of curvature of the tube characteristic. In the second, the signal is placed on a grid adjacent to the cathode and the oscillator is placed on a following electrode. The third way connects the oscillator to the grid adjacent the cathode and the signal to a later electrode. The conversion operation of all three is readily analyzed from the signal-electrode-to-plate transconductance, g_m, vs. oscillator-electrode voltage curve. It should be noted that this is true even when a local oscillator is included in the same tube structure.

By a Fourier analysis of the signal-grid transconductance vs. oscillator-electrode voltage curve, the conversion transconductance is found in the same way that the fundamental component of plate current in an amplifier is found by analysis of the plate-current vs. control-grid voltage curve. This has already been brought out in other papers[1, 2]. Mathematically, letting $t = 0$ to be the time at which the fundamental component of oscillator-electrode alternating voltage crosses zero, the conversion transconductance, g_c is given by

$$g_c = \frac{1}{2\pi} \int_0^{2\pi} g_m \sin \omega t \, d\,(\omega t)$$

where ω is the angular frequency of the oscillator. Roughly, a close estimate of the maximum conversion transconductance of practical tubes is given by 28 per cent of the maximum signal-grid transconductance attained over the oscillator cycle. By making use of such an analysis together with the results of recent studies of tube fluctuation noise, an interesting comparison may be made among modulators of the three basic types.

Let us consider four fictitious modulators or mixers, all made with similar cathode and first grid structures, the first a triode, the second a pentode, and the third and fourth hexodes*. These are shown sche-

[1] A New Tube for Use in Superheterodyne Frequency Conversion Systems: Nesslage, Herold, and Harris; *Proc. I.R.E.*, Vol. 24, pg. 207, Feb. (1936).

[2] The Operation of Frequency Converters and Modulators; E. W. Herold; to be published.

* The considerations also apply to heptodes.

matically in Figure 1. The triode and the pentode are to be operated with both oscillator and signal on the control grid, while the hexodes are operated, in the one case, with the oscillator on an outer electrode and, in the other, with the oscillator on an inner electrode. The signal-grid characteristics of each of these types is shown at the right, in every case as a function of the oscillator-electrode voltage. In the first two, the latter electrode is identical with the signal electrode so that the curves show simply the transconductance, g_m, and plate current, I_b, vs. control-grid voltage. In the other two the transconductance of one grid (the signal grid) is plotted against the voltage on another, namely the oscillator grid. In each case, the shape of the curve is typical of practicable tubes. In any event, the conclusions to be drawn are not altered by minor variations in curve shape. It should be remembered that, when the signal is placed on the same electrode as the oscillator, this electrode is not permitted to draw excessive grid current. Thus, the curves need not be extended into the positive grid region in the first two cases.

To the right of Figure 1 is given a tabular summary of conversion transconductance and noise for the four fictitious tubes. A brief discussion of the derivation of the values will be in order. If the transconductance of the triode at zero bias be called g_o and an oscillator is applied to the grid at approximately the optimum point, a Fourier analysis of the transconductance vs. bias curve gives a conversion transconductance of about 0.28 g_o. Going to the pentode, it is assumed that the screen current is 20 per cent of the total current so that the pentode curves are lowered by this amount. The conversion transconductance is correspondingly reduced. It should be noted that g_o is the cathode transconductance not the plate transconductance (in the triode the two are the same). In the hexode first detector with signal applied to the first grid, experience indicates that the plate current is not greatly in excess of 50 per cent of the total current when the oscillator grid is at its maximum positive excursion. The curves shown on Figure 1 are plotted against No. 3 grid voltage and are typical of the shape usually obtained. In such a tube, the oscillator-electrode control is by current division so that the transconductance and plate-current curves are similarly shaped. Analysis indicates a conversion transconductance of 0.14 g_o.

The lower hexode of Figure 1 cannot be compared directly with the other tubes because there is no direct dependence of the signal-grid (No. 3 grid) characteristics on the cathode and first grid structure. In general, the maximum No. 3 grid transconductance depends on the uniformity of the structure and on the No. 3 grid area. In most practical tubes a partial virtual cathode is formed between No. 2 and

No. 3 grids. The curve shapes drawn are typical of such tubes. It is difficult to attain a maximum transconductance (g_x in the figure) which exceeds the g_o of the other tubes without the tube being extremely critical to voltage variations. It is not likely that the hexode with signal on the outer grid offers much more possibility as regards conversion transconductance than the previous hexode. It will be observed that the conversion transconductance is approximately 28 per

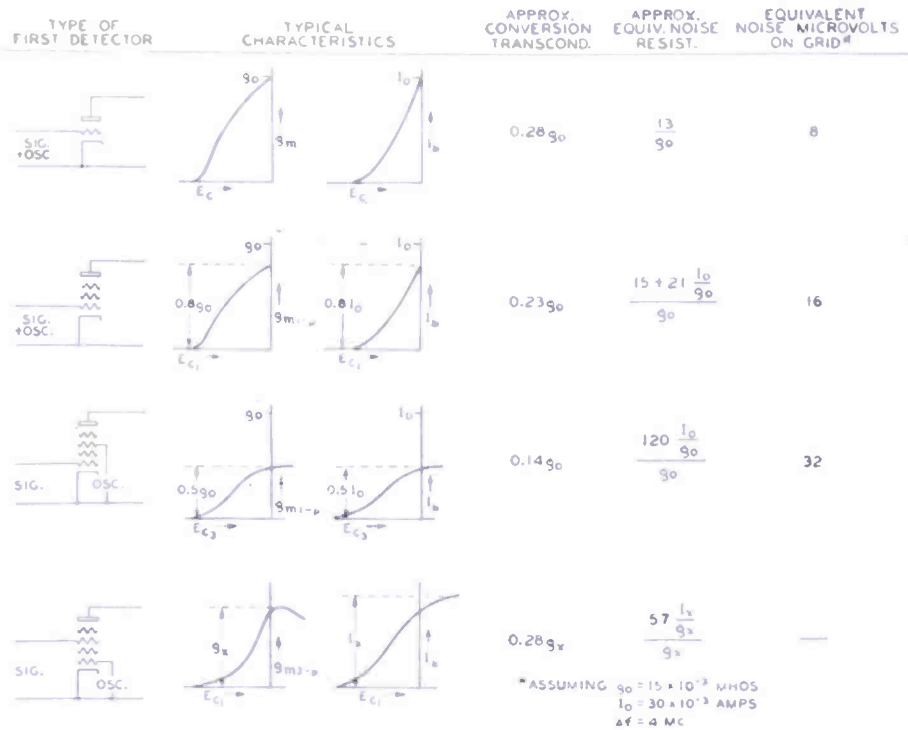

Fig. 1—Comparison of four fictitious modulators presumed to have similar cathode and first-grid structures.

cent of the maximum transconductance attained over the oscillator cycle in all the four illustrations of Figure 1, even though the curve shapes are markedly different.

The mean-squared noise of the modulator part of the converter stage may be shown to be the time-average value of the mean-squared noise over an oscillator cycle. Thus, if $\overline{i_{pn}^2}$ be the mean-squared noise current in the plate of a tube at a fixed bias, the application of an oscillator voltage will vary this noise current periodically. The resultant noise at an intermediate frequency remote from that of the oscillator will be the average over the oscillator cycle, or

$$\overline{i_{i\text{-}f}^2} = \frac{1}{2\pi}\int_0^{2\pi} \overline{i_{pn}^2}\, d\,(\omega t)$$

If an expression for $\overline{i_{pn}^2}$ is known for a tube, therefore, it becomes possible to estimate the noise such a tube will produce when used as converter or mixer. Expressions for $\overline{i_{pn}^2}$ in the triode and in the pentode have been derived[3].

The noise fluctuations of a tube are conveniently expressed in terms of an equivalent noise resistance, R_{eq}, at the grid. This resistance is independent of band width and gives a direct indication of the mean-squared noise voltage per unit band-width of the grid. For these reasons it is a convenient means for comparing the noise-to-signal ratio for various tubes. The value of the resistance is obtained by equating the thermal noise of a resistor at room temperature to the equivalent grid-noise voltage of the tube. Thus, for a converter or mixer,

$$R_{eq} = \frac{1}{4kT_R} \; \frac{\overline{i_{i\text{-}f}^2}}{g_c^2 \; \Delta f}$$

where T_R is the temperature of the room, usually taken as 293°K. The values of equivalent noise resistance in Figure 1 were found in this way.

For the triode, the plate noise is approximately given by

$$\overline{i_{pn}^2} = 0.64 \;(4kT_K)\; g_m \;\Delta f$$

where k is Boltzman's constant, T_K is the absolute temperature of the cathode ($\approx$ 1000°K) and Δf is the band-width. Under converter conditions, therefore, the triode will have

$$\overline{i_{i\text{-}f}^2} = \frac{1}{2\pi} \int_0^{2\pi} \overline{i_{pn}^2}\; d\;(\omega t) = 0.64\;(4kT_K)\;\overline{g_m}\;\Delta f$$

where $\overline{g_m}$ is the average transconductance over the oscillator cycle. Computation of $\overline{g_m}$ for the triode shows:

$$\overline{g_m} = 0.47\; g_o$$

Using the value for g_c in terms of g_o permits the equivalent noise resistance of Figure 1 to be found; namely

$$R_{eq} = \frac{13}{g_o}$$

The equivalent noise resistance for the triode, and for the other tubes as well, as computed in this way should be considered as a lower limit

[3] Shot Effect in Space-Charge Limited Tubes: B. J. Thompson and D. O. North; Paper given at Rochester Fall Convention, the Inst. of Radio Engineers, Nov. (1936); cf. also paper by Thompson, North, and Harris in this issue.

rather than as a norm. Noise within the tube other than that exhibited in the plate circuit has been neglected and will, in general, increase practical equivalent noise resistances.

In the pentode, fluctuations in the current distribution between screen and plate are also present. In the usual tube with nearly uniform screen grid, an approximate expression for the noise[3] is

$$\overline{i_{pn}^2} = \left[0.64 \; (4kT_K) \; g_{m1\text{-}p} \frac{I_b}{(I_{c2} + I_b)\,\sigma} + 2e \frac{I_{c2}\, I_b}{I_{c2} + I_b} \right] \Delta f$$

where e is the charge on the electron, I_{c2} is the screen current, and σ is the ratio of total transconductance to the conductance of an equivalent diode. The factor $\frac{I_b}{(I_{c2} + I_b)\,\sigma}$ is very closely unity when the screen current is around ¼ of the plate current as in the usual tube and will be so taken here. Since the ratio of screen to plate currents in a pentode is approximately constant, we may find the noise under converter conditions by the approximate expression,

$$\overline{i_{i\text{-}f}^2} = \left[0.64 \; (4kT_K) \; \overline{g_{m1\text{-}p}} + 2e \frac{\overline{I_b}}{1 + \frac{I_b}{I_{c2}}} \right] \Delta f$$

where $\overline{I_b}$ is the average value of the plate current over the oscillator cycle (i.e., the actual operating plate current). Analysis of the typical curves of Figure 1 with the oscillator voltage applied at about the optimum point shows $\overline{g_m} = 0.37\; g_o$ and $\overline{I_b} = 0.28\; I_o$ where I_o is the maximum value of the total current. Since it has been assumed that $I_{c2}/I_b = 0.25$, the mean-squared noise can be determined in terms of I_o and g_o. The equivalent noise resistance will be found to have the value

$$R_{eq} = \frac{15 + 21\; I_o/g_o}{g_o}$$

The plate noise, $\overline{i_{pn}^2}$, of a hexode or heptode is not so easily determined on theoretical grounds because of the possibility of a partial virtual cathode in the tube and the presence of secondary emission. For this analysis, however, secondary emission may be considered as negligibly small (it can be made so by a suppressor grid, for example) and virtual cathode formation may be neglected, at least for the type

[3] ibid.

of tube in which the signal is placed on an inner grid and the oscillator voltage on an outer grid. Under these assumptions, the heptode or hexode plate noise should be primarily current-distribution noise which is approximately equal to

$$\overline{i_{pn}^2} = 2e \frac{I_b I_{c2}}{I_b + I_{c2}} \Delta f$$

Under converter conditions, then,

$$\overline{i_{i\text{-}f}^2} = 2e \frac{\Delta f}{2\pi} \int_0^{2\pi} \frac{I_b I_{c2}}{I_b + I_{c2}} d\ (\omega t)$$

Since $I_b + I_{c2} = I_o$, which is approximately a constant, it may easily be shown that the integral becomes

$$\overline{i_{i\text{-}f}^2} = 2e \left(\overline{I_b} - \frac{1}{I_o} \overline{I_b^2} \right) \Delta f$$

where $\overline{I_b^2}$ is the average of the square of the plate current over the oscillator cycle while $\overline{I_b}$ is, of course, the average plate current. It is found that $\overline{I_b} = 0.20\ I_o$ and $\overline{I_b^2} = 0.09\ I_o^2$ for the typical curve of Figure 1. Expressing the results in terms of $\overline{I_b}$ indicates $\overline{i_{i\text{-}f}^2} = 0.55$ $(2e)\ \overline{I_b}$ which approximately checks measured values.* Use of this formula gives an equivalent noise resistance for the heptode with outer-grid oscillator injection which is

$$R_{eq} = \frac{120\ I_o/g_o}{g_o}$$

The hexode or heptode with the signal on an outer grid cannot be considered as free from virtual-cathode formation and so is not amenable to similar treatment. Experimental data, however, indicate that such a tube will have a noise (with oscillator applied) of

$$\overline{i_{i\text{-}f}^2} = 2e\ \overline{I_b}\ F^2\ \Delta f$$

where F^2 is often from 0.5 to 0.6 when secondary electrons from the screen are suppressed, and somewhat higher otherwise. The mean plate current for the typical curve of Figure 1 is

$$\overline{I_b} = 0.41\ I_x$$

* i.e., in tubes with negligible secondary emission from the screen grid. In tubes having secondary emission, the noise-per-unit plate current is usually greater.

where I_x is the maximum plate current over the oscillator cycle. The equivalent noise resistance for this tube was computed assuming $F^2 = 0.55$ and becomes

$$R_{eq} = \frac{57\, I_x/g_x}{g_x}$$

The last column of Figure 1 was computed by assuming $g_o = 15 \times 10^{-3}$ mhos, and $I_o = 30 \times 10^{-3}$ amps., values which are typical of a high-transconductance tube design. The column shows the grid voltage which is equivalent to the square root of the mean-squared noise voltage over a 4-Mc band. This voltage may be considered as limiting the useful sensitivity of the receiver.

It is now possible to compare the four fictitious tubes of Figure 1 for use in the television converter stage.* As to noise, the figure is self-explanatory and, in receivers with no r-f stage, there is little choice except between the first two. The conversion gains of the four types working into loads typical of television-receiver practice is also in favor of the first two. The triode, however, has a practical disadvantage over the pentode in that feedback must be taken into account. Since the i-f circuit is capacitive at signal frequency, an input resistance at signal frequency is obtained which is excessively low unless neutralization is used or the grid-plate capacitance is very low in comparison with the i-f tuning capacitance. The feedback may also increase the noise appreciably. Thus, although it would be possible to use the triode, the pentode is probably a safer choice.

Regarding input loading at high frequencies, the first three types will have only minor differences because the cathode and signal-grid structures were assumed to be similar. It should be remembered, however, that cathode injection of the oscillator is not permissible because of the additional input loading due to the added cathode inductance. The last type with an outer grid used for the signal will be substantially free of positive input loading (the input resistance of such a tube is usually negative).

Interaction of oscillator and signal circuits is greatest when their voltages are placed on the same grid, especially since cathode injection is contra-indicated. Least interaction may be obtained in the third type, the hexode with oscillator on an outer grid, which may be made substantially free of interaction. Among the undesirable effects which may be attributed to interaction are radiation and alignment difficulties. Radiation may be reduced to some extent by selective circuits

* More detailed information on the operation of the different types of modulator will be found in Reference 2.

between antenna and grid while alignment difficulties can be minimized by coupling the signal circuit loosely to a non-critical portion of the oscillator circuit. The latter procedure is permitted in a high-transconductance tube because of the small oscillator voltage required.

It appears to be difficult to combine the function of local oscillator in the same mount with a modulator when operation is to extend to such high frequencies. There is some advantage in this respect in the last of the types, the hexode with oscillator applied to an inner grid. This statement has been borne out by the relatively larger number of types and greater popularity of such converters in the broadcast field. Because of noise and interaction due to "space-charge coupling", this last type is not too well suited for television use.

It may be concluded that, although none of the four fictitious tubes meets all of the television receiver requirements, the triode or pentode with oscillator and signal on the same grid will give highest gain and

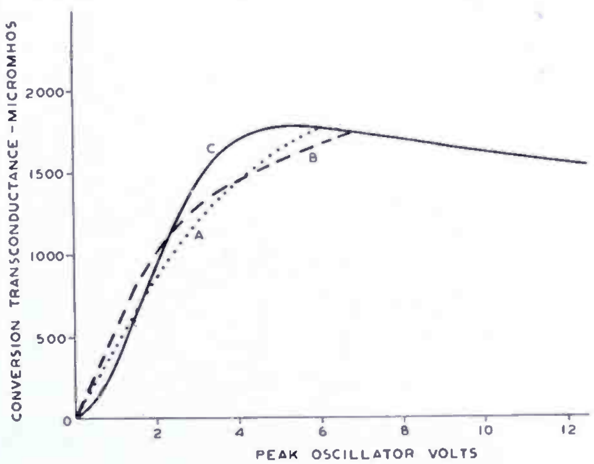

Fig. 2—Conversion transconductance of a typical variable-mu pentode. Signal and oscillator voltages applied to the control grid. Curve A, fixed-bias operation; curve B, cathode-resistor bias; curve C, grid-leak bias. Curves A and B have been drawn only up to the grid-current point.

lowest noise, and these types are, therefore, highly advantageous, particularly when no r-f stage is used.

Practical Data on Television Converter Stage

In a practical circuit using a pentode first detector, the methods of applying the oscillator and of obtaining bias are of considerable importance. It is highly desirable that variations of effective oscillator voltage, which are unavoidable when channels are switched, have a minimum effect on gain. It is also desirable that the oscillator voltage never swing the control grid so far positive as to draw large amounts of grid current. A means of achieving both ends consists in the use of automatic bias. This may be obtained either by a cathode self-bias resistor or a high-resistance gridleak and condenser or both. An illustration of the improvement which may be made in this way is shown in Figure 2 which is taken from another paper*. For the curve labelled (A) a fixed bias was used at approximately an optimum point. The curve is stopped at the grid-current point. Curve (B) shows cathode-resistance bias and curve (C) a high-resistance gridleak used for bias. It is evident that least critical operation is obtained with gridleak bias. With such operation, the bias is obtained by rectification of the effective oscillator voltage. The gridleak may be made part of an avc filter, but it should be made considerably higher than the resistance common to other tubes in order to avoid biasing them excessively.

The cathode current of a high-transconductance tube used with gridleak bias reaches excessive values when the oscillator voltage is

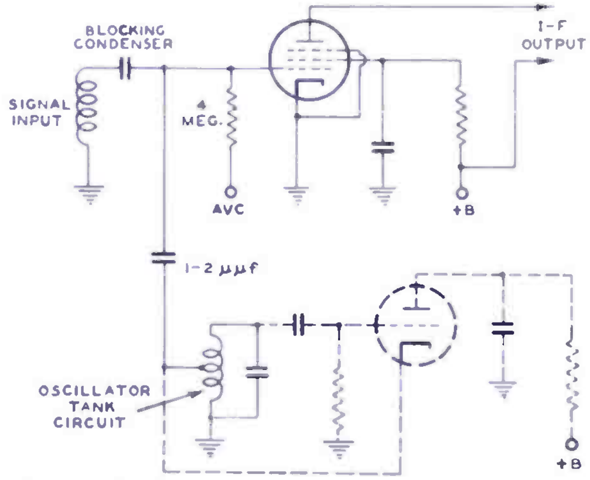

Fig. 3—Circuit of a typical converter stage using a high-transconductance pentode with a separate oscillator.

accidentally removed. This may be avoided by the use of series-screen resistor operation. Series-screen operation makes the gain even less dependent on oscillator voltage than the best of the curves of Figure 2.

Figure 3 shows a practical circuit using a pentode with gridleak bias and a series-screen resistance. The oscillator is applied across the signal circuit through a small condenser which is tapped down on the oscillator tank circuit. By this means the reaction of the signal circuit on the oscillator circuit may be kept small. The type of oscillator circuit used is not of great importance. For the sake of completeness, however, a Hartley oscillator circuit is drawn in dotted lines.

Table I shows data pertinent to television converter operation, which have been taken on a number of commercially available tube types. Two types of pentagrid tubes commonly used in broadcast receivers are included for comparison. It is evident that they are

* See Reference 2.

TABLE I

Tube Type	Conv. Transconductance Micromhos	Equiv. Noise Res. Ohms Measured	Equiv. Noise Res. Ohms Calculated	Equiv. Grid. Noise* Micro-volts	60 Mc Input Resistance Ohms	60 Mc Signal Grid Current Micro-amperes
6SA7 Pentagrid Converter	450	210,000	220,000	116	− 10,000	5
6L7 Pentagrid Mixer	400	210,000	230,000	116	+ 2,300	10
6J5 Triode	1,000	5,800	3,700	20	(†)	2
1853 Pentode	1,900	13,000	18,000	29	+ 8,000	1
1852 Pentode	3,600	3,000	3,400	14	+ 2,500	1

* $\Delta f = 4$ *Mc* † Depends on feedback.

unsuited for television use because of their high noise and low gain. The use of avc is not practical for these because of the high signal-grid current which is caused by a transit-time effect. The other types in the table had oscillator and signal applied to the control grid and were used with gridleak bias as in the circuit of Figure 3. It will be noted that the chief disadvantage of the pentode 6AC7/1852 is its low input resistance. This low resistance is mainly due to cathode-lead inductance and may be neutralized, at least partially, by several circuit arrangements.

The measured equivalent-noise resistances of the table were taken with an oscillator operating at 60 Mc and an i-f amplifier operating at 10 Mc. The measurements were made with a saturated diode connected in the plate circuit and conventional procedure. The data were referred to the grid circuit by separate measurement of conversion transconductance. The calculated equivalent noise resistances were obtained from the equations given in Figure 1, using the appropriate formula for each type of operation. The application of the equations was made by computing the values of g_o, I_o, etc., from the conversion transconductance and operating plate current, using the relations found for the typical curve shapes of Figure 1. Rough agreement with measured values is seen in every case except the triode. It is believed that small, but unavoidable feedback in the triode increased the measured noise; the data are evidence that it is difficult to realize the low value of triode first-detector noise predicted by theory.

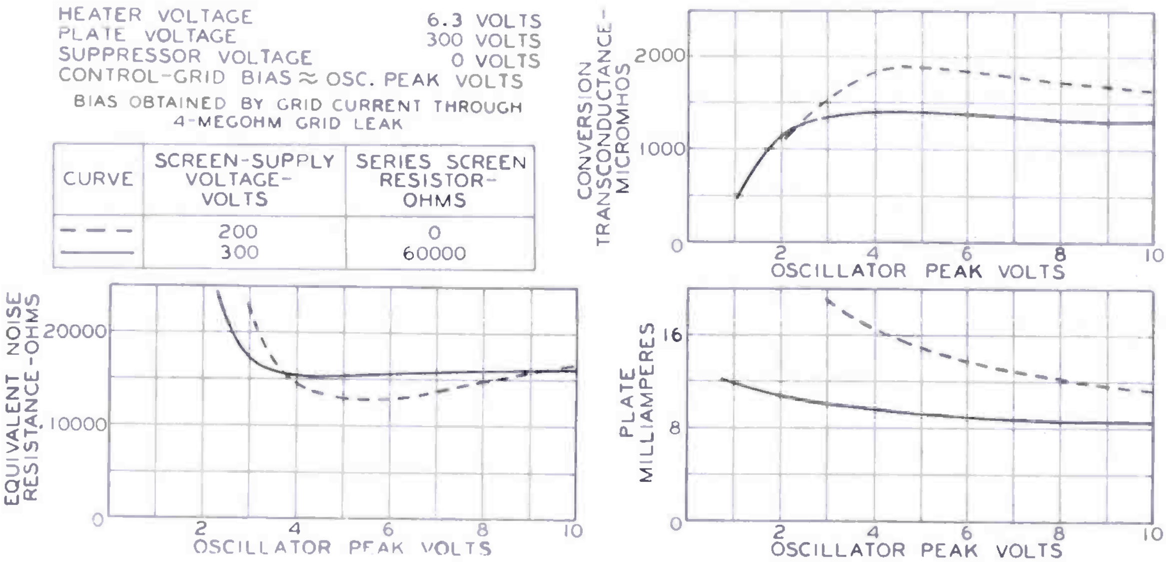

Fig. 4—Typical first-detector characteristics of the type 6AB7/1853.

It should be mentioned that the noise data of Table I were taken with a low impedance in the signal-grid circuit. The noise values must, therefore, be considered as minimum values because they neglect noise currents which are induced in the signal-grid itself by the passage of electrons,[4] and which may be appreciable at television frequencies.

More detailed data on the conversion performance of types 6AB7/1853 and 6AC7/1852 are shown by the curves of Figures 4 and 5. The data were measured using a 4-megohm gridleak and are shown for both fixed and series-screen supply. In neither case is the

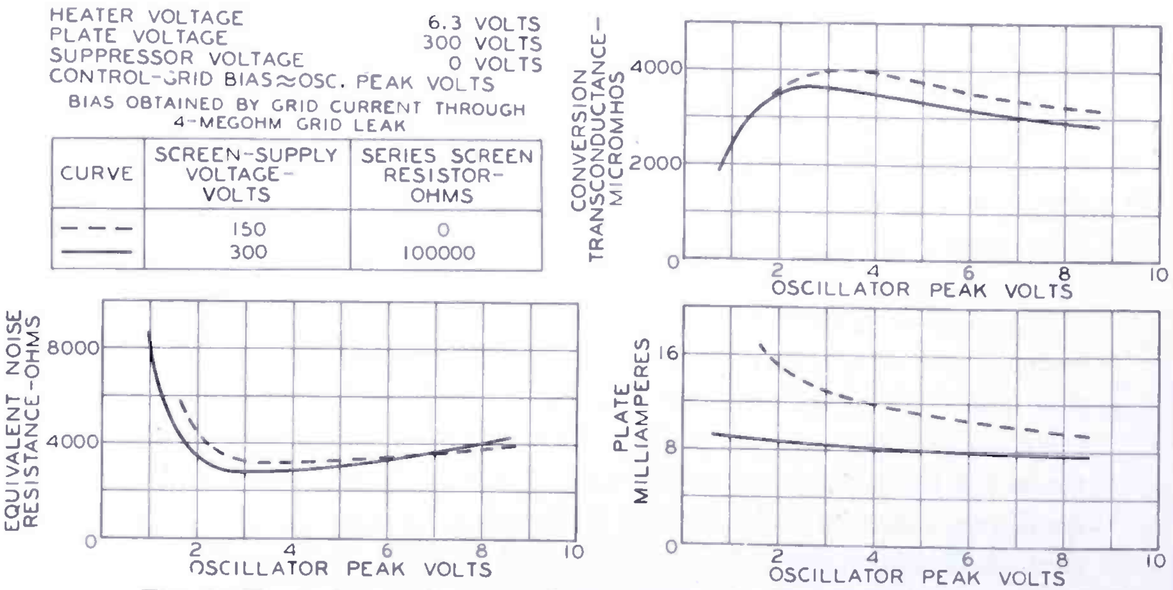

Fig. 5—Typical first-detector characteristics of the type 6AC7/1852.

[4] S. Ballantine: "Schrot-Effect in High-Frequency Circuits"; *Journ. of Frank. Inst.*, Vol. 206, pp. 159-167, Aug. (1928).

required oscillator voltage critical as to value. The conversion-transconductance curves for series-screen supply are somewhat flatter than those for fixed supply. Series-resistor-screen supply has the additional advantages that the plate current is held to reasonable values at low oscillator excitation and that variations between tubes are somewhat minimized.

The 3000-ohm equivalent-noise resistance of the 6AC7/1852 used as first detector is remarkably low and compares favorably with that of most amplifier tubes. The type 6AB7/1853, for example, when used as an amplifier, has an equivalent noise resistance of about the same magnitude. It is interesting to compare the potential signal-to-noise ratios of two receivers, each with a 6AC7/1852 in the converter stage followed by a 6AB7/1853 i-f amplifier, but the second having a 6AB7/1853 r-f stage added. If the i-f stage gain is assumed to be sufficiently high (e.g., > 5), the noise of further i-f stages may be considered negligible. Calling the gain of the first detector stage A_D and the gain of the r-f stage A_{RF} and assuming each tube as having an equivalent-noise resistance of R_{eq} ohms it is found that

R_1 = Noise equivalent resistance of receiver without

$$\text{r-f stage} = R_{eq}\left(1 + \frac{1}{A_D^2}\right)$$

R_2 = Noise equivalent resistance of receiver with

$$\text{r-f stage} = R_{eq}\left(1 + \frac{1}{A_{RF}^2} + \frac{1}{A_{RF}^2\, A_D^2}\right)$$

It is evident that if A_{RF} is infinite, the noise resistance of the receiver with an r-f stage is at a minimum. Even with this ideal and impossible condition, the improvement in signal-to-noise ratio over the receiver without an r-f stage will be only $\sqrt{1 + \frac{1}{A_D^2}}$ or less than 6 per cent for $A_D > 3$. It is impractical, therefore, in a television receiver, to obtain marked improvement in signal-to-noise ratio by adding an 6AB7/1853 r-f stage to a receiver with an 6AC7/1852 first detector, no matter what the r-f stage gain may be. An improved signal-to-noise ratio can only be obtained by using a 6AC7/1852 for the r-f stage.

Acknowledgment

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A NEW METHOD FOR DETERMINING SWEEP LINEARITY

By

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Summary—This paper contains the description of a new method for determining sweep linearity in magnetic deflection systems. Basically, the method consists in observing, on an oscilloscope screen, the voltage wave developed across a small pickup coil positioned within the deflection yoke. This wave is the differential of the yoke flux wave, and is, therefore, a measure of the slope of the sawtooth flux wave, and also of the spot velocity.

The latter part of the paper includes a description of an instrument, which embodies the principles discussed in the first section.

In any television deflection system, whether it be used in conjunction with a Kinescope, Iconoscope or Monoscope, it is essential that the distribution or positioning of the multiplicity of elements in the received picture be the same as in the originally televised scene. This condition can best be attained by employing certain prescribed wave shapes in both the horizontal and vertical sweeps. In the case of receivers, wherein all points on the Kinescope screen are nearly equidistant from the center of deflection, the sweep voltages or currents should vary linearly with time, to produce constant spot velocity during the scanning intervals. In Monoscopes and Iconoscopes slightly different conditions exist, because the beam is caused to scan a flat plate, with the result that the angular velocity of the beam must vary somewhat during the scanning cycle, to produce constant spot velocity.

Regardless, however, of the type of scanning waveforms required, it is essential that there be available some easily applied method for ascertaining the departure of the actual scanning waves from their prescribed optimum shapes. In this connection, it is pertinent to note that the majority of cases in which one wishes to determine sweep wave forms is associated with receivers, in which constant sweep velocity is desired. Therefore, the discussion to follow will be concerned principally with measurements of sweep linearity, although it should be understood that the method to be described is also applicable to non-linear sweep studies.

Methods Now in Use for Determining Sweep Linearity

One method, useful in conjunction with Kinescope deflection circuits, has for its basis the application of harmonics of the sweep frequencies to the Kinescope grid, while the sweeps are operating. The source of these harmonic signals generally consists of a chain of frequency multipliers, excited initially by voltages at the line or field base frequencies.

The result of applying these waves to the Kinescope grid is a pattern on the Kinescope screen, composed of a series of vertical or horizontal bars. The positioning (whether the bars are vertical or horizontal) and the number of bars are determined by the frequency of the applied grid voltage. Vertical bars will appear if the waves applied to the grid have frequencies that are harmonics of 13,230 cycles. Horizontal bars are the result of the application of frequencies harmonically related to 60 cycles, but less than 13,230 cycles.

Thus, if the 100th harmonic of the line frequency were applied to the Kinescope grid, the screen pattern would consist of 100 vertical bars (some of these, say 10 per cent, would occur during the sweep return time). The horizontal distribution is determined by observation of the horizontal spacing of the bars.

There are several defects inherent in this method, which make it difficult to apply when accurate results are desired. One of these is the inaccuracy associated with visual determination of the bar distribution. It is quite difficult, practically, to note crowding of the bars unless the effect is considerably marked. In addition there is no direct way of denoting the departure from sweep linearity, unless it is expressed in terms of the number of bars per inch over different portions of the scanned area. Secondly, the method involves considerable equipment, for a bar generator, locked in with the synchronizing signals, is required. Thirdly, the limited number of bars appearing during the return trace makes accurate determination of the fly-back time somewhat difficult.

It is important to note that the bar method of determining sweep linearity is useful only in conjunction with a Kinescope. It might be thought, at first, that the method could be applied to Iconoscopes and Monoscopes by impressing harmonics of the sweep frequencies on their grids. It should be noted, however, that to realize the advantages of this arrangement it would be necessary to have available, as a visual indicating device, a monitor Kinescope whose sweeps were precisely linear, or one in which the departures from linearity were accurately known.

A second, and less involved, method for determining distribution, useful in magnetic deflection systems, involves the direct observation of the wave shape of the deflecting coil current on an oscilloscope screen. This may be done by inserting a small resistor in the low-potential end of the yoke winding, and by applying the resultant voltage drop to the vertical amplifier of the oscilloscope.

Inaccuracies associated with this method may be due to one or more of several factors. Among these are non-linearity of the oscilloscope amplifier or time base, difficulty in observing small variations in sawtooth slope, possibility of the inserted resistor (even though small) altering the wave shape of the yoke current, and the fact that distributed capacitances across the yoke windings cause currents which play no part in the formation of the flux wave to flow through the resistor.

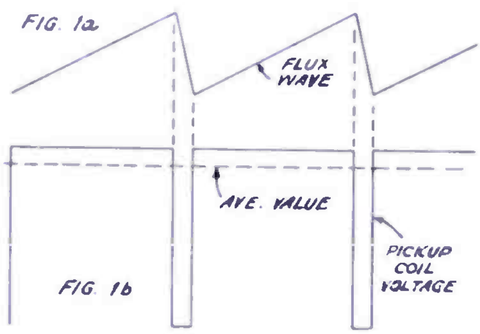

Fig. 1

Description of the New Method

The limitations to accuracy inherent in the previously described systems are not present in this method because it depends, basically, upon a determination of the slope of the actual flux wave which exists within the magnetic deflection yoke. The validity of the method is predicated upon the fact that the open-circuit voltage developed across a small pickup coil, placed in the magnetic field of the yoke, is

$$e_2 = \pm K \frac{d\phi}{dt}$$

where ϕ is the instantaneous value of yoke flux. K is a factor which is dependent upon the number of turns in the pickup coil and its proximity to the deflection coil.

The wave form of the voltage developed across the pickup coil is, therefore, the true differential or slope of the sawtooth flux wave. The pickup coil voltage wave is also an indication of the manner of variation of spot velocity on the Kinescope screen, for the spot displacement is directly proportional to the yoke flux, and therefore, $\frac{d\phi}{dt}$ may be interpreted as being equivalent to the time derivative of spot displacement, which is, of course, the spot velocity.

Let Figure 1a represent the flux wave existing within a deflection yoke; in this case, assume that it is linear during both scanning and fly-back intervals. The voltage developed across the pick-up coil is shown in Figure 1b.

The slope of the sawtooth is constant and positive during the scanning interval, consequently the pickup voltage developed (which is proportional to $\frac{d\phi}{dt}$) is also constant and positive.

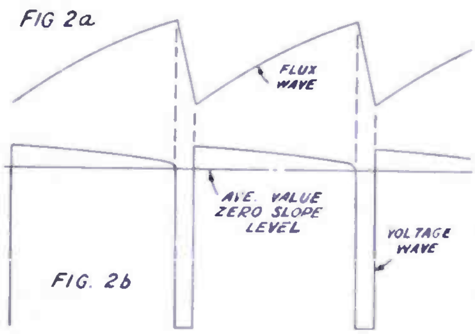

Fig. 2

During the return time the slope is negative and much greater than its scanning interval value. Therefore, negative voltage peaks exist in the wave of Figure 1b. The relative values of slope (or spot velocity) are determined by reference to a line drawn in Figure 1b which represents the average value of the voltage wave. The slope of the sawtooth at any instant is proportional to the instantaneous voltage developed with respect to the average value of the voltage wave. Thus, at the points of slope reversal in Figure 1a the slope and spot velocity are zero. In Figure 1b, these same points are found at the intersection of the voltage wave with the line indicating the average value of the wave in Figure 1b.

It follows, therefore, that the average value of the voltage wave is also the reference level for all flux wave slopes; i.e., it represents zero slope in the sawtooth, or zero spot velocity.

Actually, in practice, the flux and voltage waves are not as geometrically precise as are shown in Figures 1a and 1b. Small variations in slope are likely to exist, even in the best attainable sawtooth waves. Figures 2a and 2b show the conditions existing when the flux wave contains some overall curvature.

Although the voltage wave of Figure 2b contains all the information required for determination of the departure of the flux wave shape from linearity, it cannot be used directly for an accurate measurement of spot velocity, because the large negative voltage pulses generated during the retrace time comprise some 90 per cent of the total voltage swing, whereas the voltage variations developed during the scanning interval, in which we are particularly interested, occupy only about 10 per cent of the total developed voltage. Consequently, if the wave of Figure 2b were applied to an oscilloscope for observation, the major portion of the screen would be occupied by the retrace pulses, and the scanning voltage variations would not be accurately discernible.

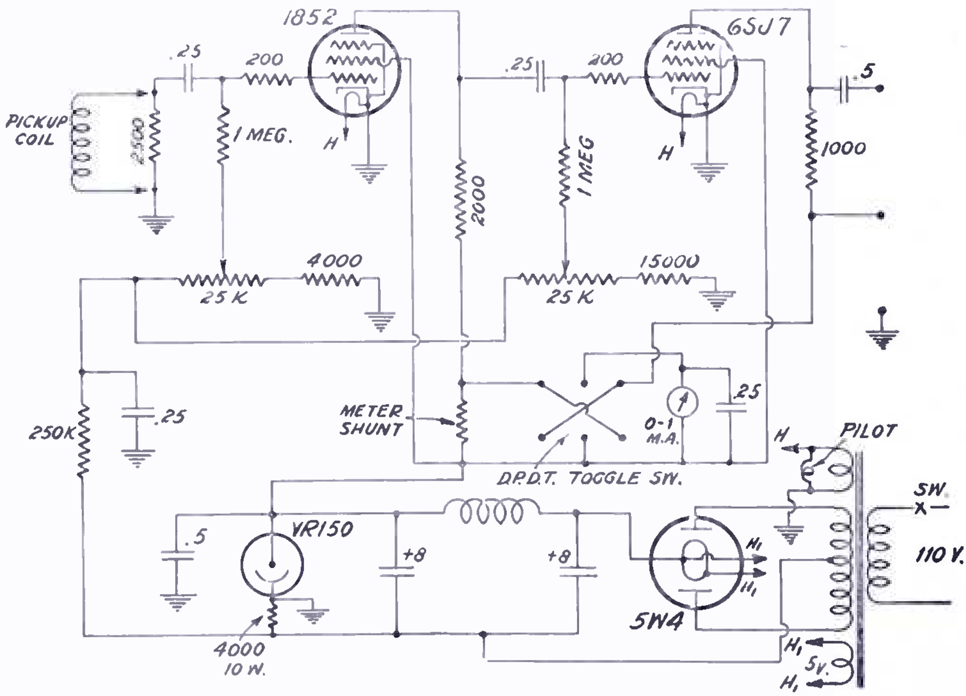

Fig. 3

Since it is with these latter variations that we are most concerned, it becomes necessary to clip off the retrace pulses from the pick-up voltage wave before applying it to an oscilloscope. In this way, if the wave is clipped at a point close to its average value or zero slope level, the entire oscilloscope screen may be used for studying the velocity variations during the scanning interval. Note that a certain degree of clipping action could be obtained by overloading the oscilloscope amplifier, but this would not be practicable, because it would introduce an indeterminacy as to the location of the zero slope level in the clipped voltage wave. This level may be discerned quite readily, if the wave is clipped externally to the oscilloscope. A knowledge of its location on the clipped wave is essential to the determination of the relative slope variations in the flux wave.

In the instrument which embodies the principles discussed previously (for which a complete circuit diagram is shown in Figure 3) a small pickup coil (10 μh for the horizontal sweep and 2 mh for the vertical) is positioned within the deflection yoke (the vertical output tube in the deflection system is removed if the horizontal distribution is to be studied, and vice versa). The yoke flux linking the pick-up coil induces a voltage of the general type shown in Figure 2b, wherein the polarity and amplitude may be adjusted by rotating the yoke about the pickup coil. A peak-to-peak voltage of about 0.5 volt is obtained, and this is applied, with the retrace pulses acting in a positive direction, to the grid of an amplifier tube, the 1852 shown

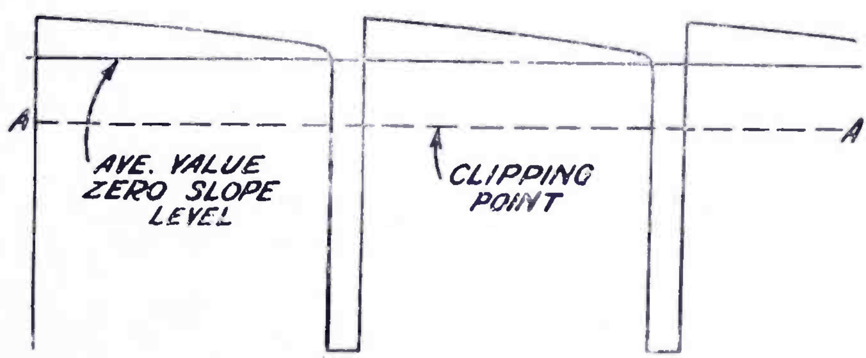

Fig. 4

in Figure 3. The amplifier output, which appears in reversed phase, has an amplitude of about 6-8 p-p volts. It is essential that this amplification be effected with no resultant clipping or rectification, so that the true position of the average value of the wave may be maintained for subsequent calibrating purposes.

To check the presence of rectification a 0-1 ma meter is provided in the instrument; this meter, in conjunction with a switch and a shunt to reduce the meter sensitivity by a factor of 10, may be placed in the 1852 amplifier plate circuit. There should be no change in plate current due to the application of the signal, for proper operation. Figure 4 represents the voltage wave appearing across the plate load of the 1852 for a typical sweep condition, i.e., sweep velocity decreasing during the scanning interval.

This wave is applied to the grid of the following tube in the unit, a 6SJ7, which acts as a clipper to remove the negative retrace peaks. The clipper tube is biased at a point which causes the retrace peaks existing below the line A-A in Figure 4 to extend beyond plate-current cut-off, with the result that they do not appear in the plate circuit of the clipper tube. This action is shown in Figure 5.

Since the wave of Figure 4 is applied to the clipper tube's grid through a conventional RC coupling network, the average value of the grid voltage assumes a position in Figure 5 which is coincident with the steady bias of the clipper tube, i.e., the d-c component of the output of the amplifier tube is lost in transmission through the coupling condenser.

The plate-current wave of Figure 5 flows through the load resistor in the clipper tube's plate circuit to create an output voltage wave as shown in Figure 6.

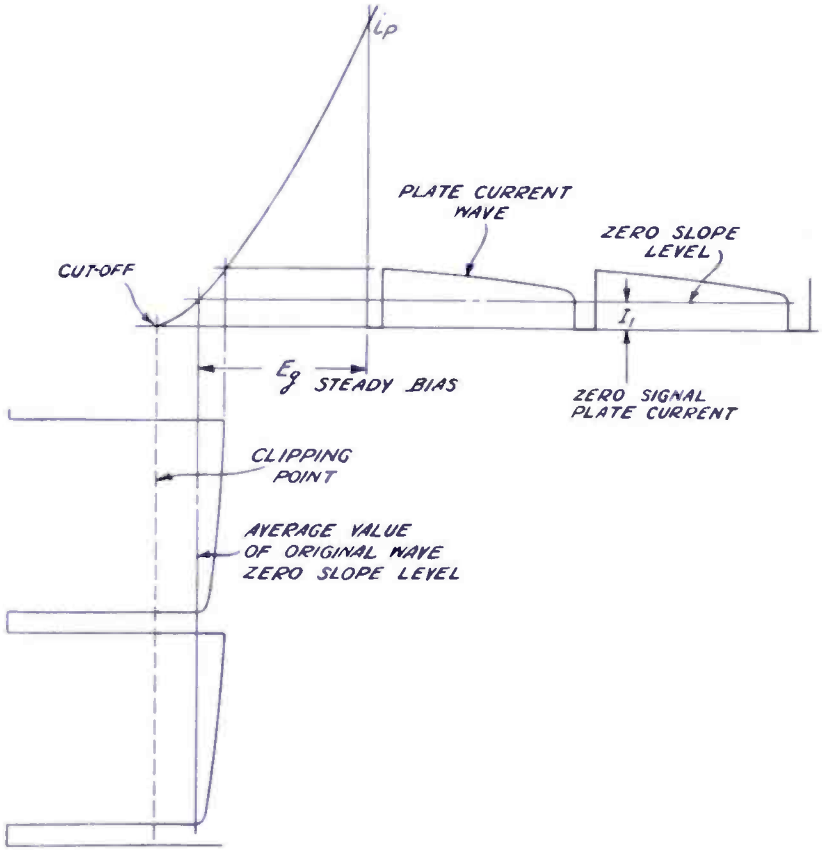

Fig. 5

The position on the clipped voltage wave of Figure 6 of the level corresponding to zero-sweep velocity is determined from Figure 5 by noting that the zero-signal plate current of the clipper tube represents the difference between the clipping point (zero current) and the steady bias point, which, in turn, coincides with the average value of the original voltage wave. Consequently, on the clipped voltage wave, the zero-velocity point may be located by measuring down from the clipping point an amount equal to $I_1 R_L$, the zero-signal drop in the clipper's plate circuit.

It will be noted, in examining Figure 5, that there is some distortion of the clipped output voltage wave due to curvature of the i_p-e_g characteristic of the clipper tube. This effect tends to magnify the voltage variations representing changes in sweep velocity during the scanning interval, and, in effect, makes the device more critical of departures from linearity in the sweep. The curvature effect may be taken into account, however, by transferring, to the oscilloscope screen, a non-linear scale which is proportional to the departure from linearity of the i_p-e_g characteristic of the clipper.

There are certain pertinent points about the unit, as exemplified in the diagram of Figure 3, which merit some comment. First, the 0-1 ma meter is used for two purposes; one of these is to read the 1852 plate current both with and without applied signal, to detect the presence of undesired rectification. The second is to measure the zero-

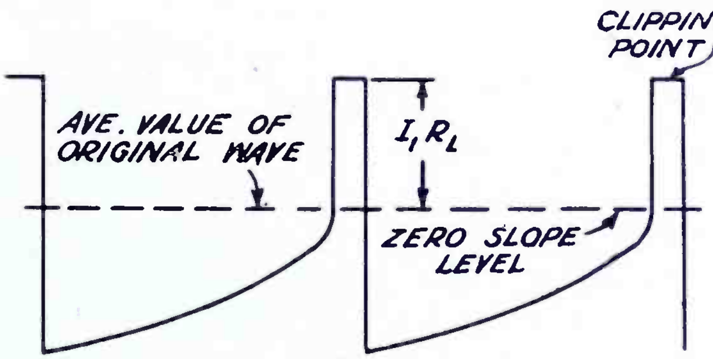

Fig. 6

signal clipper plate current as a means for determining the position of the zero-velocity level on the clipped output wave. The power supply of the unit is self-contained and is regulated at 150 volts with a VR-150. This is done to minimize interstage coupling and, at the same time, to provide a low d-c impedance source for the screen of the clipper tube, to preclude changes in its i_p-e_g characteristic due to rectification in the screen grid. A source of negative voltage is also available. This provides a means of controlling the bias on the two tubes; in the 1852 this control affords a means of regulating the gain and, in the clipper, the variable bias provides control of the clipping point.

The frequency and phase response of the amplifier and clipper must be sufficiently good to permit the accurate reproduction of square waves at both the line and field frequencies. Failure to fulfill this requirement will introduce misleading effects, in the form of tilt in the clipped wave at the field frequency, and overshoot at the discontinuous points at the line frequency. These effects could be erroneously interpreted as being due to velocity variations in the sweeps under observation. In the experimental unit, the desired transient response has been attained by using low values of plate load resistors (to extend the high-frequency range to about 500 kc), and by employing adequate values of RC time constants in the grid coupling circuits for proper operation at the field frequency.

Figures 7 and 8 are two photographic views of the experimental model of the instrument. The two pick-up coils are mounted individually in $1\frac{3}{8}''$ diameter plug-in coil forms. The axes of the coils are positioned at right angles to the axis of the yoke, which is slipped over the assembly. Rotation of the yoke then provides control of the coupling between yoke and pick-up coil. Care was taken to minimize the shunt capacitance in the pick-up coil used at the line frequency to prevent resonance effects. This coil is a single layer solenoid of

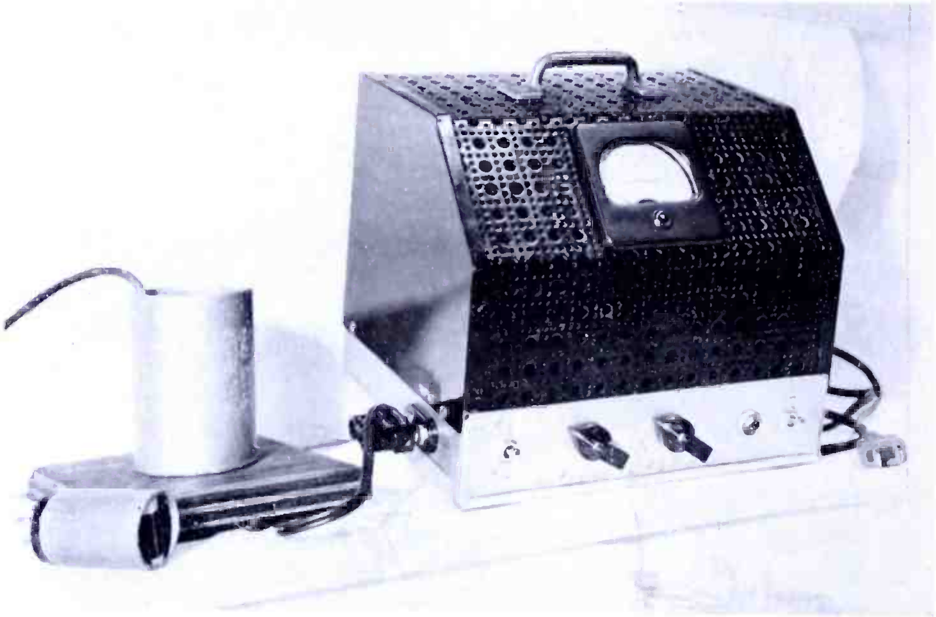

Fig. 7

10 μh wound on a $\frac{1}{2}''$ form. The field frequency coil is bank wound. Its self-inductance is 2 millihenrys.

The technique of operating the instrument is straightforward. To evaluate the velocity variations in a given horizontal deflection circuit the sweep amplitude is adjusted to the desired level. The yoke is removed from the neck of the Kinescope and is placed over the pick-up coil assembly (with the 10 μh coil in place). The vertical sweep is turned off to prevent the introduction of a field frequency component into the pick-up coil voltage. Before the pick-up coil voltage is applied to the unit the bias of the 1852 is adjusted so that the d-c plate current is about 7 ma. Then the clipper tube bias is adjusted to produce a d-c plate current of 0.2-0.4 ma. The coil voltage is then applied to the input terminals of the unit, and by rotation of the yoke, the amplitude of the input voltage is adjusted to a level which just fails to cause a change in the 1852 plate current. The retrace interval peaks are positive in the input wave.

With the oscilloscope connected across the instrument's output terminals, the voltage wave will have the general appearance as shown in Figure 6. To locate the zero-velocity level on the oscilloscope screen it is necessary only to measure down from the clipping point an amount corresponding to I_1R_L, the zero-signal drop in the plate resistor of the clipper. A horizontal line drawn through this point defines the reference level for all relative slope or velocity measurements. The variations in sweep velocity are represented by the irregularities in the bottom portion of the clipped voltage wave.

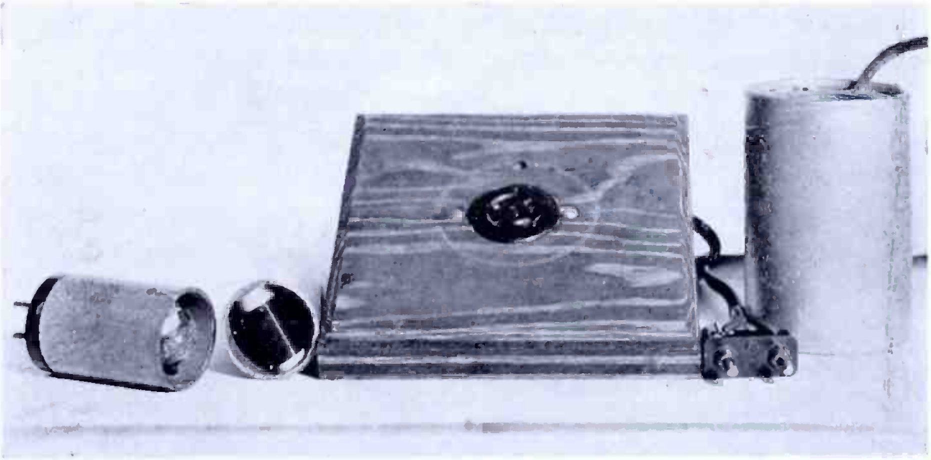

Fig. 8

As stated previously, the non-linear i_p-e_g characteristic of the clipper tube causes the voltage variations in its output load to be exaggerated somewhat during the scanning interval. This is of little significance and may actually aid in making velocity variations more apparent. If, however, accurate evaluation is necessary, it may be obtained by the use of a properly weighted, non-uniform scale laid against the oscilloscope screen.

In working with the device it will be found that changes in the deflection circuit which will produce almost unnoticeable changes in the coil-current wave form will appear in greatly magnified form on the differentiated wave. This is a useful feature, for, although the distribution may appear close to perfect on a current wave-form basis, or by inspection of a fixed pattern on the Kinescope screen, it is well to remember that the velocity variations are important in the event that moving scenes are reproduced, particularly if there is much localized horizontal motion. Furthermore, inspection of a fixed pattern may lead to erroneous conclusions, since the applied pattern may not have good distribution.

In adjusting the sweep linearity, either horizontally or vertically, one should strive for a purely rectangular differentiated wave. If the bottom of the wave is tilted it means a gradual overall departure from linearity, i.e., curvature in the sawtooth. Localized linearity variations will appear as localized ripples in the output of the linearity measuring unit.

Measurement of return time is made by determining the horizontal distance occupied on the oscilloscope screen by one period of the differentiated wave, and by taking the inverse ratio of this distance to the distance occupied by the return time, measured at the line corresponding to zero-sweep velocity. Under some conditions a certain amount of facility in measurement is afforded by using a sine wave for horizontal sweep in the oscilloscope.

The oscilloscope used in conjunction with this instrument must be capable of amplifying, without phase or frequency distortion, square waves at the line and field frequency. It is also necessary to have some means for calibrating the vertical deflection in terms of applied voltage, to permit the accurate determination of the zero-sweep velocity point.

One question which might arise involves the possibility of tapping across the deflection transformer secondary to obtain a differentiated voltage for linearity measurements, instead of using the mutual inductance method suggested here. If this is done, the voltage developed is not merely $L\dfrac{di}{dt}$, but, rather, $L\dfrac{di}{dt} + iR$, where L and R are respectively the effective inductance and resistance seen across the secondary of the output transformer, and i is the instantaneous yoke current. Consequently, even though the effective distribution is precisely uniform, the output voltage wave will not be purely rectangular, but will have a certain amount of slope during the scanning time. This is due to the passage of the sawtooth current through the effective resistance of the yoke and transformer circuit, as indicated by the iR term in the expression.

THE SERVICE RANGE OF FREQUENCY MODULATION

By

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Summary—An empirical ultra-high-frequency propagation formula is correlated with experimentally confirmed frequency modulation improvement factors to develop a formula for the determination of the relation between signal-noise ratio and distance in a frequency-modulation system. Formulas for the distance between the transmitter and the point of occurrence of the threshold of frequency-modulation improvement, as well as formulas for the signal-noise ratio occurring at this distance, are also developed.

Results of calculations and listening tests are described which evaluate the signal-noise ratio gain obtained by applying pre-emphasis to the higher modulation frequencies of both amplitude and frequency modulation systems.

Examples using a typical set of conditions are given and discussed.

IT HAS BEEN shown that the propagation characteristics of frequency modulation are such as to confine the use of this type of modulation insofar as telephony is concerned to the ultra-high frequencies which do not use the ionosphere as a transmission medium.[1] Consequently, for most purposes, calculations of service range of a frequency modulation transmitter may be based totally on the propagation characteristics of ultra-high-frequency waves and in this paper that base will be used. Formulas and empirical data now available make it possible to calculate the field strength at the receiver if transmitter power, antenna heights, and distance are known. Consequently, the signal-noise ratio for amplitude modulation may be determined if the field strength of the noise is known. From this known amplitude-modulation signal-noise ratio, the signal-noise ratio obtainable with a given frequency-modulation system may be determined by multiplying the corresponding amplitude-modulation signal-noise ratios by the frequency-modulation improvement factors as given in the author's previously published paper[2] on frequency-modulation noise characteristics. It is the purpose of this paper to perform this correlation of these propagation and frequency-modulation improvement formulas and develop formulas so that from the known constants of the frequency-modulation system, the signal-noise ratio at a given distance may be directly calculated.

Amplitude-Modulation Signal-Noise Ratio vs. Distance

The following empirical formula has been given by H. H. Beverage[3] for calculating the field strength when the receiver is within the optical distance of the transmitter:

$$E \text{ (r-m-s volts per meter)} = \frac{88\sqrt{W}\,ah}{\lambda D^2} \qquad (1)$$

where W = effective watts radiated = power in antenna times antenna power gain over a one-half wave dipole,
a = the receiving antenna height in meters,
h = the transmitter antenna height in meters,
D = the distance in meters,
λ = the wavelength in meters.

This formula is to be used for calculating the field strength for distances within the horizon only. For distances beyond the horizon, Beverage used a graphical method of plotting the curve of field strength versus distance. In this graphical method the field strength versus distance curve according to equation (1) was plotted for distances out to the horizon and then the curve was continued for distances beyond the horizon, but with a slope of $1/D^n$ instead of $1/D^2$. The exponent "n" was determined empirically and varies with frequency in the manner shown in Figure 1 which is reproduced from Beverage's paper.

In place of the graphical construction of the curve for distances beyond the horizon, the formula given by (1) may be revised to be applicable to all distances, whether they be within or outside of the horizon, as follows:

$$E \text{ (r-m-s volts per meter)} = \frac{88\sqrt{W}\,ah\,D_h^{\,n-2}}{\lambda D^n} \qquad (2)$$

in which the exponent "n" is equal to two for distances within the horizon and is chosen from the curve of Figure 1 for distances beyond the horizon. D_h is the distance to the horizon in meters and is equal to $2.21\sqrt{h} + 2.21\sqrt{a}$ where a and h are the receiving and transmitting heights in meters. Thus, where $D < D_h$, $n = 2$, and where $D > D_h$, n is taken from Figure 1.

When the units of the formula given by (2) are converted to feet, microvolts, miles, and megacycles, the formula becomes:

$$E \text{ (r-m-s microvolts per meter)} = \frac{0.01052\sqrt{W}\,ahf\,D_h^{\,n-2}}{D^n} \qquad (3)$$

where $W =$ effective watts radiated $=$ power in antenna times antenna power gain over a one-half wave dipole,

$a =$ receiving antenna height in feet,

$h =$ transmitting antenna height in feet,

$D =$ distance in miles,

$D_h =$ distance to the horizon in miles $= 1.22\ \sqrt{h} + 1.22\ \sqrt{a}$,

$f =$ frequency in megacycles.

In comparing the calculated curves with the experimental data, Beverage states that "scattering and absorption, even in open country,

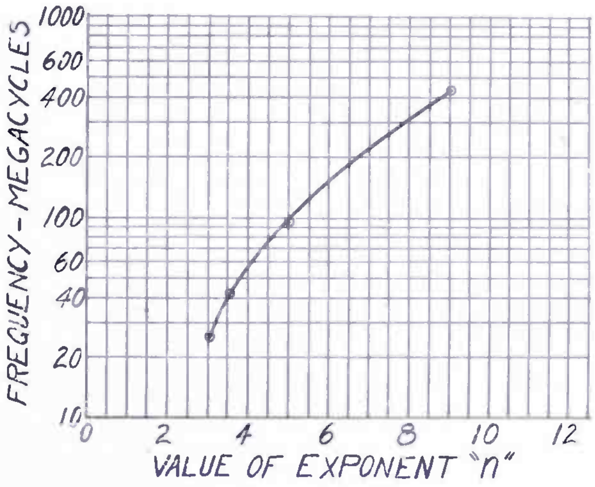

Fig. 1—Variation of exponent "n" in propagation formula when transmission is beyond the horizon.

tend to reduce the average intensity to something in the order of thirty to sixty per cent of the calculated value." In the following derivations an average experimental factor of forty-five per cent will be included to take into account this absorption and scattering.

It will be noted that this formula gives the average field intensity and does not take into account fading. More recent work by MacLean and Wickizer[4] shows the range of fading which may be expected for one set of transmission conditions. Thus, for transmission conditions reasonably close to the case treated by MacLean and Wickizer, the signal intensities at the fading minimums may be determined by applying a correction obtained from Figure 12 of the MacLean and Wickizer paper which is reproduced herewith as Figure 2. In the present paper, the formulas will be derived for the case of the average signal intensity and the fading correction will be applied to the examples given.

The peak carrier-noise ratio* obtained at a given distance may be determined by converting the voltage given by (3) into peak values (multiply by 1.414) and dividing by the peak noise voltage, N. N is the noise field strength as determined by means of a field-strength meter having a pass band characteristic equal to twice the band width of the audio spectrum which it is desired to receive. The experimental scattering and absorption factor may be taken into account by multi-

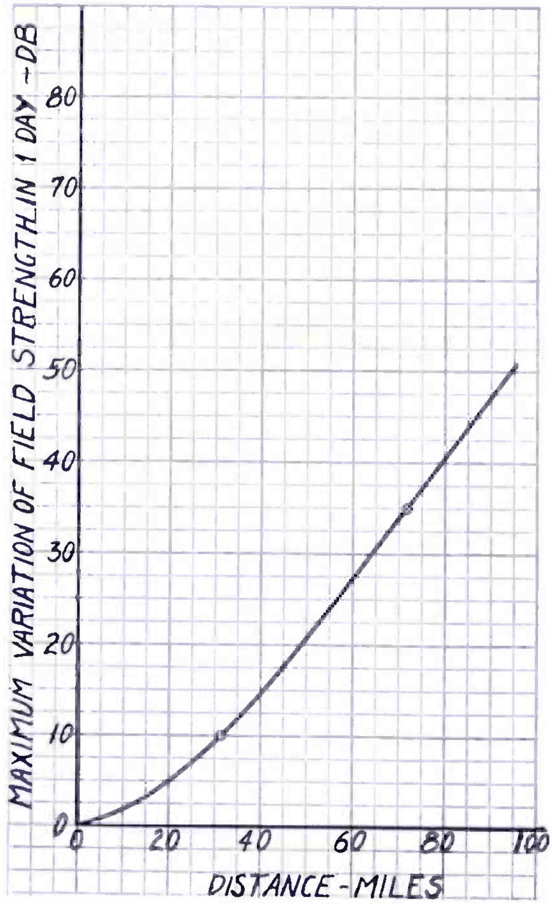

Fig. 2—Fading range of 50-megacycle transmission, transmitter antenna height = 1300 feet.

plying by 0.45. Thus, the amplitude-modulation peak carrier-noise ratio, which is approximately equal to the amplitude-modulation peak signal-noise ratio at the receiver output, for the case of one hundred per cent modulation, is given by:

* Throughout this paper carrier-noise ratio will refer to the ratio between the carrier and noise voltages as measured at the output of the intermediate-frequency channel of the receiver. Signal-noise ratio will refer to the ratio between the signal and noise voltages at the output terminals of the receiver. In an amplitude-modulation receiver the signal-noise ratio is usually substantially equal to the carrier-noise ratio, but in a frequency-modulation receiver the two quantities may differ greatly.

$$C/N = \frac{0.0067\sqrt{W}\,ahf\,D_h^{n-2}}{N\,D^n} = S_a/N_a \tag{4}$$

where C = peak carrier-field strength,
N = peak noise-field strength as above defined,
S_a = peak audio-signal voltage, amplitude-modulation receiver,
N_a = peak audio-noise voltage, amplitude-modulation receiver.

Frequency-Modulation Signal-Noise Ratio vs. Distance

The author's previously cited paper[2] on frequency-modulation noise characteristics gives the improvement factors effected by a frequency-modulation system over an amplitude-modulation system. For the case of fluctuation noise the factor is $\sqrt{3}$ times the deviation ratio μ. (The deviation ratio, μ, is equal to F_d/F_a where F_d is the peak-frequency deviation due to modulation and F_a is the width of the audio channel of the system.) This factor holds for equal carriers fed to the two receivers and for the condition of a peak carrier at least 4 decibels above the peak noise in the frequency-modulation receiver intermediate-frequency channel. Thus, the signal-noise ratio at the output of the frequency-modulation receiver may be found by multiplying the signal-noise ratio given by (4) by the improvement factor or:

$$S_f/N_f = \sqrt{3}\,\mu\frac{0.0067\sqrt{W}\,ahf\,D_h^{n-2}}{N\,D^n}$$

$$= \frac{0.0116\,\mu\sqrt{W}\,ahf\,D_h^{n-2}}{N\,D^n} \text{ (fluctuation noise)} \tag{5}$$

where S_f = peak audio-signal voltage, frequency-modulation receiver,
N_f = peak audio-noise voltage, frequency-modulation receiver.

When the received noise is impulse noise the improvement factor is equal to twice the deviation ratio or $2\,\mu$. The corresponding signal-noise ratio is then:

$$S_f/N_f = \frac{0.0134\,\mu\sqrt{W}\,ahf\,D_h^{n-2}}{N\,D^n} \text{ (impulse noise)} \tag{6}$$

It will be noted that in deriving the signal-noise ratio formulas of (5) and (6), the power gain normally effected by frequency modulation at the transmitter is automatically taken care of by the fact that W appears in both the amplitude- and frequency-modulation formulas. In the case of amplitude modulation the value of W used would normally be less than the corresponding value used in the frequency-modulation formulas by a factor equal to the frequency-modulation power gain at the transmitter.

Effect of the Improvement Threshold

The presence of the phenomena called the "improvement threshold" places a rather definite maximum service range on a frequency-modulation system. The improvement threshold occurs at the point where the peak voltages of the noise and carrier in the intermediate-frequency channel of the frequency-modulation receiver are equal. The experimental work of the author's previously cited paper[2] shows that the full frequency-modulation improvement for fluctuation noise is not obtained until the carrier is at least four decibels above the improvement threshold or where the carrier-noise ratio is about four decibels. The nature of the improvement threshold is such that the signal-noise ratio drops rapidly as the carrier falls below the four decibel carrier-noise ratio. For the case of fluctuation noise, which is of a continuous nature, the noise smothers the signal in a manner which has been described in detail before[2]. Consequently the distance at which the improvement threshold occurs for fluctuation noise may be taken as the maximum service range of the frequency-modulation system. However, for impulse noise of the type which consists of sharp impulses having a low rate of recurrence, the situation is somewhat different. It has been pointed out to the author by V. D. Landon of RCA Manufacturing Company, that for the condition of no modulation present, if the receiver is carefully tuned so that the incoming carrier is approximately synchronized with the oscillation frequency of the impulse, the frequency-modulation improvement is maintained for impulse noise which is stronger than the carrier. This effect is shown in the experimentally determined curve of Figure 3 which also shows the effect of a slight detuning from the point of synchronism. For these curves, the peak carrier-noise ratio and the peak signal-noise ratios were measured by means of an oscilloscope coupled to the intermediate-frequency channel output and the audio output of the receiver, respectively. Curve B was taken with the receiver carefully tuned to the impulse noise minimum. Curve A was taken with the carrier detuned about 20 per cent of the maximum frequency deviation of the receiver used. It will be noted that the improvement threshold manifests itself sharply when the receiver is detuned. Hence, for all except the very low passages of modulation the threshold would be present since application of frequency modulation corresponds to a momentary detuning. However, the reduction of the noise during the idle periods of the modulation is undoubtedly very helpful.

The curves of Figure 3 also show the value of carrier-noise ratio required to obtain the full frequency-modulation improvement for impulse noise. It can be seen that for all practical purposes it can be assumed that the full improvement is obtained at a peak carrier-noise ratio of unity.

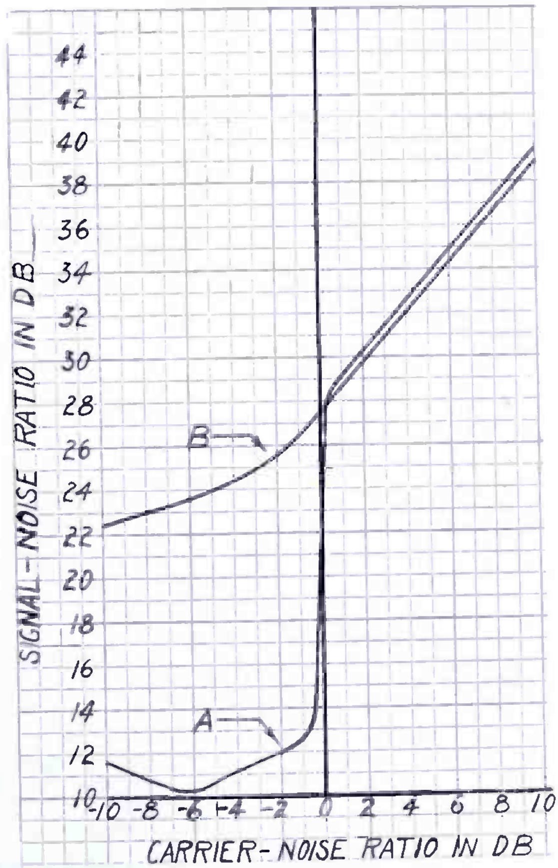

Fig. 3—Measured signal-noise ratio versus carrier-noise ratio characteristics with impulse noise. Receiver deviation ratio = 2.8.
Curve A = receiver detuned by an amount equal to 20 per cent of the maximum frequency deviation.
Curve B = receiver tuned for impulse noise minimum.

A further characteristic of the effect of impulse noise when it is stronger than the frequency-modulation carrier is the noise-silencing action which is present. This produces a sort of a minimum signal-noise ratio for carrier strengths below the improvement threshold when the carrier and pulse frequency are not synchronized. It has been shown[2] that this minimum signal-noise ratio is equal to the deviation ratio of the system. Thus, it may be seen that although the signal-noise ratio will become poorer below the improvement threshold, the signal may still be serviceable. On the other hand, if the impulse noise is of a continuous nature like fluctuation noise, the noise discriminates against the signal below the improvement threshold and, therefore, smothers the signal in the same manner that fluctuation noise does.

The formula for the distance at which the improvement threshold occurs may be obtained by developing a formula for the distance from the transmitter at which carrier-noise ratios of four and zero decibels are obtained in the intermediate-frequency channel of the frequency-modulation receiver for fluctuation and impulse noises respectively. First—the equation for the carrier-noise ratio in the frequency-modulation intermediate-frequency channel must be determined. This may be derived from (4) which gives the peak carrier-noise ratio as defined for an amplitude-modulation system. For a given radiated carrier, the carrier-noise ratio in the frequency-modulation receiver is less than that defined by (4) by a factor which depends upon the ratio of the effective band widths of the two receivers and upon the type of noise.

For the case of fluctuation noise,

$$C_o/N_o = C/N \times \frac{1}{(F_{fi}/2F_a)^{1/2}} \text{ (fluctuation noise)} \tag{7}$$

where C_o = carrier level in i-f amplifier of the frequency-modulation receiver,

N_o = noise level in i-f amplifier of the frequency-modulation receiver,

F_{fi} = i-f band width of frequency-modulation receiver,

F_a = band width of audio spectrum it is desired to receive.

For convenience let $Z_f = F_{fi}/2F_a$ for fluctuation noise. Substituting in (7), we obtain:

$$\frac{C_o}{N_o} = \frac{C}{N} \times \frac{1}{\sqrt{Z_f}} \text{ (fluctuation noise)} \tag{8}$$

The equation corresponding to (8) for impulse noise, which varies directly with the band width instead of as the square-root of the band width, will be:

$$\frac{C_o}{N_o} = \frac{C}{N} \frac{1}{Z_i} \text{ (impulse noise)} \tag{9}$$

where $Z_i = F_{fi}/2F_a$ for impulse noise.

The factors $\sqrt{Z_f}$ and Z_i, which are the ratios between one-half the intermediate-frequency channel width and the audio channel width of the frequency-modulation receiver for the two types of noise, are noise-determining ratios and, therefore, must be expressed in terms of *equivalent* channel widths. The equivalent channel widths are different for the two types of noise; hence, a separate factor is used for each type of noise. For fluctuation noise, the equivalent channel width is determined by dividing the area under the energy response curve (the selectivity curve plotted with the ordinates squared) by the height of the curve at resonance. This gives the width of the rectangular channel which would be equivalent to the actual round-topped channel. For impulse noise, the equivalent channel width is determined by dividing the area under the amplitude response curve by the height of the curve at resonance.

A further simplification of the ratios, Z_f and Z_i, may be effected by expressing them in terms of the deviation ratio, μ. Thus,

$$Z_f = K_f \mu \text{ (fluctuation noise)} \tag{10}$$

$$Z_i = K_i \mu \text{ (impulse noise)} \tag{11}$$

$$\text{where } K_f = F_{fi}/2F_d \text{ (fluctuation noise)} \tag{10a}$$

$$\text{and } K_i = F_{fi}/2F_d \text{ (impulse noise)} \tag{11a}$$

in which F_d = maximum applied frequency deviation.

The factors K_f and K_i express the ratio between the equivalent band widths of the intermediate-frequency channel and the total plus and minus peak frequency deviation of the frequency-modulation system. In other words they specify how far out on the intermediate-frequency selectivity curve the frequency deviation may be carried.

In order to determine the formula for the carrier-noise ratio in the frequency-modulation intermediate-frequency channel for the case of fluctuation noise, (4) and (10) may be substituted in (8) which gives:

$$\frac{C_o}{N_o} = \frac{0.0067\sqrt{W}\,ahf\,D_h^{\,n-2}}{N\,D^n\sqrt{K_f\mu}} \text{ (fluctuation noise, peak values)} \tag{12}$$

Likewise, for impulse noise (4) and (11) may be substituted in (9) which gives:

$$\frac{C_o}{N_o} = \frac{0.0067\sqrt{W}\,ahf\,D_h^{\,n-2}}{K_i\,\mu\,D^n\,N} \text{ (impulse noise, peak values)} \tag{13}$$

(12) and (13) give the carrier-noise ratio in the intermediate-frequency channel of the frequency-modulation receiver for the two types of noise. Since it is known that the improvement threshold

occurs when these carrier-noise ratios are equal to four and zero decibels, respectively, the distance at which the improvement threshold will occur for a given set of transmission conditions may be determined by equating (12) to 1.585 (four decibels) and (13) to unity (zero decibels), and solving for the distance. Thus,

$$D_i = \left(\frac{0.0042 \sqrt{W}\, ahf\, D_h^{n-2}}{N \sqrt{K_f \mu}} \right)^{1/n} \quad \text{(fluctuation noise)} \qquad (14)$$

$$D_i = \left(\frac{0.0067 \sqrt{W}\, ahf\, D_h^{n-2}}{N K_i \mu} \right)^{1/n} \quad \text{(impulse noise)} \qquad (15)$$

in which D_i indicates the distance at which the improvement threshold occurs.

A study of (14) and (15) shows the effect of a variation of the transmission conditions. For both types of noise the distance to the improvement threshold is directly proportional to the $1/2n$ power of the watts radiated. Hence, for the higher radiation frequencies, where the exponent "n" is large, the improvement threshold distance increases more slowly with increase in power. For fluctuation noise, the improvement threshold distance is inversely proportional to the $1/2n$ power of the product of the factor K_f and the deviation ratio. Thus, as the deviation ratio is increased, the improvement threshold distance decreases and decreases at a less rapid rate when it is beyond the horizon and for the higher radiation frequencies where the exponent "n" is larger. The importance of making the factor K_f as small as possible, by carrying the frequency deviation out on the selectivity curve as far as possible, is also indicated. When the noise is impulse noise (formula 15) the improvement threshold distance is inversely proportional to the "n"th root of the product of the factor K_i and the deviation ratio. Consequently, with this type of noise, the improvement threshold distance decreases more rapidly as these factors are increased.

The distances given by (14) and (15) may be substituted in (5) and (6) to find the frequency-modulation signal-noise ratio existing at the improvement threshold as follows:

$$S_f/N_f \text{ (at } D_i) = 1.585 \sqrt{3}\, \mu \sqrt{K_f \mu} = 2.74 \sqrt{K_f}\, \mu^{3/2} \quad \text{(fluctuation noise)} \qquad (16)$$

$$S_f/N_f \text{ (at } D_i) = 2\, \mu\, K_i\, \mu = 2\, K_i\, \mu^2 \quad \text{(impulse noise)} \qquad (17)$$

Effect of Pre-Emphasizing the Higher Modulation Frequencies

As has been pointed out for the case of amplitude modulation[5], the use of pre-emphasis circuit at the transmitter and a de-emphasis circuit at the receiver produces an overall gain in signal-noise ratio. This gain depends upon the fact that the higher modulation frequencies of voice and program material are of such a small amplitude that their accentuation does not increase the peak voltage of the applied modulating wave as much as the restoring circuit at the receiver reduces the noise. Thus, when the pre-emphasis circuit is inserted at the transmitter, the peak voltage of the modulating wave may increase somewhat so that the modulation level must be lowered, but this loss at the transmitter is overshadowed by the gain at the receiver. Consequently there are two quantities which must be evaluated—the loss at the transmitter and the gain at the receiver.

In order to determine the loss introduced by the pre-emphasis circuit at the transmitter, the following experimental observations were made: The equivalent of a two-string oscillograph was arranged by using two electronically switched amplifiers to feed the vertical plates of an oscilloscope. The outputs of the amplifiers were common, with one of the separate inputs being fed by program material through a pre-emphasis circuit. The gain of the two amplifiers was equalized at a low modulation frequency where the pre-emphasis circuit was not effective. The two amplifiers were alternately switched on by means of a 60-cycle square-wave form and the oscilloscope sweep circuit synchronized with the 60 cycles. The resulting pattern on the oscilloscope screen consisted of two segments of sweep one of which was actuated directly from the program material and the other through the pre-emphasis circuit. Hence, an accurate method of comparing the peak voltages of the two waves could be provided by inserting an attenuator in the pre-emphasized circuit and setting for equal peak voltages indicated by the two traces on the oscilloscope.

The observations made with the electronic-switch oscillograph covered all types of program material and were made for two audio-band widths of 5 and 12 kilocycles. The results of the observations indicated that in general the insertion of the pre-emphasis circuit increases the peak voltage of the level about 2.5 decibels for both the 5 and 12-kilocycle band widths. On certain material such as guitar, harmonica, and piano solos, the level is raised about 4.5 decibels total, but the occasion of such rises is rather infrequent and their duration very short. Hence, a permanent attenuation of 2.5 decibels might be inserted and the volume-limiting equipment relied upon to take care of the occasional higher peaks.

The action of the de-emphasis circuit in reducing the noise at the receiver is somewhat greater for the case of frequency modulation than it is for amplitude modulation. The reason for this is shown in Figures 4 and 5 which show the noise spectrums obtained in the output of the amplitude and frequency-modulation receivers, respectively, with and without the use of a de-emphasis circuit such as would be used with a pre-emphasizing circuit in accordance with R.M.A. Television Transmission Standard M9-218 at the transmitter. In the case of amplitude modulation, the normally evenly distributed noise is concentrated at the lower modulation frequencies. In the case of frequency modulation, the triangular noise spectrum is changed to a spectrum

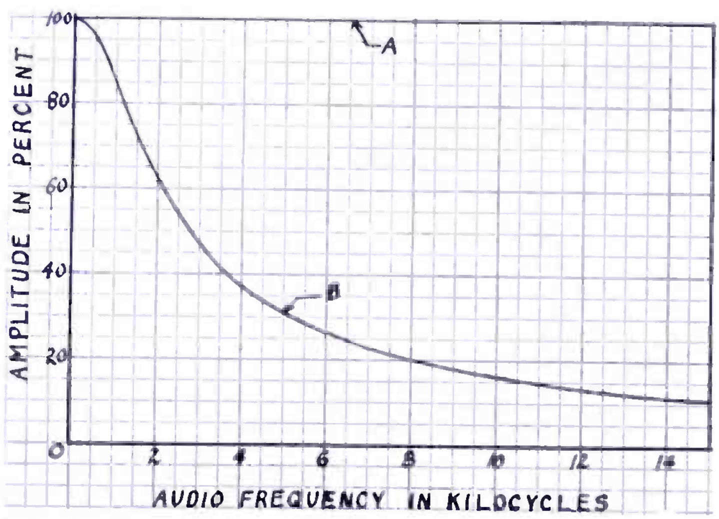

Fig. 4—Amplitude-modulation audio noise spectrums.
Curve A = no de-emphasis used.
Curve B = de-emphasis in accordance with R.M.A. Television Transmission Standard M9-218.

which is practically flat except for the falling off at the lower modulation frequencies.

In determining the relative figures of merit for the noise spectrums of Figures 4 and 5, or, in other words, the noise gains produced by the de-emphasis circuit, there are two determinations which are of importance. The first is the objective comparison which has to do with the relative strengths of the noise as would be measured on a meter. The second is the subjective comparison which takes into consideration the manner of utilizing the signal in the presence of the noise. For program or voice reception, the subjective comparison would be determined by a listening test.

The objective comparison of the noise spectrums of Figures 4 and 5 may be calculated for fluctuation noise by comparing the squared-ordinates areas of the spectrums. Such a comparison gives the ratio of the energies passed by the two spectrums; the root-mean-square voltage ratio is the square-root of this area ratio. The corresponding comparison for impulse noise may be calculated by comparing the areas of the spectrums. The ratio of the areas of the spectrums gives the peak voltage ratio directly for impulse noise. These areas may be obtained by integration or they may be plotted and a planimeter used. The curves of Figure 6 show the results of such a determination by the use of integration. These curves give the signal-noise ratio gain effected by the de-emphasis circuit alone. To obtain the overall gain due to pre-emphasis, the transmitter loss of 2.5 decibels must be subtracted from the values indicated by the curves.

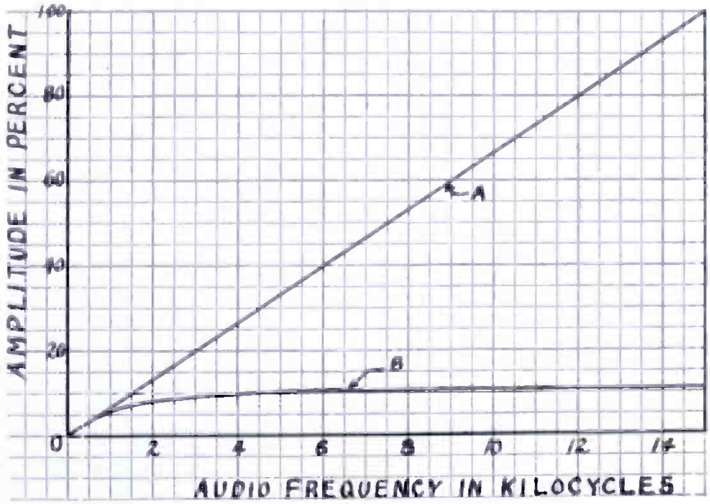

Fig. 5—Frequency-modulation audio noise spectrums.
Curve A = no de-emphasis used.
Curve B = de-emphasis in accordance with R.M.A. Television Transmission Standard M9-218.

In order to obtain the subjective gain effected by the de-emphasis circuit, listening tests were conducted in which the de-emphasis circuit was switched in and out and the annoyance effect of different types of noise compared. Both fluctuation noise and impulse noise were obtained from amplitude- and frequency-modulation receivers to mix with program to produce a signal-noise ratio. The de-emphasis circuit was then switched in and out of the noise while an attenuator was varied to balance the annoyance effect. Audio band widths of 5 and 12 kilocycles were used. The averaged observations of two observers indicated the rather unexpected result that for program and music reception, the subjective effect is practically the same as the objective effect. That is to say that the objective gains as portrayed by Figure 6 may also be taken as the subjective gains that would be realized in a listening test and the overall gain is obtained by subtracting the transmitter loss of 2.5 decibels from the values obtained from Figure 6.

For speech reception, where the noises were balanced for equal intelligibility, it was found that there was little or no gain effected by the de-emphasis circuit. For the case of amplitude-modulation noise, the use of pre-emphasis would apparently entail a slight loss in intelligibility when the 2.5 decibel transmitter loss was subtracted. With frequency-modulation noise, an intelligibility gain of a few decibels would be realized.

In connection with the listening tests another interesting observa-

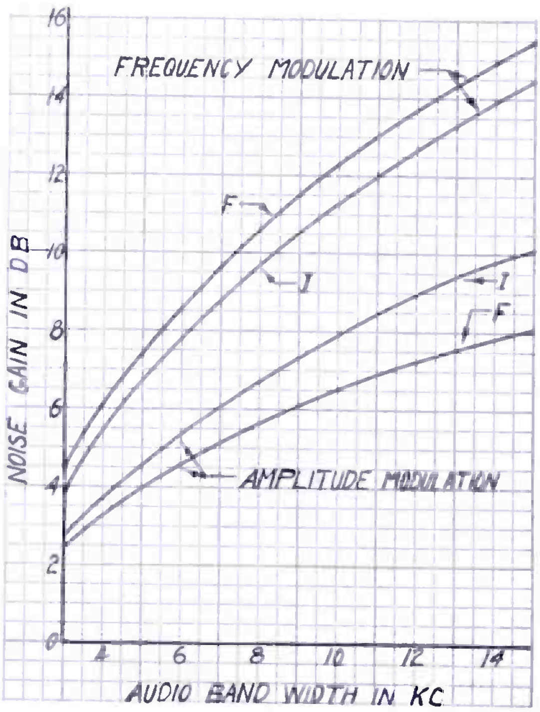

Fig. 6—Calculated noise gains effected by the de-emphasis circuit.
Curves F = fluctuation noise.
Curves I = impulse noise.

tion was made regarding the relative annoying effects of the triangular frequency-modulation noise spectrum and the rectangular amplitude-modulation noise spectrum using tube hiss as a noise source. In this test the noises were first balanced with a meter and then they were mixed with speech and the relative levels readjusted until they impaired the intelligibility of the speech the same amount. The averaged observations of three observers showed that about 8 decibels more noise could be tolerated with the triangular frequency-modulation noise than with the rectangular amplitude-modulation noise.

EXAMPLE

The curves of Figures 7, 8, 9, and 10 have been calculated for the following assumed transmission conditions for a high-fidelity broadcast system:

Transmitting antenna height	= 800 feet.
Receiving antenna height	= 30 feet.
Audio channel	= 15 kilocycles.
Frequency	= 42 megacycles.
Maximum frequency deviations	= 20 and 75 kilocycles.

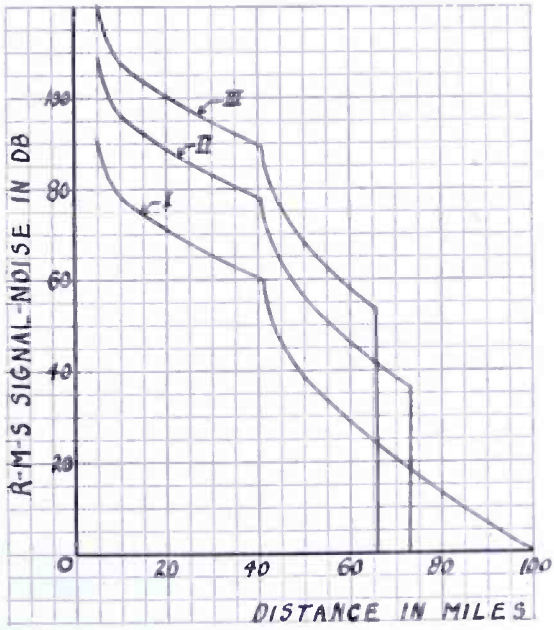

Fig. 7—R-M-S signal-noise ratio versus distance. Fluctuation noise = 1 peak microvolt per meter.
Curve I = amplitude modulation, 500 watts radiated.
Curve II = frequency modulation, 1000 watts radiated, maximum frequency deviation = 20 kilocycles.
Curve III = frequency modulation, 1000 watts radiated, maximum frequency deviation = 75 kilocycles.

Powers of one and 100 kilowatts radiated were used for frequency modulation and these powers were halved for the corresponding amplitude-modulation calculations. This two-to-one power gain effected at the frequency-modulation transmitter was taken instead of the usual four-to-one factor since it represents the gain that would be effected by a frequency-modulation system over the most efficient amplitude-modulation system which is the high-level type of modulation system. If the modulator tubes were paralleled with the final amplifier tubes in such a system, the increase in power would be two-to-one. The usual factor of four-to-one assumes the more inefficient types of low-level modulation.

The values of signal-noise ratio obtained for all of the curves were corrected by subtracting one-half the maximum fading range as portrayed by Figure 2. Formula (4) was used to calculate the average signal-noise ratio for amplitude modulation and the fading correction was applied to obtain the minimum signal-noise ratio. It was assumed that the transmission conditions of the fading correction curve were near enough to the conditions of this example to allow this direct correction without interpolation.

Where the propagation curves of the corresponding amplitude-modulation system were to be determined as well as those of the frequency-modulation systems, which was the case in this example, the simplest procedure was to draw the amplitude-modulation curves and then construct the frequency-modulation curves which follow lines parallel to the amplitude-modulation curve, but at higher signal-noise ratio levels for distances within the improvement threshold distances. To do this the average amplitude-modulation signal-noise ratios for fluctuation noise were calculated from (4). These were corrected for fading and the pre-emphasis gain of 5.6 decibels was added. For the low-deviation frequency-modulation system, the fluctuation noise gain is equal to 1.73 times the deviation ratio, μ ($\mu = 20/15 = 1.33$) or 7.2 decibels. The use of pre-emphasis adds another 7.4 decibels to the frequency-modulation gain as compared to the amplitude-modulation system with pre-emphasis since the total gain due to pre-emphasis on the frequency-modulation system is 13 decibels. The transmitter power gain adds another 3 decibels. Hence the curve for frequency modulation with a deviation of 20 kilocycles is $7.2 + 7.4 + 3 = 17.6$ decibels higher than the corresponding amplitude-modulation curve for the region within the threshold of improvement distance of the frequency-modulation system. The frequency-modulation system with a 75-kilocycle deviation has a larger deviation ratio ($\mu = 75/15 = 5$) so that its total gain over the amplitude-modulation system with pre-emphasis is $18.7 + 7.4 + 3 = 29.1$ decibels.

For the case of impulse noise, the frequency-modulation gains are equal to twice the deviation ratio and the pre-emphasis gains are also different. For amplitude modulation, the pre-emphasis gain is 7.5 decibels and for frequency modulation it is 4.5 decibels more or a total of 12 decibels. Hence, the gain of the frequency-modulation system using a 20-kilocycle deviation is $8.5 + 4.5 + 3 = 16$ decibels. The corresponding gain for the system using a deviation of 75 kilocycles is $20 + 4.5 + 3 = 27.5$ decibels.

In order to determine the signal-noise ratios at the improvement threshold distances, the gains due to pre-emphasis must be added to the signal-noise ratios calculated from (16) and (17). The pre-emphasis gains to be added in this case are the total gains of 13 and 12 decibels for fluctuation and impulse noise respectively. Before formulas(16) and (17) can be applied, the factors K_f and K_i must be evaluated. These factors are the ratios between the equivalent band widths of the frequency-modulation receiver intermediate-frequency channel and the total plus and minus frequency deviation. It is obvious that, in order to obtain the greatest distance to the improvement threshold, the equivalent band width must be made as small as the frequency deviation of the system will allow. The limitations encountered are the introduction of harmonic distortion and the introduction of amplitude modulation due to the frequency variation exceeding the flat-topped portion of the selectivity characteristic. In practice the limiter tends to take care of this departure from a flat-topped selectivity characteristic. Some rather preliminary measurements have indicated that the deviation may extend out to about 2.5 decibels down (down to 75 per cent) on the sides of the selectivity curve without producing harmonics which are too high for high-fidelity reception.

A study of typical intermediate-frequency equivalent band widths has shown that the band width 2.5 decibels down is about 90 per cent of the equivalent band width for fluctuation noise and about 78 per cent of that for impulse noise. The factors K_f and K_i are equal to the reciprocals of these percentages or 1.1 and 1.3, respectively.

The above evaluations of K_f and K_i are based on somewhat incomplete distortion measurements and assume no guard band to take care of tuning drift in the case of the frequency-modulation systems. Further work is undoubtedly necessary in these respects, but it is believed that this evaluation will serve the purposes of this paper.

With these evaluations of K_f and K_i, formulas (16) and (17) simplify to:

$$S_f/N_f \text{ (at } D_i) = 2.9\, \mu^{3/2} \text{ (fluctuation)} \tag{16a}$$

$$S_f/N_f \text{ (at } D_i) = 2.6\, \mu^{2} \quad \text{(impulse)} \tag{17a}$$

Applying (16a) to the system with a deviation of 20 kilocycles and adding the gain due to pre-emphasis of 13 decibels gives 25.9 decibels for the peak signal-noise ratio at the improvement threshold distance. The same calculation for the case of a 75-kilocycle deviation gives 43.2 decibels. For impulse noise (17a) is used and a pre-emphasis gain of 12 decibels is added to give ratios of 25.2 and 48.3 decibels for the two values of frequency deviation.

The curves of Figures 7 and 8 for fluctuation noise assume a thermionic agitation and tube-hiss noise level equivalent to one peak microvolt per meter (0.224 r-m-s microvolts per meter). This corresponds to an r-m-s noise voltage of 0.52 microvolts in series with a dummy antenna at the input terminals of the amplitude-modulation receiver. Such a noise voltage is about that which would be obtained with good design using an 1852 radio-frequency amplifier tube. All of the ratios for these curves were converted from peak to r-m-s ratios to correspond to the general practice in considering this type of noise. This was done by adding 10 decibels to the ratios. The figure of 10

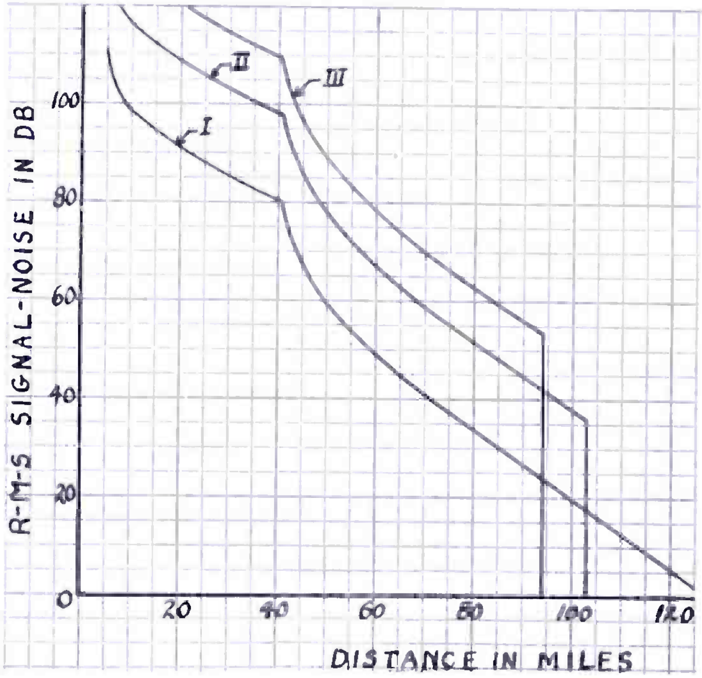

Fig. 8—Same as Fig. 7, but with amplitude modulation radiated power = 50 kilowatts, frequency-modulation radiated power = 100 kilowatts.

decibels is obtained from the data of the author's previous paper[2]. In that paper the crest factor of fluctuation noise is evaluated at 13 decibels. Subtracting the 3 decibel crest factor of the signal gives what might be termed the crest factor of the signal-noise ratio as 10 decibels.

The curves of Figures 9 and 10 for impulse noise assume a noise field intensity of 100 peak microvolts per meter for an effective band width of 30 kilocycles (audio band width = 15 kilocycles). This intensity is about that which would be received with horizontal polarization from the ignition system of the average automobile at a location about 125 feet from the road over which the automobile travels. The curves for this type of noise have been plotted with peak signal-noise ratios since root-mean-square values have little significance due to the very high and variable crest factor.

At distances beyond the improvement threshold for impulse noise, the curves are plotted for the condition of full modulation in which the improvement threshold manifests itself. Under this condition, the signal-noise ratio is limited to a value which is equal to the deviation ratio when pre-emphasis is not being used. It will be remembered that this ratio is what might be termed the "silencing" signal-noise ratio since the noise tends to punch holes in the signal, but when the noise pulses have a relatively slow rate of recurrence, this condition is quite

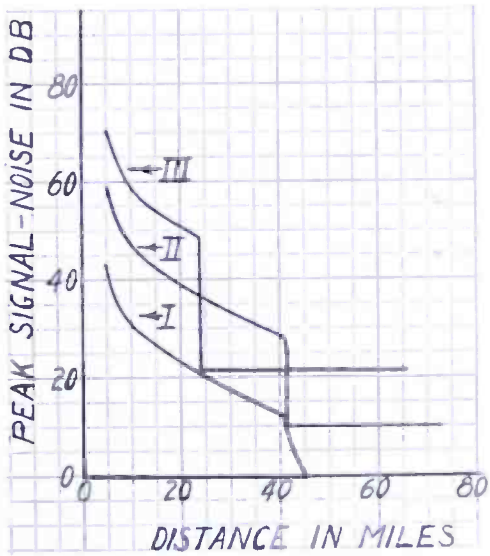

Fig. 9—Peak signal-noise ratio versus distance. Impulse noise = 100 peak microvolts per meter.
Curve I = amplitude modulation, 500 watts radiated.
Curve II = frequency modulation, 1000 watts radiated, maximum frequency deviation = 20 kilocycles.
Curve III = frequency modulation, 1000 watts radiated, maximum frequency deviation = 75 kilocycles.

tolerable. On the other hand, if the pulses have a high rate of recurrence, the noise tends to smother the signal and service is limited to the improvement threshold distance as in the case of fluctuation noise. When pre-emphasis is being used, a gain is added to this limited signal-noise ratio which is equal to the pre-emphasis gain of 7.5 decibels for amplitude-modulation noise. The pre-emphasis gain for amplitude modulation noise is used in this case since the noise in the silencing condition has lost its triangular spectrum characteristic of the frequency-modulation noise received above the improvement threshold. Thus, for the system with a 20-kilocycle deviation, the signal-noise ratio in the silencing condition beyond the improvement threshold distance is $2.5 + 7.5 = 10$ decibels. For the system with a 75-kilocycle deviation it is $14 + 7.5 = 21.5$ decibels.

At distances beyond the improvement threshold for impulse noise when the modulation is in the idle condition, the signal-noise ratios are somewhat higher than those shown by the curves due to the synchronization effect pointed out by V. D. Landon. This results in a considerable reduction of the annoyance effect of the noise. Hence, the curves may be taken as somewhat pessimistic, but satisfactory for comparison purposes.

Discussion

It is apparent from the curves that the maximum distance is served if the maximum frequency deviation is such that the minimum

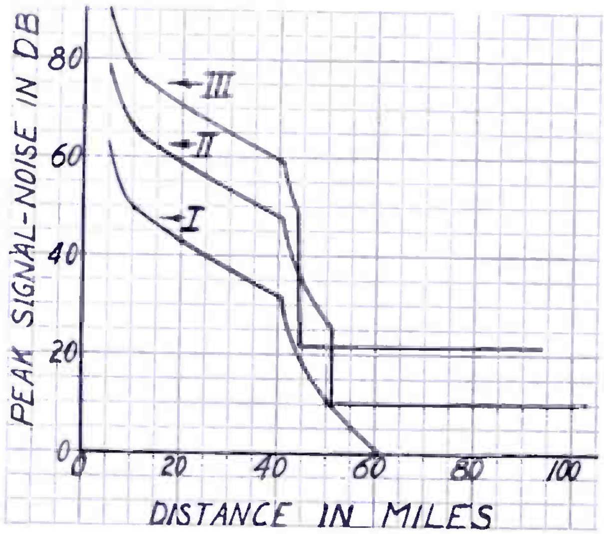

Fig. 10—Same as Figure 9, but with amplitude-modulation radiated power = 50 kilowatts, frequency-modulation radiated power = 100 kilowatts.

tolerable signal-noise ratio exists at the improvement threshold distance. Hence, the choice of the optimum deviation hinges on the definition of the minimum tolerable signal-noise ratio. If the figure of 30 decibels root-mean-square for fluctuation noise is taken as acceptable (this figure has appeared in the literature[6,7] as a commercially satisfactory signal-noise ratio), it is seen from Figures 7 and 8 that a maximum deviation of 20 kilocycles is more than adequate for all distances out to the improvement threshold distance since the lowest signal-noise ratio, which occurs at the improvement threshold distance for this type of noise, is 36 decibels. A deviation of less than 20 kilocycles is probably inadvisable since this would prevent the possible future transmission of an audio fidelity of zero to 20 kilocycles.

It is seen from Figures 7 and 8 for fluctuation noise that at the distance corresponding to the improvement threshold for a deviation of 75 kilocycles, a broadcasting service using a deviation of 20 kilocycles yields a signal-noise ratio of 42 decibels. For the same distance, a service using a 75-kilocycle deviation yields a signal-noise ratio of 53 decibels. At this point it is apparent that a 20-kilocycle deviation produces a signal-noise ratio which is comfortably above what would be considered commercially satisfactory, with an attendant conservation of band width in the available portion of the frequency spectrum. However, if it were assumed that a signal-noise ratio of 53 decibels is necessary, this ratio may be obtained with a deviation of 20 kilocycles at a somewhat shorter distance from the transmitter. For the conditions of the curves of Figures 7 and 8 this distance would be 20 per cent shorter. On the other hand, a 20-kilocycle deviation is capable of furnishing what would be considered better than acceptable service out to a distance which is about 10 per cent greater than the maximum service range for a 75-kilocycle deviation.

When the noise is impulse noise as portrayed by Figures 9 and 10, it can be seen that the general shape of the curves are similar to those for fluctuation noise. If the frequency of recurrence happens to be high so that the noise is continuous, the silencing properties of the frequency modulation cannot be taken advantage of. Instead, the noise, if stronger than the carrier, will depress the signal so as to smother it. For this type of noise a low-deviation service will have a greater range than a high-deviation service by an amount that is considerably larger than the corresponding case for fluctuation noise. This is especially true where the transmitter power is low enough to cause the improvement threshold to occur within the horizon as is the case with the conditions of Figure 9.

If the impulse noise happens to be automobile ignition where the rate of recurrence of the pulses is rather infrequent and the time duration of the impulses short, the noise-silencing properties of the frequency-modulation system may be taken advantage of when the noise is stronger than the carrier. With this type of noise, the effectiveness of the silencing action increases as the deviation is increased. However, for this type of noise, even a system with a low deviation produces a noise output which has a rather low annoyance value.

It is apparent that the service obtainable where fluctuation noise predominates, may be predicted with a fair degree of accuracy. However, owing to the highly variable character and distribution of impulse noise, predictions are difficult regarding this type of noise. Hence, it is felt that it is highly desirable to conduct field tests to further compare the results of different deviations under actual service conditions.

Conclusions

From the above considerations, it seems apparent that if fluctuation noise is the primary limitation, a frequency-modulation broadcasting system using a deviation of 20 kilocycles will produce a signal-noise ratio greater than the values normally considered acceptable, for all distances out to the distance at which the improvement threshold occurs. If the limitation is impulse noise a similar relationship exists, but it is felt that further field work is desirable to study the results under actual service conditions.

The effect of increasing the deviation of the system is to reduce the distance at which the threshold effect is realized, but at the same time the signal-noise ratio at distances equal to or less than that distance is improved.

The interests of band width conservation are, of course, best served by a choice of the lowest value of deviation that will yield an acceptable signal-noise ratio out to the threshold distance. It would also seem that the same choice will yield the desired service at the lowest cost.

In order to evaluate carefully these effects under a large number of actual service conditions, a series of field tests are being undertaken, using receivers designed for optimum performance at each of the different values of deviation under consideration.

Acknowledgment

The guidance and helpful suggestions of Mr. H. O. Peterson, and the assistance of Mr. R. E. Schock in the experimental work of this paper, are gratefully acknowledged.

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APPLICATION OF ABELIAN FINITE GROUP THEORY TO ELECTROMAGNETIC REFRACTION

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Summary—This paper presents the direct relationship between the group theory and indices of refraction encountered in the solution of optical problems. The properties and advantages of the group theory are made available by using the double-subscript notation when designating the indices of refraction. The notation is developed from fundamental concepts of refraction at a plane surface and applied to two specific refraction problems in the following order:

a. refraction in the Kennelly-Heavside layer (radio waves)
b. electronic refraction (electron optics)

The criterion to be used in determining the applicability of the method developed in this paper to any refraction problem is reduced to a simple equation.

I. Introduction

IN THE solution of electromagnetic refraction problems, such as those encountered in film recording, television and the refraction of radio waves, it is necessary to consider several indices of refraction lying in the path of propagation. Generally, the indices of refraction are different for adjacent media thereby requiring extreme care when the notation used in current literature is employed. By modifying the system of notation for the index of refraction, the inherent mathematical properties of the quantities involved become available to facilitate the required calculations. These mathematical properties follow immediately from the theory of Abelian finite groups.[1] The modified notation referred to is that which involves double subscripts and will be introduced by considering monochromatic light passing through different media. Several refraction problems will be analysed to illustrate the advantages of this notation.

A brief summary of the definition of an Abelian finite group is appropriate in this portion of the paper. Such a group is defined in the following manner:

Let G be a system of elements and R a rule of combination for uniting any pair of the elements in a given order. If a and b are two

[1] Finite Groups—H. Hilton—Oxford University, Clarendon Press.

elements of G, the combination or product is denoted by ab. The elements in the system G are said to form a group and the system is said to be a group if the following conditions are satisfied:

1. If a and b are elements of G, then ab is also an element of G.
2. If a, b, c, are elements of G, then $(ab)c = a(bc)$
3. If a is an element of G, there is an element a^1 of G, called the inverse of a such that $aa^1 = 1$
4. For an Abelian group $ab = ba$

In the discussion which follows, a correlation between the indices of refraction and the elements of a group will be made in detail.

II. Refraction at a Plane Surface

Assume that there are two different media in contact and in order to distinguish one from the other designate the first as medium 1 and the second as medium 2. It is known from experience that, as light passes from medium 1 to medium 2 the direction of propagation is altered and the results of investigation show that the ratio of the sine of the angle of incidence to the sine of the angle of refraction is a constant for all angles of incidence. The constant is designated as the index of refraction of medium 2 with respect to medium 1. This experimental phenomenon is explained on a theoretical basis by assuming the velocity of light changes abruptly as light passes from medium 1 to medium 2. The quantitative relationship between these velocities states that the ratio of the velocity of light in medium 1 to the velocity of light in medium 2, is equal to the index of refraction of medium 2 with respect to medium 1. In order to indicate that two media are involved in the specification of the index of refraction, write

$$\mu_{12} = \frac{V_1}{V_2} \tag{1}$$

where, $V_1 =$ velocity of light in medium 1.
$V_2 =$ velocity of light in medium 2.

μ_{12} is read as, "the index of refraction of medium 2 with respect to medium 1," meaning at the same time that light is passing from medium 1 to medium 2. The latter fact is indicated by the order in which the subscripts appear.

Now, reverse the direction of propagation of light and allow the light to pass from medium 2 to medium 1, and in accordance with the above notation.

$$\mu_{21} = \frac{V_2}{V_1} \tag{2}$$

From equation (1) and (2) the important relation becomes:

$$\mu_{12} = \frac{1}{\mu_{21}} \tag{3}$$

also

$$\mu_{12} \cdot \mu_{21} = 1 \tag{4}$$

Equation (3) stated in words is:

> the index of refraction of medium 2 with respect to medium 1 is equal to the reciprocal of the index of refraction of medium 1 with respect to medium 2.

The relation as expressed by equation (3) indicates that a quantity may be shifted to the denominator from the numerator or vice versa by merely interchanging the subscripts.

From equation (4), notice the following algebraic simplification which is one of the important features of the double-subscript notation. This simplification will be discussed in more detail when considering three different media.

$$\mu_{12} \cdot \mu_{21} = \mu_{11} = \frac{V_1}{V_1} = 1 \tag{5}$$

or

$$\mu_{21} \cdot \mu_{12} = \mu_{22} = \frac{V_2}{V_2} = 1 \tag{6}$$

Equation (5) may be interpreted geometrically by referring to Figure 1 where medium 2 has parallel plane surfaces.

The product of the two indices of refraction indicate the light is traveling from medium 1 to medium 2 and then from medium 2 to medium 1, the over-all index of refraction reducing to unity. This means there is no resultant deviation of the light ray or in other words the final ray is parallel to the original ray.

Now consider the more general problem of determining the index of refraction at a surface separating medium 2 and medium 3 with the individual indices of refraction given with respect to medium 1. First permit the light ray to pass from medium 2 to medium 3 as shown in Figure 2.

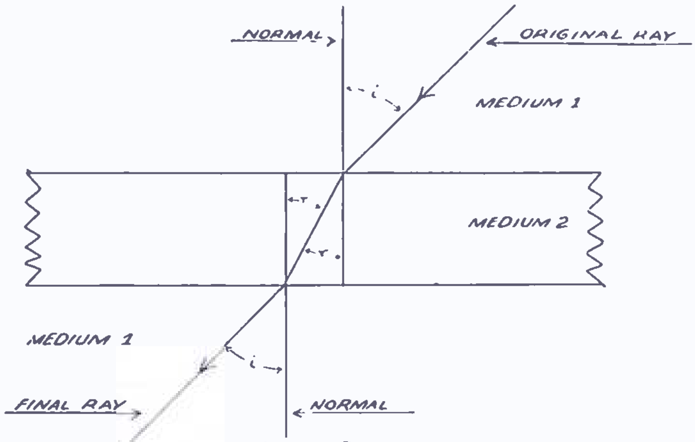

FIGURE 1

Using the double-subscript notation, the index of refraction becomes:

$$\mu_{23} = \frac{V_2}{V_3} \tag{7}$$

When numerator and denominator are divided by V_1,

$$\mu_{23} = \frac{\dfrac{V_2}{V_1}}{\dfrac{V_3}{V_1}} = \frac{\mu_{21}}{\mu_{31}} = \frac{\mu_{13}}{\mu_{12}} = \mu_{21} \cdot \mu_{13}$$

$$\mu_{23} = \frac{\mu_{13}}{\mu_{12}} = \frac{\sin i}{\sin r} \tag{8}$$

also $$\mu_{12} \cdot \mu_{23} = \mu_{13} \tag{9}$$

From equation (9) a very important rule of multiplication is obtained. Designating 1 and 3 as outer subscripts and 2 and 2 as inner subscripts, the rule is:

When multiplying one index of refraction by another, arrange the subscripts in such a manner that two inner subscripts are alike, thereby permitting the product to be substituted by a single index of refraction with its subscripts equal to and in the same order as the outer subscripts of the product.

Then,
$$\mu_{12} \cdot \mu_{23} \cdot \mu_{31} = 1$$

For a total of n media, the expression is:

$$\mu_{12} \cdot \mu_{23} \cdot \mu_{34} \cdots \mu_{n1} = 1 \qquad (10)$$

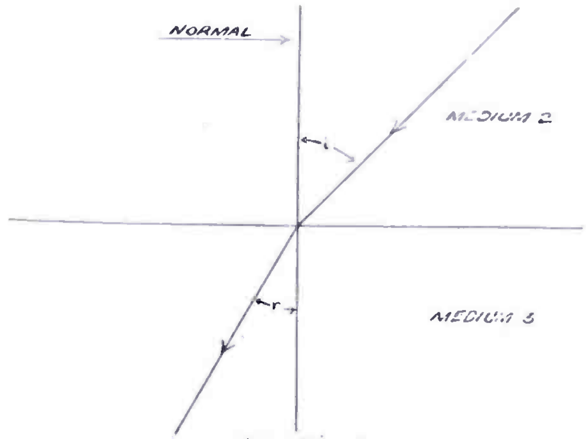

FIGURE 2

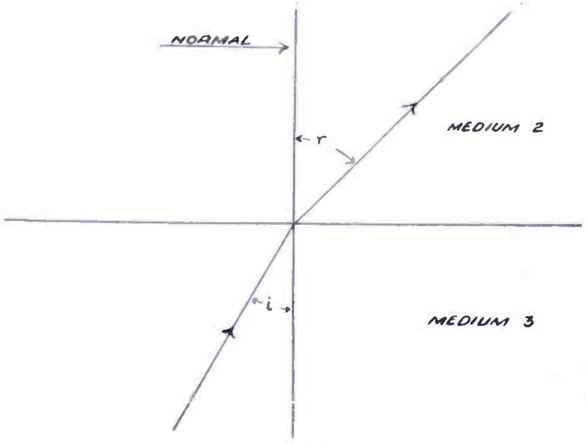

FIGURE 3

Equations (3) and (10) exemplify a unique property of Abelian finite groups.

As a second part of this problem, reverse the direction of propagation as shown in Figure 3 and compute the new index of refraction using the double-subscript notation.

The desired index of refraction is:

$$\mu_{32} = \frac{1}{\mu_{23}} \tag{11}$$

From equation (8),

$$\mu_{32} = \frac{1}{\mu_{23}} = \frac{\mu_{12}}{\mu_{13}} \tag{12}$$

Therefore, equations (8) and (12) are the solutions to the proposed problems involving media 2 and 3. If the product of the index of refraction μ_{32} and sin i is greater than unity, total reflection results.

III. Refraction in the Kennelly-Heaviside Layer

Refraction of radio waves in an ionized medium satisfies equation (10) and therefore is expressible in terms of this notation. The approximate formula for the index of refraction of radio waves in an ionized medium as given by P. O. Pederson[1] is:

$$\mu = \sqrt{1 - \frac{81\,N}{f^2}} \tag{13}$$

where, N = electron density in electrons per cubic centimeter
f = frequency in kilocycles

Consider a vacuum as medium 1 and two adjacent regions within the Kennelly-Heaviside Layer as medium 2 and 3 respectively, located such that the direction of propagation is from medium 2 to medium 3. Then in terms of the double-subscript notation, equation (13) is expressed as:

$$\mu_{12} = \sqrt{1 - \frac{81\,N_2}{f^2}} \tag{14}$$

$$\mu_{13} = \sqrt{1 - \frac{81\,N_3}{f^2}} \tag{15}$$

where, N_2 = electron density in medium 2
N_3 = electron density in medium 3

[1] "The Propagation of Radio Waves," P. O. Pederson, *G. E. C. Gad*, Copenhagen, Denmark.

From equation (8),

$$\mu_{23} = \frac{\sin i}{\sin r} = \sqrt{\frac{f^2 - 81\,N_3}{f^2 - 81\,N_2}} \tag{16}$$

which indicates total reflection within the Kennelly-Heaviside Layer occurs when,

$$\frac{\sin i}{\mu_{23}} = \mu_{32} \sin i > 1 \tag{17}$$

or when,

$$\sqrt{\frac{f^2 - 81\,N_2}{f^2 - 81\,N_3}} \cdot \sin i > 1 \tag{18}$$

and not necessarily at the region of maximum electron density.

IV. Electronic Refraction

Refraction of electrons when passing from one electro-static field to another also satisfies equation (10) and may be expressed in terms of the notation developed in this paper. This fact can be shown by considering a number of thin surfaces separating regions having electro-static potentials V_1, V_2, V_3, V_n respectively and determining the relation between the electron velocities and the corresponding potentials. When an electron is moving with a speed v_2 in region (2) before passing through a surface and a speed v_3 in region (3) after passing through the surface, then*

$$\frac{v_3}{v_2} = \sqrt{1 + \frac{(V_3 - V_2)e}{\frac{1}{2}mv_2^2}} \tag{19}$$

where e = electronic charge
m = electronic mass
v_2 = electron velocity corresponding to the potential V_2 influencing the electron

Then

$$\tfrac{1}{2}mv_2^2 = V_2 e \tag{20}$$

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* *Electron Optics in Television*—Maloff and Epstein (McGraw-Hill Book Co. Inc.).

Substituting equation (20) into (19),

$$\frac{v_3}{v_2} = \sqrt{\frac{V_3}{V_2}} \tag{21}$$

The index of refraction as defined by equation (7) becomes:

$$\mu_{23} = \frac{v_2}{v_3} = \sqrt{\frac{V_2}{V_3}} \tag{22}$$

By a similar process of reasoning concerning regions (3) and (4),

$$\mu_{34} = \sqrt{\frac{V_3}{V_4}} \tag{23}$$

Therefore, if μ_{24} were desired, it is only necessary to perform the product

$$\mu_{23} \cdot \mu_{34} = \mu_{24} \tag{24}$$

which gives,

$$\mu_{24} = \sqrt{\frac{V_2}{V_4}} \tag{25}$$

If the regions (1) and (2) are reasonably short, the two separating surfaces may be considered as a "thin" electro-static lens with a focal length of

$$f = \frac{4V_1}{\dfrac{V_2}{d_2} - \dfrac{V_1}{d_1}} \tag{26}$$

where d_1 = distance from zero potential surface to first separating surface

d_2 = distance between the two separating surfaces

Equation (26) may now be expressed in convenient form by dividing numerator and denominator by V_2 thus giving:

$$f = \frac{4\mu_{12}^2}{\dfrac{1}{d_2} - \dfrac{\mu_{12}^2}{d_1}} \tag{27}$$

or

$$\frac{1}{f} = \frac{1}{4\mu_{12}^2}\left(\frac{1}{d_2} - \frac{\mu_{12}^2}{d_1}\right) \tag{28}$$

The corresponding formula for the focal length of a "thin" glass lens in air is:

$$\frac{1}{f} = (\mu_{12} - 1)\left(\frac{1}{R_1} + \frac{1}{R_2}\right) \tag{29}$$

Where R_1 = radius of curvature of the first surface
R_2 = radius of curvature of the second surface

Thus, with the aid of equations (28) and (29) it is possible to compare the performance of "thin" electro-static lenses with "thin" glass lenses.

The foregoing analysis illustrates how the application of the group property clarifies the relations involved in refraction problems. It can be applied to any refraction problem which satisfies equation (10).

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