RCA REVIEW

A Quarterly Journal of Radio Progress

INDEX

Papers and Authors

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DEMONSTRATING BROADCAST FACSIMILE AT THE NEW YORK WORLD'S FAIR

BY

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BECAUSE of the tremendous publicity accorded television in recent years, many other developments in the communications art have been relegated to a somewhat obscure background, insofar as the general public is concerned. Two of the very important "other developments" are photoradio and facsimile reproduction.

These two techniques of transmitting and reproducing photographs and other printed matter are very similar in many respects, differing mainly in the form of the reproduced material. In photoradio the final product is usually a photographic film, while in broadcast facsimile reproduction it consists of printed paper.

A great many communications engineers have become aware of the rapid strides in this field, by means of the works of Ranger, Artzt, Young, and others. However, the public at large has had comparatively little information concerning this art. A striking illustration of this is the amazement and interest exhibited by the visitors who thronged the broadcast facsimile exhibit in the RCA Building at the New York World's Fair.

A complete broadcast facsimile system includes the following:

1. A scanner which converts the density variations in newsprint into corresponding variations in an electrical signal. In current practice, the output consists of an audio-frequency signal.
2. A broadcast transmitter modulated by the output of the scanner, and
3. A receiver which produces print from the facsimile signals in a manner analogous to the reproduction of sound in an ordinary radio receiver.

For the RCA facsimile demonstration system at the New York World's Fair it was deemed unnecessary to use a transmitter since wire lines were available for connecting the scanning equipment directly to the various receivers.

By means of eight receivers located in the RCA and the Missouri State Buildings, a continuous facsimile program was made available to Fair visitors. This program took the form of a radio newspaper, the *RCA Radio Press*, published with the cooperation of the *New York Herald-Tribune*. The *Radio Press* came out in three editions daily. The morning "World's Fair Edition" included a program of Fair events for the day and news of the various exhibits. The source for this information was the Fair Publicity Bureau. An afternoon and a night edition acquainted visitors with important national and international news, and often scored "scoops" over the New York

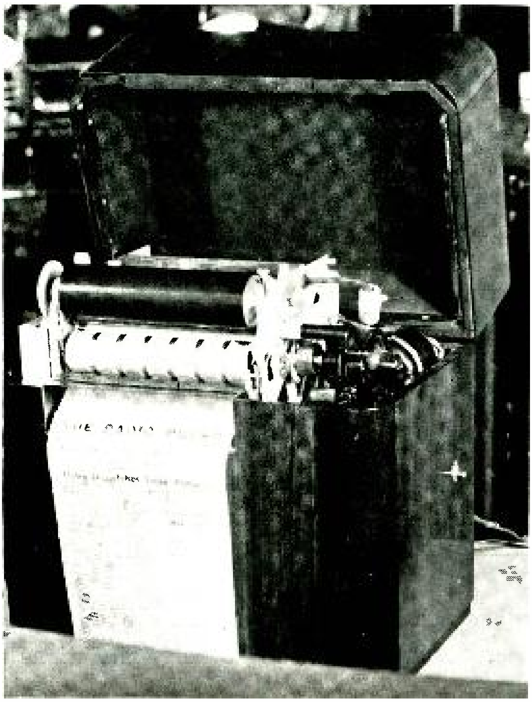

Fig. 1—Facsimile receiver for the home reproducing afternoon edition of *The Radio Press*.

daily papers. Figure 1 shows a facsimile receiver reproducing an afternoon edition of the *Radio Press*. One hour each day was devoted to the radio edition of the *St. Louis Post-Dispatch* run in conjunction with the *Post-Dispatch* exhibit in the Missouri State Building.

What is the sequence of events leading to the production of a facsimile newspaper? Let us take a brief trip through the facsimile exhibit. In the editorial office of the *Radio Press*, a constantly operating teletype machine records the latest front-page news, at the same time it is received at other newspaper offices throughout the country.

A member of the *New York Herald-Tribune* editorial staff chooses the news matter with the greatest reader interest and after editing and condensing it passes it on to an operator of the Coxhead Vari-typer.

This ingenious machine performs functions similar to those of a linotype unit in a newspaper plant, including that of automatically spacing the type to fit a line of print. The material is now in a form ready for assembly on a special 12-in. by 8.5-in. form. This form becomes the master-copy which the facsimile system will reproduce on the display receivers. The master-copy is comparable to ordinary typewriter copy produced with ten-point type.

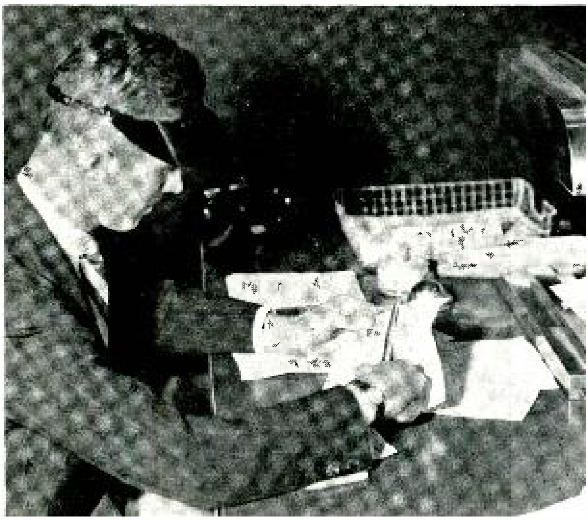

Fig. 2—*New York Herald Tribune* editor at work on an edition.

The facsimile scanner is shown in Figure 3. Scanning is accomplished by placing the master-copy on the cylindrical drum which is revolved before a light spot. The reflected light actuates a phototube and a signal is produced whose amplitude varies directly with the amount of light reflected from the copy. The scanner is a complete self-contained unit including the actual scanning apparatus; a timer; and three standard panel units—the compensating amplifier, the power-supply panel, and the voltage regulator. The output of the compensating amplifier consists of the carrier tone modulated by impulses from the scanning machine. In a complete system, this is the signal which modulates the radio-frequency carrier.

In the demonstration system, the output of the compensating amplifier entered two transmission lines. One line supplied signals to the various receivers on display in the RCA Exhibit Building. These included four home receivers and a special receiver for the horizontal *Post-Dispatch* copy, on display in the main facsimile exhibit; a single receiver on the second floor; and a special home receiver designed for the "living room of tomorrow." The other line terminated in a *St. Louis Post-Dispatch* exhibit receiver at the Missouri Building, nearly

Fig. 3—Facsimile scanner.

a mile away. The main facsimile exhibit in the RCA Building is illustrated by Figure 4.

The facsimile exhibit was extremely successful in serving three distinct purposes:

1. It was instrumental in introducing many visitors to their first view of home facsimile reproduction, and in acquainting people with the practicability of a broadcast system.
2. It supplied visitors with eagerly-awaited news on the history-making events as they occurred daily.

3. It served as a source of additional field data on the actual operation of the equipment in continuous service. The various units were on duty thirteen hours a day, seven days a week from April 30, 1939 to October 31, 1939, the close of the Fair's first year. Astonishingly little maintenance was required, and additional proof was obtained of its flexibility and simplicity of operation.

Fig. 4—A section of the facsimile exhibit showing scanner, "horizontal" receiver, and home receiver in operation.

Public reaction was on the whole very favorable. A great many people had never heard of this development. Among the popular questions asked were: "What is the range of reception?" "Will facsimile supplant our present daily newspapers?" and "How much will a home receiver cost?"

A DETERMINATION OF OPTIMUM NUMBER OF LINES IN A TELEVISION SYSTEM

BY

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Summary.—*In a television system horizontal resolution and vertical resolution are equally costly in video-band width. That is, on the basis of a fixed-band width, the vertical resolution increases with the number of scanning lines and the horizontal resolution decreases a corresponding amount—and vice versa. Also if a given increment of resolution is available for increasing the quality of a picture which originally has equal vertical and horizontal resolution, the quality will be improved more by applying the increment equally to the vertical and horizontal resolution than by applying it to improve resolution in only one direction. This conclusion follows from the known equality of the acuity of the human eye in various directions and the random orientation of the subject matter transmitted. Hence, the optimum utilization of the transmission band requires that the number of lines be near that number which provides equal horizontal and vertical resolution.*

In applying this criterion, the horizontal and vertical resolution is calculated in terms of the amount of blur with which the most elementary hypothetical test subject is reproduced on the receiving screen. The test subject is a single abrupt transition in brightness along the surface of the subject. The transition is placed nearly at right angles to the scanning lines for the analysis of vertical resolution and parallel to the lines for the analysis of horizontal resolution.

The immediate reduction of the visibility of the line structure justifies revising the number of lines slightly upward from the number determined by the criterion of "equal vertical and horizontal resolution" for the receivers of present practicable fidelity. As the fidelity of receivers is increased a corresponding improvement in shape of the scanning spot will reduce this particular need for more lines. Hence, it is possible to choose a number of lines which will provide nearly optimum picture quality for both present receivers and future improved receivers. This number of lines is between 441 and 507 (at 30 frames per second) based upon the reception of a maximum video-frequency signal of 4.5 Mc, which is available with the vestigial side-band method of transmission within the 6-Mc channel allocated for each television transmission (including sound).

(1) PREVIOUS ANALYSIS

EVEN in the early days of television experimentation it was tacitly assumed that the transmission band for the video signal should be wide enough to provide horizontal resolution equal to vertical resolution. (The expression "resolution" as used in this paper, refers to that useful characteristic of a picture which makes the picture sharp and clear as contrasted to blurred, "fuzzy" or smeared. It has usually been measured in terms of the distance between adjacent points or bars which can just be "resolved" or distinguished in the picture; but "resolution" as we apply the term is equally as important in pictures having no fine lines or "points," but having relatively large black, white or gray areas with sharp junctions between these areas.) The transmission band intended to meet this condition was calculated by the simple formula

$$f = N^2 \, r \, a/2 \tag{1}$$

where N = number of scanning lines
f = maximum video frequency in c.p.s.
a = picture-aspect ratio ($= 4/3$)
r = frame-repetition rate.

This formula neglects the return time of the scanning spot between lines and between frames, but corrective factors may be applied readily. It may be seen by inspection that (1) gives the highest frequency required to transmit a checker board pattern in which the width of each square is equal to the line pitch when it is necessary only that each square be reproduced as a dot without retaining the square shape. It was assumed in Eq. (1) that the vertical location of the checker board pattern was always such that the scanning lines coincided with the rows of squares.

In 1934, Kell, Bedford, and Trainer[1] recognized that useful television subject matter would have details that generally did not coincide with the scanning lines and that, therefore, the vertical resolution would depend upon the relative positions of the scanning lines with respect to the picture detail. They made observations using a complete television system to transmit a test pattern. The pattern consisted of a tapered wedge of near-horizontal alternate black and white bars which converged and thus occupied all positions with respect to the scanning lines[2]. The average of readings made by several observers indicated that 100 scanning lines were required to make 64 black and white bars distinguishable. Then, upon the assumption that in the finest checker board pattern resolvable the width of a square must be 1/0.64 times the line pitch, the frequency-band requirement was multiplied by a factor $K = 0.64$. This constant K has been widely accepted

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[1] "An Experimental Television System," R. D. Kell, A. V. Bedford and M. A. Trainer, *Proc. I.R.E.*, Vol. 22, p. 1247, No. 11, November 1934.

[2] Such a test pattern and its application are given in "A Figure of Merit for Television Performance," A. V. Bedford, *R.C.A. Review*, July 1938. When vertical and horizontal resolution are considered separately this test pattern may be considered equivalent to a continuous series of checker board patterns of different coarseness.

by television workers although there is not universal agreement as to its numerical value.

The determination of the band width by this method is open to criticism due to several inaccurate and incomplete hypotheses. First, the influence of the size, shape, and light distribution of the scanning aperture in both transmitter and receiver, all of which are known to affect resolution, has been omitted in the derivation of the formula. Some justification for this omission is the assumption that any reasonable variations in the spot would affect vertical and horizontal resolution equally unless the aperture attenuation is compensated. Compensation for aperture attenuation improves only the horizontal resolution and, hence, alters the relations for equal vertical and horizontal resolution.

Second, the most unsound assumption is that the transmission of a series of regularly spaced squares or bars is a measure of useful resolution. Actually, the subject matter usually transmitted will be represented by changes of light intensity along a scanning line that repeat at such intervals as to be substantially non-repeating so far as high-frequency behavior is concerned. After an abrupt change from black toward white, the next change is as likely to be toward white again instead of black which would be required to complete a single cycle. It is still less likely that the subject will contain repeating simple cycles which are identical.

Fidelity in transmission of the checker-board pattern is not a complete criterion of the capability of the system to transmit properly the most elementary "building block" with which picture detail is constructed, namely: an isolated abrupt change from black to white. This is evident from the fact that a good phase characteristic is not necessary for the transmission of a checker-board pattern. On the other hand, both theory and experiment show that reasonable linearity of phase is required for the proper transmission of isolated abrupt changes from black to white, or from white to black. Even if complete phase correction is assumed, there is still no assurance that the successful transmission of repeating dots in either a vertical or horizontal direction is an infallible indication of the ability of the system to transmit useful detail.

Wheeler and Loughren[3] have used the average reproduced width of a very narrow isolated white line as a criterion of useful resolution and have reached theoretical conclusions regarding the required band width. Here again, the criticism offered is that the unit for analysis is too specialized and is not the most elementary "building block" of

[3] "The Fine Structure of Television Images," H. A. Wheeler and A. V. Loughren, *Proc. I.R.E.*, Vol. 26, No. 5, May 1938.

which picture detail is constructed. The narrow line is a more basic unit than the repeating squares of the checker board, but a line represents *two* abrupt changes, black to white and then white to black.

(2) Reproduction of an Abrupt Change in Brightness as a Measure of Resolution

Almost all investigators[4] have used "equal vertical and horizontal resolution" as a criterion for adjusting the number of scanning lines to the band width. We also propose to use the criterion of equal vertical and horizontal resolution, but we prefer a different measure of resolution.

In the present theoretical investigation the most elementary building unit of which picture detail may be constructed has been adopted as a test unit. The fidelity of reproduction of the unit in the received picture is a measure of the resolution of the picture. This unit is an "abrupt change or discontinuity in intensity of illumination" along the surface of the subject to be transmitted. An analogous unit used in electric-circuit theory is the "Heaviside Unit Function." It may be represented as a sharp transition from black to white, from white to black, or from any intermediate shade to any other shade. For convenience in the analysis the transition is considered to occur from black to white. This causes no loss of generality because resolution is independent of polarity and amplitude[5]. In the reproduced picture at the receiver the change from black to white is not abrupt, but gradual. The distance along the picture screen required for the change from black to white to be effectively completed is a measure of the "blur" in the picture. The reciprocal of this distance is then a measure of useful vertical or horizontal resolution depending upon the angular position of the test transition. As will be seen the curves showing the surface illumination along the transition may have various irregular shapes depending upon the position of the transition from black to white with respect to the scanning lines, the spot size and shape, and the amplifier amplitude and phase characteristics. (The direction of the transition is considered to be at right angles to the junction or border dividing the black and white areas.) A comparison of these shapes is necessary in order to arrive at a significant relative evaluation of vertical and horizontal resolution.

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[4] "Channel Width and Resolving Power in Television Systems," J. C. Wilson, *Jour. Television Soc.* Vol. 2, No. 2, Part II, pp. 397-420, June 1938. An extensive bibliography is included.

[5] An exception occurs in the vestigial side-band systems of transmission where a slightly different shape of transient response wave occurs if the modulation is excessive. Some account of this condition will be taken later.

In useful television picture subjects the transitions to be transmitted will have many different shapes and degrees of abruptness. It is proper however to use the unit-function type of transition as a test unit for evaluating the response to all these transitions because any transition from black to white may be represented by a series of such Heaviside unit functions. If the system will respond properly to a unit function it will also respond faithfully to any wave shape. (Amplitude linearity of response is assumed.)

(3) Vertical Resolution

The determination of vertical resolution is based upon a test-picture subject in which the upper portion is black and the lower portion is

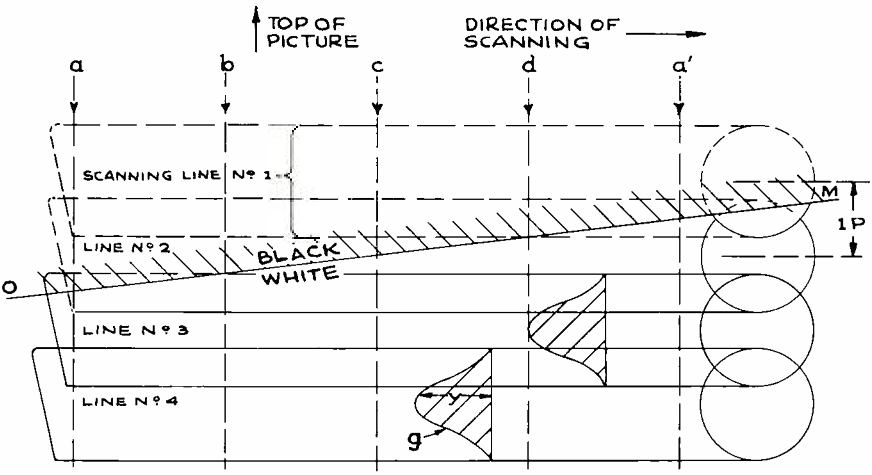

Fig. 1—A section of a scanning pattern showing the position of a black-to-white junction nearly parallel to the scanning lines. Representative transitions occur along *a*, *b*, *c* and *d*.

white with an infinitely abrupt transition between. The light intensity across the transition in the received picture will be a function of spot size in the transmitter, spot size and distribution in the receiver, distance between centers of adjacent scanning lines (that is, line pitch), and the position of the abrupt junction with respect to the scanning lines. Vertical resolution does not involve changes in intensity *along* the scanning lines and is, therefore, independent of the high-frequency response of the transmission system.

The distribution of intensity over the luminous spot of a simple cathode-ray tube has approximate circular symmetry, but is not of uniform intensity along its diameter. The intensity is greatest at the center of the spot and decreases toward the edge in consequence of aberrations in the focusing fields of the electron stream and the random initial velocities of electrons emitted from the cathode. The light distribution of the spot is such that the light intensity distribution *across* a single scanning line may be approximated by the cosine-squared function as shown in Figure 1. (This does not mean that the *intensity* along a diameter regarded as a function of the *distance* along the diameter is a cosine-squared.) The ordinates y of the curve g indicate the light intensity of the scanning line at various points across the line. The overlap of adjacent scanning lines has been assumed to be 50 per cent of the scanning-line pitch as shown. (The line pitch is the distance from center to center of the lines.)

The size of the scanning spot in the best cathode-ray receiving tubes available at the present time varies considerably with modulation of light output from minimum to maximum useful values. Also the intensity distribution within the spot changes with modulation. A constant spot is assumed for the present analysis, but the results may be properly applied to a system including a variable spot by considering the assumed constant spot to be the effective mean spot of the system. In the cathode-ray pick-up devices the scanning spot is not modulated and, hence, the assumption of a constant spot at the transmitter is entirely correct.

The black-to-white junction in the test subject is indicated by the line *OM* in Figure 1. *OM* is drawn nearly, but not quite parallel to the scanning lines in order that the junction will fall at every possible position with respect to the scanning lines. The various transitions from black to white, for example along the broken lines a, b, c, and d, are substantially vertical and, hence, are essentially measures of vertical resolution. If the junction line *OM* had been drawn exactly parallel to the scanning lines, the resulting transition would be very critical to vertical position with respect to the scanning lines and the test would lose its significance unless a variety of vertical positions and a mean of the several transitions were used. An easier method is to use the nearly parallel test junction. This not only is permissible, but is desirable because the exactly parallel test junction would represent a very special case of the subject matter.

Curve c in Figure 2 shows the calculated light intensity at the receiving screen that corresponds to the variation along the line c of Figure 1. Identical cosine-squared spots are assumed in the pickup device and the receiver. In Figure 2 the abscissa is the vertical distance along the receiving screen. The unit of distance is the scanning-line pitch. The dotted curves c_1, c_2, and c_3 show the contribution of the individual scanning lines to form the sum curve c. (Incidentally, curve c shows that the cosine-squared spot with 50 per cent overlap does not produce a flat field—one in which the scanning lines are indistinguish-

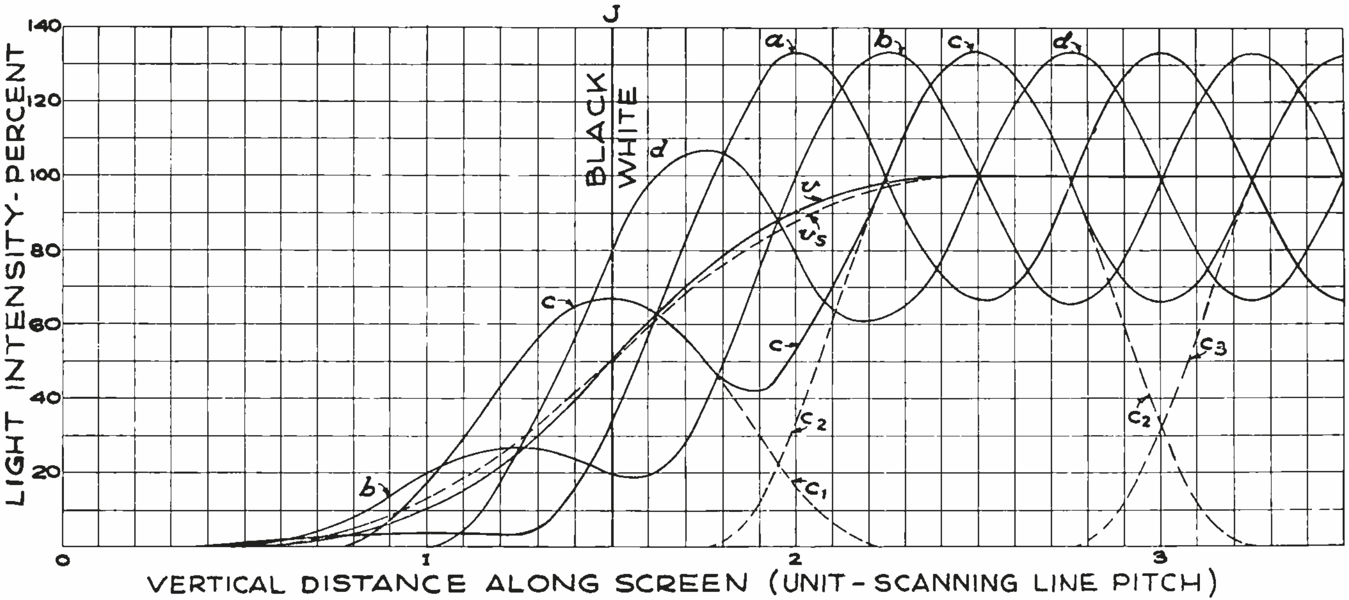

Fig. 2—Curves a, b, c, and d are plots of light intensity along a, b, c and d of Figure 1. The steepness of rise of the mean curve v is a measure of vertical resolution.

able—because the light intensity at the center of a line is twice as great as at the mid-points.)

Similarly, curves *a*, *b*, and *d* of Figure 2 show the light intensities in the transitions along the lines *a*, *b*, and *d*, respectively, of Figure 1. The relative position of the junction in the test subject is marked *J*.

Since the curves *a*, *b*, *c*, and *d* have different rates of rise it is evident that the resolution of a vertical transition depends upon the position of the test subject with respect to the scanning lines at the transmitter. It is natural to seek a single value of effective vertical resolution that will agree with an observer's impression of the vertical resolution of which a given television system is capable. Such a quest has promise of success in advance because common television experience shows that the observer does not separately scrutinize every vertical transition in a complex television picture. However, he does form an impression of the sharpness of an outline of a black, gray, or white area in a subject that depends upon the mean distribution of intensity along the outline.

It is reasonable to assume that the eye will tend to obtain an impression of the net sharpness of the junction *OM* that corresponds to a mean transition curve or arithmetic-average of all the transition curves which may occur across the junction. The basic assumption is that in effect the eye integrates the light intensities in elemental areas parallel to the junction *OM*. In addition the use of the "arithmetic-average" mean transition curve for this purpose is supported by the fact that in most real television subjects there is at least the minute amount of motion required to cause the scanning lines to intersect the junction at continually changing points. Thereby the well-known ability of the eye to integrate light values with respect to *time* effectively contributes to the effect which was at first assigned to the optical integration of elemental areas along a stationary junction.

All conditions at *a'* and *a* (Figure 1) are identical and the transitions between *a* and *a'* change smoothly and gradually. Then since *b*, *c*, and *d* are equally spaced between *a* and *a'* the average of curves *a*, *b*, *c*, and *d* of Figure 2 is reasonably near the average of all the transitions across the junction *OM*. The curve *v* (Figure 2) is a plot of the average values of the ordinates of curves *a*, *b*, *c*, and *d* and will be used as the mean vertical transition curve in the following study.

(4) Horizontal Resolution

The next step is the determination of a horizontal transition curve for comparison with the mean vertical transition curve *v* in order to arrive finally at an economic choice of the number of scanning lines for the band width available for picture transmission.

The test subject used above must be rotated substantially 90 degrees so that the junction between the black and the white areas is at right angles to the scanning lines, and the transition from black to white is *along* the scanning lines. The sharpness of the transition at the receiver may now be limited by both the signal transmission band and by the well-known aperture attenuation occasioned by the use of finite scanning apertures.

According to well-established theory the variation of the light intensity along a scanning line due to use of a finite symmetrical scanning aperture may be derived by replacing the effect of the aperture by an imaginary one of infinitesimal length and passing the signal through a hypothetical electrical network that attenuates all of the

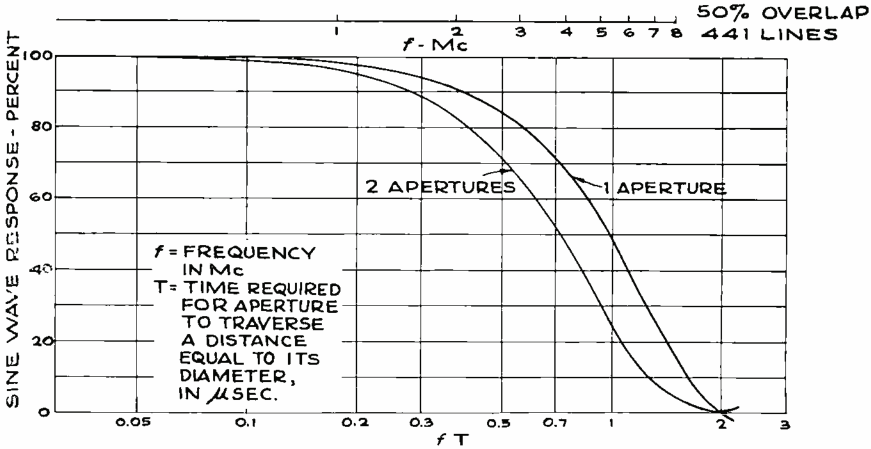

Fig. 3—Equivalent frequency response of a circular-scanning aperture having a cosine-squared transverse-light intensity.

frequencies in the picture signal by various prescribed amounts such as shown in Figure 3. This figure shows the calculated frequency characteristic of the cosine-squared aperture. The aperture introduces no phase distortion up to the frequency of zero response. The "*f*" abscissa scale applies only for a 441-line picture, a 50 per cent overlapped spot and a repetition rate of 30 frames per second. The "*fT*" scale may be applied to any system using a cosine-squared spot by inserting the proper value of *T* as defined in the figure. The frequency characteristic of the transmitting and receiving apertures may be compensated by correcting networks located in the transmission system, nearly up to the frequency at which the aperture response becomes zero. It will be seen after a determination of number of scanning lines has been made, that the required compensation for the aperture losses is reasonable to obtain for the video band available.

Present television-channel allocation requires that the picture transmitter and the accompanying sound transmitter operate within a 6-Mc

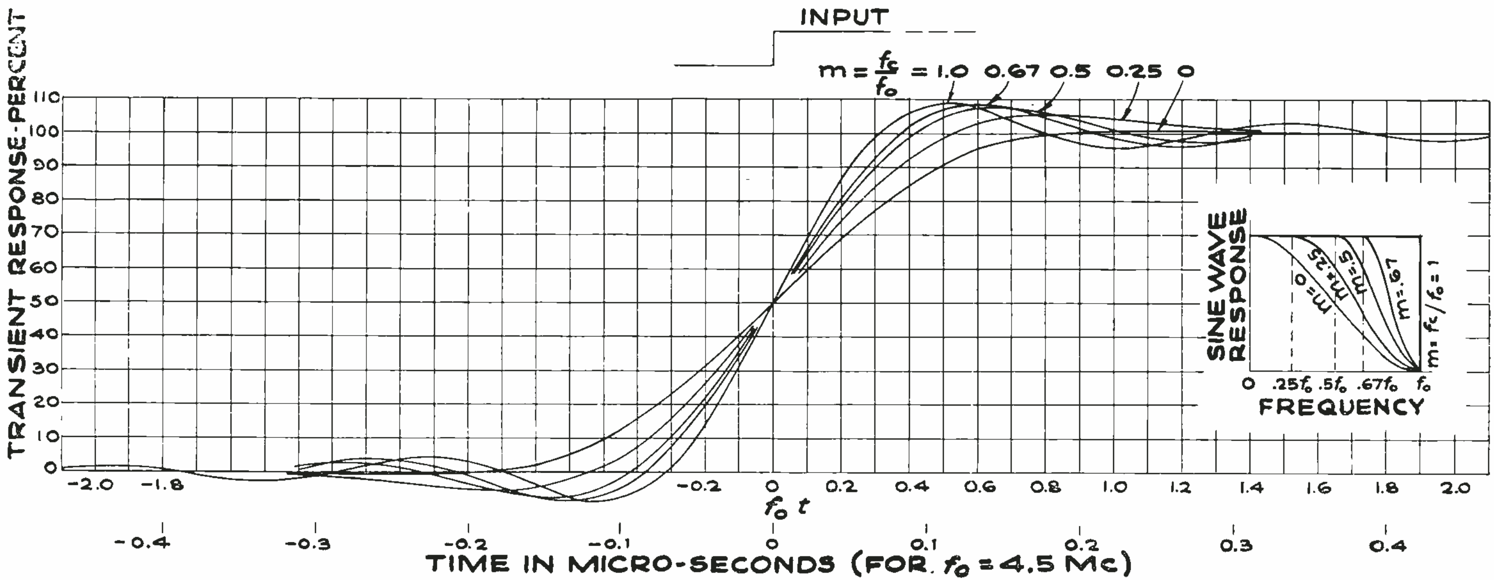

Fig. 4—Transient response of various idealized video-transmission systems having the different cut-off characteristics shown in the insert. Steepness of rise determines horizontal resolution.

band. Considerable experimental and theoretical work has determined that by the use of vestigial-side-band picture transmission the picture carrier may be located so that a maximum of 4.5 Mc video-band width may be received. The extent to which this 4.5-Mc video band is utilized for controlling the light intensity of the scanning spot in the receiver is determined by the receiver design. Probably most commercial receivers will fall materially short of ideal utilization due to the high cost of providing an overall frequency characteristic flat in amplitude and phase and having an abrupt cut-off. It is likely that cheaper receivers will have overall frequency characteristics which begin to fall at a relatively low frequency and very gradually approach zero response at 4.5 Mc. (The "over-all" characteristic is the equivalent v-f characteristic of the system which includes the transmitter and the effects of the r-f, i-f, and v-f characteristics of the receiver.) It is necessary to investigate the horizontal resolution provided by representative types of receivers characterized by certain overall frequency characteristics.

Figure 4 shows the calculated[6] transient response to a unit-function input wave for five different idealized frequency characteristics. (The input wave shown at the top of the figure is essentially the signal produced by scanning across the junction of our test subject when located in a vertical position.)

Each idealized frequency characteristic has uniform sine-wave response up to a frequency f_c beyond which the amplitude drops along a curve having a sine-wave shape to zero at f_o as shown in the insert of Figure 4. Each curve is identified by a different scale of m, the ratio f_c/f_o. The idealized characteristics have linear phase shift as evidenced by the symmetry of the transient response curves about the point of 50 per cent transient response. The origin of the abscissa scales was arbitrarily placed at the time of 50 per cent response. The generalized scale, $f_o\,t$, is applicable to systems having any value of f_o, expressed in megacycles when t is in microseconds. A specific time scale corresponding to $f_o = 4.5$ Mc is included, as our present interest is limited to systems having this maximum video frequency.

(5) Comparison of Vertical and Horizontal Resolution

It is well established theoretically that vertical resolution and horizontal resolution are equally costly in video-band width. In other words, the vertical resolution can be increased a reasonable amount by increasing the number of lines, but in consequence the horizontal resolution

[6] Several practicable methods for calculating this response are given by A. V. Bedford and G. L. Fredendall, in "Transient Response of Video-Frequency Amplifiers" *Proc. I.R.E.*, Vol. 27, No. 4, pp. 277-284, April 1939.

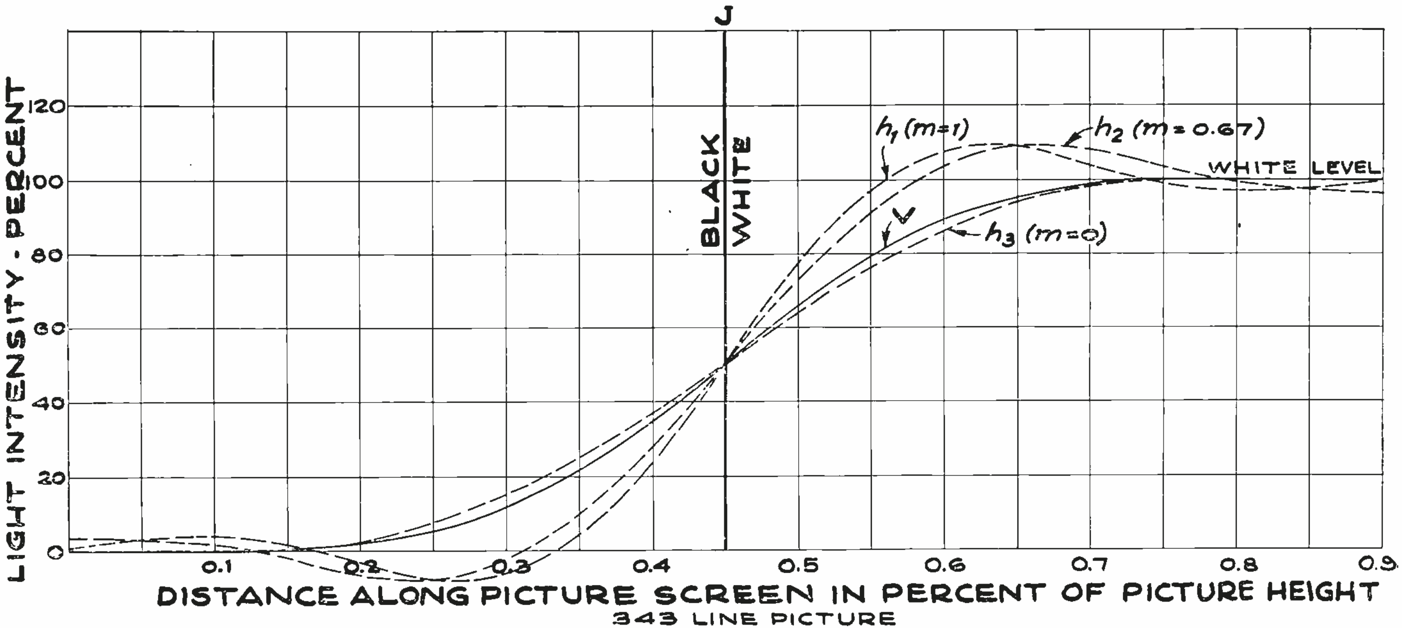

Fig. 5—Curve v of Figure 2 and several curves of Figure 4 are replotted here for comparison of the variable-light intensity vertically and horizontally along the receiving screen. 343 scanning lines and 30 frames per second.

must decrease a proportional amount if the band width and frame-repetition rate are fixed[7].

Therefore, the conditions for optimum use of the facilities available for a television service must include among other factors the choice of a number of scanning lines which will provide a favorable ratio of the vertical to the horizontal resolution. Without attempting at the moment to specify the ratio, we shall proceed to compare the vertical and horizontal resolution obtainable for different numbers of scanning lines.

Figure 5 shows the mean vertical transition curve v taken from Figure 2 and several horizontal transition curves h_1, h_2, and h_3, taken from Figure 4, replotted for a 343-line (30-frame-per-second) picture. The curves were made properly superimposable for comparison by using the abscissa scale which is common to both the vertical and horizontal transitions, namely, the distance along the picture screen[8]. The data for curves h_1, h_2, and h_3 in Figure 5 corresponding to receivers of different fidelity, were calculated from the transient response curves of Figure 4 where $f_a = 4.5$ Mc and $m = 1.0$, 0.67, and zero, respectively, the per cent response being considered directly equivalent to per cent light intensity produced by the kinescope since we have assumed the use of adequate compensation for the aperture losses[9]. The abscissas of curves h_1, h_2, and h_3 for Figure 5 were determined by letting a unit of time in Figure 4 become the distance traveled by the beam along the screen in the same unit of time. From a comparison of the rates of rise of curves h_1 and v of Figure 5 it is evident that for a 343-line picture the horizontal resolution theoretically obtainable is much better than the vertical resolution. In fact, even the relatively poor frequency characteristic $m = o$ as represented by h_3, would provide only about 10 per cent less horizontal than vertical resolution.

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[7] The two types of resolution also are related in apparatus cost since the cost of the amplifiers is a function of the v-f band width which is received and also because any steps taken to reduce the size of the scanning spot, such as operating at a higher anode voltage which increases the receiver cost, are apt to improve both vertical and horizontal resolution in the same order.

[8] The data for curve v of Figure 5 was obtained from curve v of Figure 2 by multiplying the abscissa by a factor 0.324, (because one line pitch of a 343-line picture is 100/309 per cent of the picture height) and then shifting the origin of the abscissa scale. Suitable allowance is made for loss of 10 per cent of the vertical lines due to vertical blanking and for 15 per cent loss of length of each scanning line in calculating these curves and those curves which follow involving horizontal transitions and horizontal resolution. An aspect ratio of 4/3 was used in all cases.

[9] The negative values of light intensity shown cannot exist as it means only that the kinescope is driven beyond cut-off. For transitions from *gray* to white the negative portions would be interpreted to mean another shade of gray.

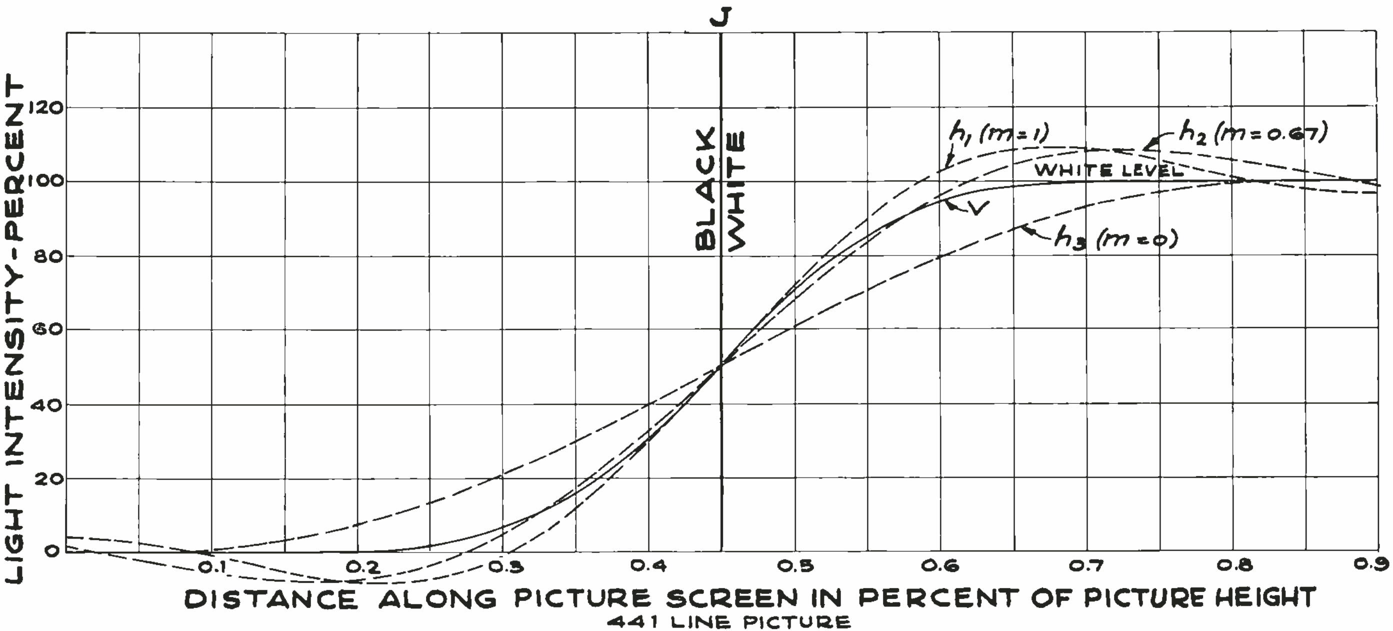

Fig. 6—Same as Figure 5 except for 441 lines.

In a similar manner Figures 6, 7, and 8 have been prepared to compare the transitions obtained in pictures having 441, 507, and 605 scanning lines, respectively. By inspection of these curves it can be seen readily that the frequency characteristics $m = 0.67$ provides about equal vertical and horizontal resolution for a 441-line picture and that the best frequency characteristic, $m = 1$, provides about equal vertical and horizontal resolution for a 507-line picture. Figure 8 shows that in a 605-line picture the vertical resolution is much better than the horizontal resolution provided by the best frequency characteristic, $m = 1$.

Before drawing conclusions as to the specific number of scanning lines for the greatest utilization of the available transmission bands, it is desirable to summarize the major observations drawn from inspection of Figures 5, 6, 7, and 8 in another set of curves.

In Figure 9, the vertical resolution V in arbitrary units has been plotted against the number of lines in the picture. Curve V is a straight line through the origin because the number of scanning lines determines the vertical resolution explicitly when the size of the scanning spot is constant in terms of the scanning-line pitch. The horizontal-resolution curve H_1 has been drawn with ordinates which vary inversely as the number of lines or such that the product of H_1 and V is constant. This relation between horizontal resolution and vertical resolution is valid for any other reasonable criterion for measuring resolution as well as the reproduction of an abrupt junction from black to white. Curve H_1 (for $m = 1$) was made to intersect curve V at that number of lines (507) for which the vertical resolution and the horizontal resolution are essentially equal as seen by inspection of Figure 7. This condition determines curve H_1 uniquely. Then we may say that the curves V and H_1 of Figure 9 are plots of vertical and horizontal resolution in the *same* arbitrary units when measured by the criterion of response to an abrupt junction.

The curve H_2 (for $m = 0.67$) is similar to H_1 except that the intersection with V is at 441 lines as indicated by the close agreement of curve h_2 and v in Figure 6. Curve H_3 also was drawn such that its ordinates vary inversely as the number of lines, but the value of H_3 was made about 10 per cent lower than V at 343 lines since by inspection of Figure 5, h_3 is about 10 per cent less steep than v for this number of lines.

(6) Picture Repetition Rate

Since the speed of the scanning spot along the scanning lines is proportional to the picture-repetition rate, the steepness of the curves h_1, h_2, and h_3 of Figures 5, 6, 7, and 8 and the ordinates of H_1, H_2, and

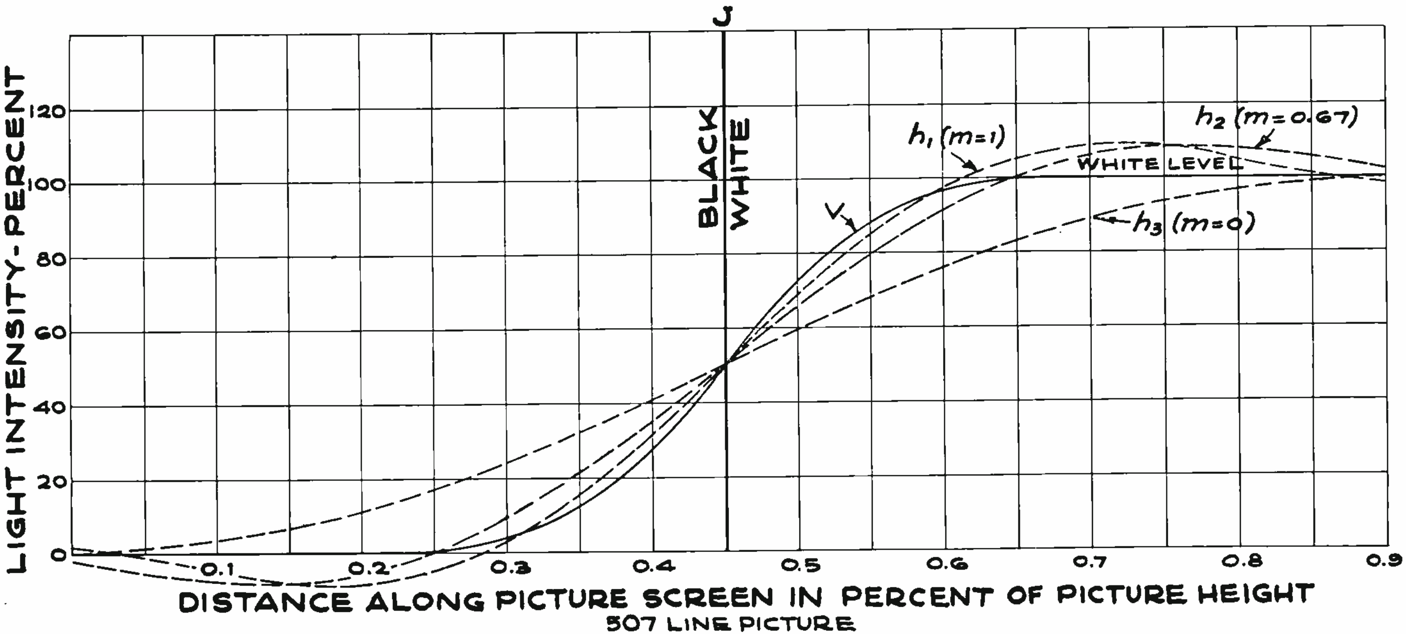

Fig. 7—Same as Figure 5 except for 507 lines.

H_3 of Figure 9 would vary inversely as the repetition rate. The net effect is that the number of lines which will provide a prescribed ratio of horizontal and vertical resolution will vary inversely as the *square root* of the picture-repetition rate.

The present study has been based upon the assumption of a frame frequency of 30 per second, interlaced. The corresponding field frequency is 60 per second, a number which has the advantage of dividing into the fundamental and harmonic frequencies of the prevalent 60-cycle power circuit a whole number of times. Let us briefly digress to review the possible advantages and disadvantages of several other frame frequencies.

A system using 24-frames and 48 interlaced fields per second appears attractive due to increased resolution or saving in frequency band. However, if the present study had been based upon 24 frames per second the suggested number of lines would be only 12 per cent higher. Unfortunately the 24-frame system is vulnerable to several serious defects. As a consequence of cross-talk in the transmitting equipment and the receiver, the kinescope beam is subject to small spurious modulation and deflection at 60- and 120-cycles. The beating with the 48-cycle field deflection causes adjacent fields to have different brightnesses and different positions on the screen. The result is 12- and 24-cycle flicker and motion in various areas of the scanning pattern as discussed in an earlier paper.* In addition to the usual types of cross-talk between electric circuits, the kinescope beam is deflected by the 60-cycle stray magnetic field of the power transformer.

By the use of additional electric filtering, electric shielding, magnetic shielding and careful transformer placement the cross-talk can be reduced to tolerable values, but the residual spurious deflection (though not readily detectable) will still reduce the theoretical vertical and horizontal resolution by at least a part of the theoretical gain. The 24-frame interlaced picture will also have appreciable inter-line flicker due to the low repetition rate (particularly for the brighter pictures) that will contribute further to eye fatigue.

One source of interference tending to cause flicker at a rate of 24 cycles per second that could not be avoided by the television engineer is the 120-cycle photo-electric pick-up by the camera from the 60-cycle light source generally present at the site of outside pick-up programs. The authors are convinced that the defects of a 24-frame system enumerated above appreciably outweigh the *theoretical* gain of only 12 per cent in horizontal and vertical resolution.

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* "Scanning Sequence and Repetition Rate of Television Images," R. D. Kell, A. V. Bedford and M. A. Trainer, *Proc. I.R.E.*, Vol. 24, No. 4, pp. 559-576, April 1936.

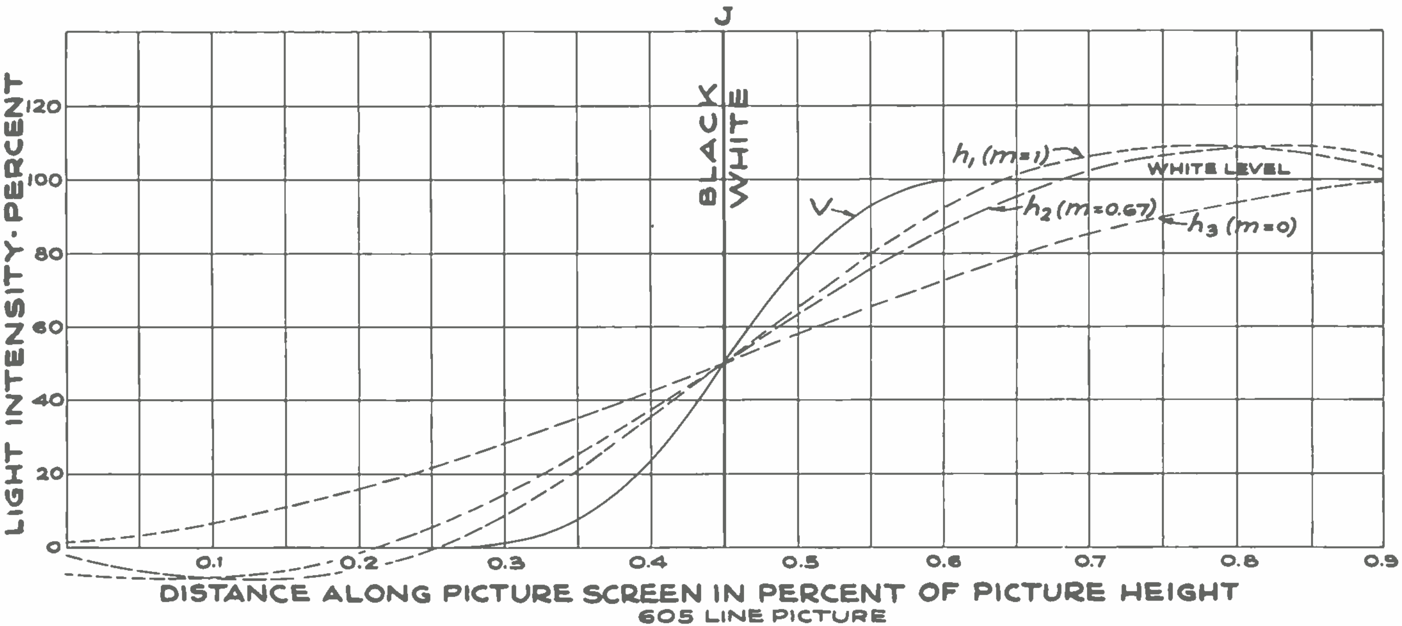

Fig. 8—Same as Figure 5 except for 605 lines.

A frame frequency of 15 per second has been seriously considered because it has the desired relation with the power frequency and because it theoretically would allow an increase of about 41 per cent in the vertical and the horizontal resolution within the available video band. Of course the low-field frequency of 30 would cause an intolerable amount of flicker unless sufficient light-storage were used in the receiver, such as conceivably might be provided by a suitable luminescent screen having long retentivity. However, if the screen material had a gradual rate of decay of light in common with the thousands of materials already reported, the picture would be badly smeared when the subject moved. On the other hand if at a future date an ideal picture-storage device should become available for providing uniform brightness for each entire frame period followed by an abrupt replacement by the next frame, flicker would be eliminated, but rapid motion in the subject would appear rather jerky. It is therefore very probable that a 30-frame-per-second interlaced system will afford a greater net service to the public.

(7) Choice of Number of Scanning Lines

It is clear that if the vertical resolution and horizontal resolution are *grossly* different, most of the excess value of the greater is lost due to the observer's tendency to choose a viewing position where his eyes, rather than the lower resolution, limit his realized resolution. At this viewing position it is evident that the observer obtains no appreciable benefit from the excess portion of the greater resolution. If the difference between the two resolutions is relatively small, say 20 per cent, the excess is not entirely lost in every case, because it is known that observers do not or can not generally adjust their viewing distance so critically that all of the excess of the higher resolution is unappreciated. Nevertheless, the various viewing distances used will tend to vary about the position where the eye limits the realized resolution. Hence, even when the excess of one resolution is small, the benefit from the excess portion of resolution will be less than from a similar-valued portion of the lesser resolution.

Generally the subjects to be transmitted by a television system for domestic service will be varied and will require substantially equal vertical and horizontal resolution for satisfactory results. Then upon the assumption that the vertical and horizontal borders and outlines of objects and areas in the picture are of equal value in defining the objects and areas, any increase of definition of either vertical or horizontal borders or outlines at the expense of the other definition must detract from the completeness with which the object is defined as a whole.[10] Therefore, from the point of view of resolution only, the number of lines should be such as to provide "equal vertical and horizontal resolution." However, there are several other qualifying and limiting conditions affecting the choice of number of scanning lines which will be discussed; but for the moment we shall assume that a condition of equal vertical and horizontal resolution makes optimum use of available facilities and determines the number of lines.

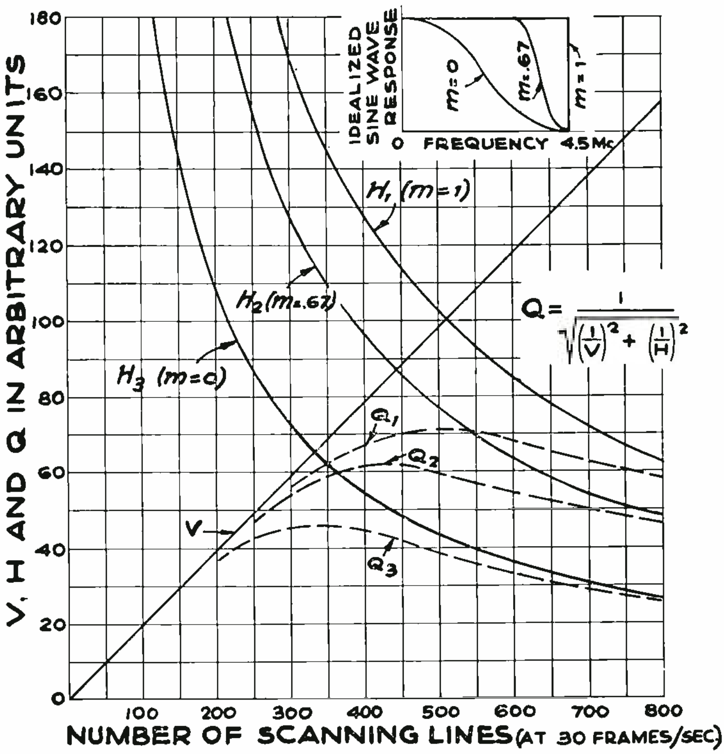

Fig. 9—The V and H-curves are vertical and horizontal resolution, respectively, provided by receivers having the idealized over-all frequency characteristics shown in the insert. The Q-curves are intended to indicate the corresponding picture quality as a function of the number of scanning lines.

By application of only the criterion of "equal resolutions" the optimum number of lines may still be 507, 441, 290, or some other value depending upon the response characteristic of the typical receiver

[10] Certain tests made in a recording "facsimile" system indicate that a typed page has maximum legibility when the horizontal resolution of the system is somewhat greater than the vertical resolution. This is due to the predominance of vertical strokes in the average typed page and the close spacing of the letters along the typed line compared with the line-to-line spacing. In order to accommodate many such transmissions most efficiently the vertical and horizontal resolutions are generally made unequal in facsimile systems.

assumed as seen by the points where V intersects H_1, H_2, and H_3 in Figure 9. It is expected that the fidelity of response of commercial receivers will vary with the price class of the receiver and with the future state of development of the art. Then since only a single number of scanning lines must be chosen to operate with all the types of receivers, it is desirable to know the relative quality of the pictures of receivers, it is desirable to know the relative quality of the pictures received by the different receivers for different numbers of lines.

Curve Q_1 (for $m = 1$) in Figure 9 is intended to express an index of picture quality as a function of the number of the lines and a fixed-video response. According to curve Q_1, the quality is reduced as the number of lines is varied above and below that corresponding to equal values of H_1 and V_1. Q is calculated from the formula

$$Q = \frac{1}{\sqrt{\left(\frac{1}{V}\right)^2 + \left(\frac{1}{H}\right)^2}} \tag{2}$$

The assumption is that the vertical blur, $1/V$ and the horizontal blur, $1/H_1$ may be regarded as vectors at right angles. The formula effectively adds these vectors and obtains the reciprocal of the sum, a number which appears to be related to the net resolution provided by V_1 and H_1. Admittedly, this procedure is arbitrary, but it has some logical support and has been partially verified as an equation of practical significance by a series of specific viewing tests.[11]

Curves Q_2 (for $m = 0.67$) and Q_3 (for $m = o$) were calculated by using the values of H_2 and H_3 instead of H_1 in Eq. (2). Each Q-curve is a maximum at the number of lines which provide equal vertical and horizontal resolution. Due to insufficient evidence presented here concerning the reliability of Eq. 2, the curves Q_1, Q_2, and Q_3 will be used only as a qualitative aid in visualizing how picture quality may be impaired when H and V are appreciably unequal. None of our conclusions will depend upon numerical values taken from the Q-curves.

There are several factors which tend to reduce the horizontal resolution of a receiver operating under practical receiving conditions:

(1) Multiple-path reception of the radio wave. (This depends upon terrain, obstacles, and reflectors such as buildings. No known remedy gives complete relief in many receiving locations so that

[11] In these tests a calibrated test pattern consisting of converging bars (see Reference 2) was reproduced by a television receiver having a fixed-band width with the test bars placed at 45° to the scanning lines, such as to indicate resolution in a diagonal direction. The number of scanning lines was changed in small steps over a wide range. The average values of readings by five observers were plotted and found to conform quite closely to curve Q_3 of Figure 9.

multiple-path reception is probably a permanent factor of consequence.)

(2) Accumulated-phase and amplitude errors in the transmitting and receiving system[12]. (This is likely to be appreciable particularly in chain programs in which the residual imperfections of equalization for the repeaters and links have an opportunity to "add up.")

(3) Imperfect aperture compensation. (Since aperture compensation requires an increase in the high-frequency response without the introduction of non-linear phase shift, rather complex correcting circuits are required. Due to cost considerations, a compromise solution is likely. Precompensation in the transmitter may offer a partial solution of this problem.)

(4) Vestigial side-band transmission. (Kell and Fredendall[13] and other writers have shown that inherent imperfections in the horizontal transitions occur, due to the absence of most of one of the side bands, even when only ideal filters are used. This effect is important only when high modulation is used.

It is impossible to evaluate accurately the effects enumerated above, but it is reasonably estimated that they may be on an average of such value that a receiver which is perfect for the available 4.5-Mc video band (as indicated in Figure 9 by $m = 1$) would provide actual horizontal resolution poorer than that shown by H_2 (for $m = 0.67$).

Considered only from the point of view of the overall-frequency characteristic, a receiver having the characteristic which we have indicated by $m = o$, is very poor compared with the measured response of a certain existing commercial receiver of good quality. However, the measured transient response of the commercial receiver proved to be very comparable in rate of rise to the theoretical transient response for the $m = o$ receiver, due to the phase distortion associated with circuits employed for rejecting adjacent channel interference. Nevertheless, this receiver demonstrated that it was capable of reproducing a satisfactory 441-line picture. It is to be expected that most commercial receivers will fall short of even the effective fidelity indicated by the curves for $m = 0.67$. In order not to penalize the receivers of this general fidelity excessively, the number of lines should not be much above 441 (see Figures 6 and 9).

It is known that as the number of scanning lines is increased the objectionableness of the line structure of the picture is reduced. Even though the limitation of resolution is generally subject to more serious criticism than the visibility of the line structure it should still be

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[12] See Reference 6.

[13] R. D. Kell and G. L. Fredendall "Selective Side-Band Transmission in Television." *RCA Review*, Vol. IV, No. 4, pp. 425-440, April 1940.

economical to shift the choice of the number of scanning lines slightly upward from that determined solely by considerations of resolution.

The line structure of the picture can be altered by changing the size, shape, and light distribution of the scanning spot. The circular receiving spot with cosine-squared distribution of intensity was chosen for this study because such a spot permits high-average light intensity in present practicable kinescopes. A size of spot giving 50 per cent overlap was chosen as a reasonable compromise between loss of resolution for a larger spot on one hand, and a loss of light and a more apparent line structure on the other hand.

A uniform rectangular spot with height equal to the line pitch may be considered ideal since it produces an entirely flat field and provides higher resolution than other configurations, such as a cosine-squared spot with *100 per cent* overlap[14], which is also known to produce a flat field. A vertical-transition curve v_s, for the rectangular spot, is plotted in Figure 2 (v_s corresponds to v which is for the cosine-square spot with 50 per cent overlap). Since the rate of rise of v_s is essentially the same as that of curve v the subsequent curves of Figures 4 to 9 would apply substantially as well for the rectangular spot.

It is of interest to note from Figure 7 that if the full possibilities of the 4.5 Mc video band (represented by $m = 1$) become realizable and the rectangular scanning spot becomes practicable—as results of future engineering development—the optimum number of lines would be about 507.

In view of the several considerations above, we conclude that the use of a number[15] of scanning lines between 441 and 507 at 30 frames per second allows optimum use of the channels available for a television-broadcast service in the United States. In reaching this conclusion we have satisfaction in the belief that adoption of the number of lines suggested will allow a nearly optimum performance of receivers of present commercial quality and at the same time will not penalize future receivers of improved quality.

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[14] Recommended by H. A. Wheeler and A. V. Loughren, Reference 3. A flat field is one in which no visible line structure is present.

[15] The number of lines should preferably be multiples of small odd whole numbers in order to facilitate construction of simple electronic synchronizing-signal generators. Numbers 441, 495, 507, 525, 539, 567, 605, 625, etc. satisfy this condition.

It is interesting to note that the number of scanning lines 507 results in $K = 0.85$ when inserted in equation (1) after revision as follows:

$$f = N^2 r a K (1 + t_H) (1 - t_V) / 2$$

where $t_H = 0.15$ = the fractional part of horizontal-sweep period for return time) and $t_V = 0.10$ = the fractional part of vertical-sweep period allowed for vertical-return time.

A 500-MEGACYCLE RADIO-RELAY DISTRIBUTION SYSTEM FOR TELEVISION

By

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Summary—*This paper reviews the development and operation of a radio-relay system for television-program distribution. Television programs from W2XBS located on the Empire State Building in New York City were delivered to Riverhead, Long Island, through radio repeating stations near Hauppauge and Rocky Point. The amplification in the repeaters was accomplished without demodulation and re-modulation in the repeater equipment. Radio carrier frequencies between 400 and 500 Mc were employed in the radio links. The carrier was frequency modulated directly by the video signals. The paper describes some of the problems involved in designing radio networks to interconnect television broadcasting stations and describes some of the methods applied in their solution. As a result of these developments it is now feasible to provide radio networks for television-program distribution over wide areas.*

NATIONWIDE television service requires the distribution to remote areas of program material originating in any one locality as is now done by wire networks in sound broadcasting. Facilities for this service usually employ directive radio or wire networks which are capable of transmitting a modulation band ranging from 30 cycles to several million cycles per second.

Several years ago, an experimental system was set up to determine the feasibility of television relaying by radio. This work resulted in the construction and operation of a radio relay between New York City and Camden, New Jersey, in 1933.[1] At that time a 120-line picture was transmitted which required a modulation band of one megacycle.

During the last eight years the video-modulation band has increased from one megacycle to the present requirement of four megacycles. In developing a distribution system, possible future requirements should be taken into account and these may necessitate the use of even wider modulation bands.

In 1934 tubes of appreciable output at frequencies of 100 to 200 megacycles were available. To make use of these frequencies, R.C.A. Communications, Inc. installed an experimental communication circuit between New York City and Philadelphia.[2] This circuit provided valu-

[1] "The Radio-Relay Link for Television Signals," C. J. Young, *Proc. I.R.E.*, Nov. 1934.

[2] "The New York-Philadelphia Ultra High Frequency Facsimile Relay System," H. H. Beverage, *RCA Review*, July 1936.

"Practical Application of an Ultra High Frequency Radio Relay Circuit," J. Ernest Smith, F. H. Kroger and R. W. George, *Proc. I.R.E.*, Nov. 1938.

able information on operating costs, maintenance, and signal propagation at 100 megacycles. By 1938 the progress of the art permitted the design of radio-relaying equipment for frequencies as high as 500 megacycles. Consequently, in 1939 an experimental 500-megacycle television radio relay was built and operated.

To determine the feasibility of relaying by radio it is necessary to consider the proper antenna heights, antenna aperture dimensions, spacing between relay stations, and power radiated to give the most economical result. These considerations must be based upon available data on ultra-high-frequency propagation.[3] There are other factors

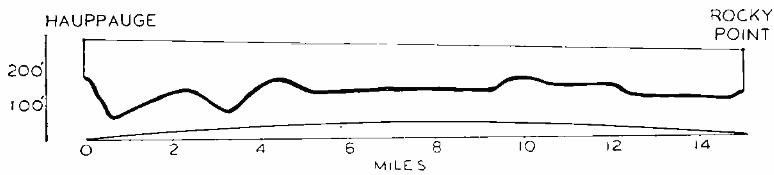

Fig. 1—Profile of terrain between Hauppauge and Rocky Point.

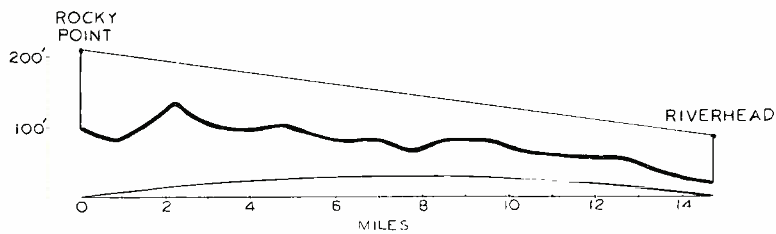

Fig. 2—Profile of terrain between Rocky Point and Riverhead.

such as signal fading, unusual interfering noise sources, and antenna efficiency which must be taken into account. Moreover, the particular terrain in question will be a factor in determining the final layout since advantageous use can often be made of elevated points thus permitting lower tower structures.

It is of interest to note that when the antennas of a fixed aperture at each end of a link are at heights sufficient to allow the direct and reflected rays to arrive at the receiving antenna with a phase angle of substantially 120 degrees, the required amplifier power gain is then inversely proportional to the square of the frequency, assuming a constant transmitter power input. This fact indicates that it would be advantageous to operate at as high a frequency as possible; especially, since the higher the frequency, the lower will be the antenna heights

[3] "Ultra High Frequency Propagation Formulas," H. O. Peterson, *RCA Review*, Oct. 1939.

necessary to bring the direct and reflected rays together with the 120-degree phase angle at the receiving antenna. It was felt that 500 megacycles for a carrier frequency would be a suitable starting point even though higher frequencies may offer additional advantages. It is expected, however, that there will be an upper limit of useful frequencies due to absorption by rain, fog, snow, and gases of the atmos-

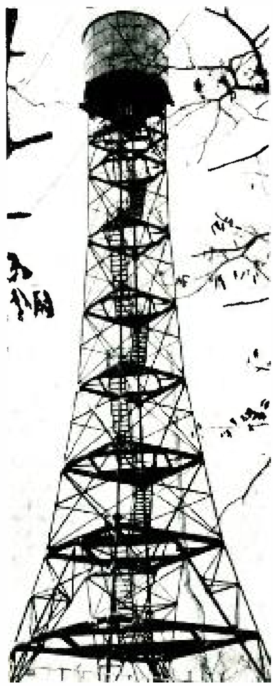

Fig. 3—Hauppauge tower and penthouse.

phere. Also, the noise level in receiving equipment to make use of such high frequencies may be greater than is now anticipated.

Frequency modulation was employed in this work due to its advantages for circuits where multipath phenomena are absent.[4] Furthermore, tubes were available which greatly minimize the equipment for producing a frequency-modulated carrier. In addition, frequency modulation permits the use of limiting and class C amplification thus, simplifying the problem of maintaining overall circuit linearity.

[4] "Freq. Modulation Propagation Characteristics," M. G. Crosby, *Proc. I.R.E.*, Vol. 24, No. 6, June, 1936.

Recent measurements[5] have shown that automobile-ignition interference is present with about the same field strength on all frequencies between 40 and 450 megacycles. This means that the interfering energy received with antennas of the same effective height and directivity will be constant with frequency. However, the antenna directivity (power gain) for a given aperture area increases in proportion to the square of the frequency; hence, a large reduction of received interference is had at the higher frequencies if the interference is not generated directly in front of the receiving antenna.

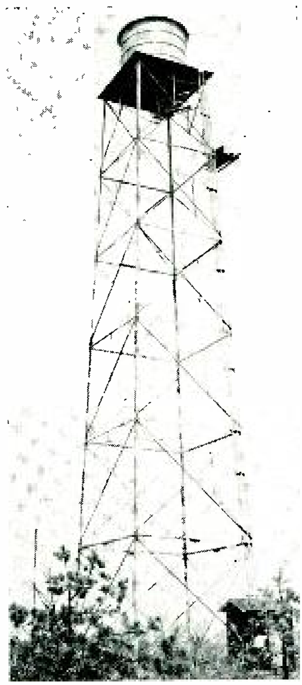

Fig. 4—Rocky Point tower and penthouse.

In Figure 1 is shown the profile of terrain between the Hauppauge terminal and the Rocky Point repeater. This profile has been plotted to show the conditions existing with an earth's radius 4/3 of its actual value in order to take into account the normal refraction occurring in the earth's atmosphere. Using this profile to estimate the magnitude of the received signal by combining the direct and reflected rays in the usual manner we obtain a value of 1.6 millivolts delivered across a 75-ohm receiver input, assuming a transmitter power of one watt and

[5] "Field Strength of Motor Car Ignition Between 40 and 450 Megacycles," R. W. George, presented at the U.R.S.I., Washington, D. C., April, 1940.

an antenna gain of 20 decibels at each station. This calculation was based on the assumption that the indirect ray was reflected from the ground whereas, actually, underbrush and trees would cause the effective reflection point to be somewhat higher. A difference of ten feet would reduce the calculated voltage by 20 per cent giving 1.3 millivolts. The actual measured value was found to be 1.2 millivolts.

In Figure 2 a similar profile is shown between Rocky Point and Riverhead. Calculation of the expected signal compared with that actually measured shows agreement within about four decibels which is as close as the accuracy of the profile will allow in this case.

Fig. 5—Riverhead receiving antenna.

Although extensive continuous observations have not been made over such fifteen-mile paths, it is felt that signal fading would be quite small, as no appreciable fading has been observed during the tests of this television relay. A circuit of 30 miles in length operated on 500 megacycles with good optical clearance has shown fading of more than ten decibels to occur rarely and then only for short periods of less than an hour.

Assuming that fading is produced by varying amounts of refraction which results in varying the path-length difference between the direct and reflected rays, then it would be expected that minimum fading would occur when, under average conditions, the path-length difference is ½ wavelength. This condition brings the direct and reflected rays in phase at the receiving antenna which results in the strongest possible field. Changes of refraction conditions in the atmosphere will only alter the phase angle between the two components and will alter the resultant field strength only slightly. The opposite condition occurs when the two components are nearly in phase opposition at the receiving antenna, as in this case a small change in phase angle gives a large change in the resultant field. It is not usually economical to place the antennas at a sufficient height to bring the direct and reflected rays in phase so that a compromise with antenna heights should give a path-length difference of 1/6 wavelength or 120 degrees. This condition results in a received field equal to that which would be obtained in free space where only the direct ray would be present.

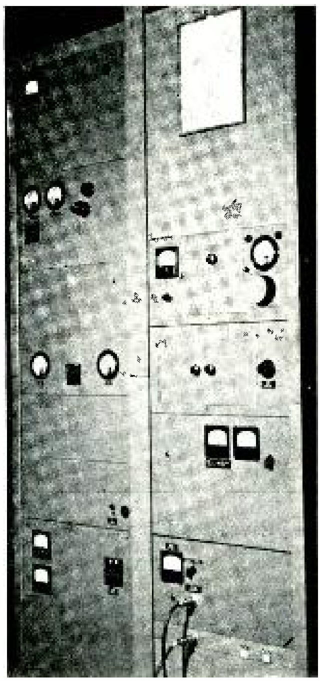

Fig. 6—500-Megacycle terminal transmitter.

In order to choose the proper amplifying system it becomes necessary to know the amount of gain to be incorporated in each repeater amplifier. The term repeater as used here is considered to be the apparatus between input and output antennas. The maximum gain that can be used is determined by the ratio between the maximum repeater-output power and the noise power appearing at the repeater input. If this ratio is 120 decibels and a signal-to-noise ratio of 50 decibels is desired, then the repeater gain would be 70 decibels. It is, of course, necessary to use the proper size of antennas, antenna heights, and station spacings to bring the signal-to-noise ratio at the repeater input to the desired value.

At the present state of development, the 500-megacycle receivers having an r-f band of eight megacycles give an equivalent noise-power input of 1.4 times 10^{-12} watts. The signal power required for a 50-decibel signal-to-noise ratio is then 1.4 times 10^{-7} watts. If the maximum power output of the repeater amplifier is 1.4 watts we will require a repeater gain of 10^{7} or 70 decibels.

We have seen that our repeater amplifier should have an overall gain in the neighborhood of 70 decibels and this amplifier must have a flat bandpass of at least eight megacycles. Experience has shown that

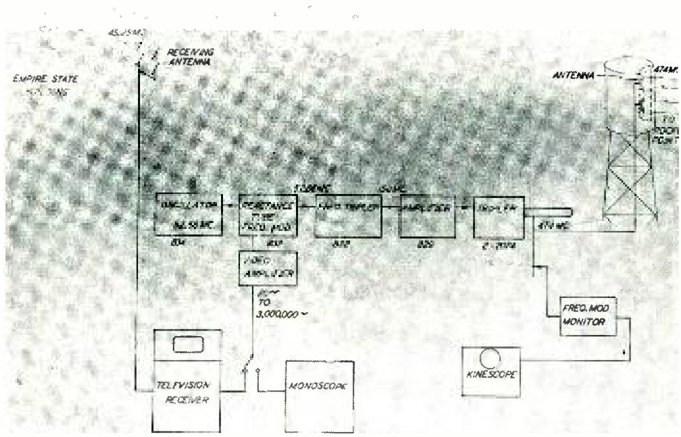

Fig. 7—474-Mc terminal transmitter at Hauppauge, N. Y.

stable operation can be maintained by converting the 500-megacycle signal to a lower intermediate frequency in the neighborhood of 100 megacycles where the major portion of the gain is readily realized. An output of about 0.7 watt can be obtained on either the same 500-Mc carrier frequency or an adjacent frequency by a high, level converter. A single stage of amplification is sufficient to raise this power level to about two watts.

The system herein described used a repeater having an input frequency of 474 megacycles and an output frequency of 460 megacycles.

The relay system as demonstrated consisted of a terminal station at Hauppauge, a repeater station at Rocky Point, and a terminal at Riverhead, all located on Long Island. The tower and antenna structures at Hauppauge, Rocky Point and Riverhead are shown in Figures

3, 4, and 5. A spacing of 15 miles between stations made this circuit 30 miles long. Television signals as broadcast from the Empire State Building on 45.25 megacycles were received at Hauppauge on a receiver

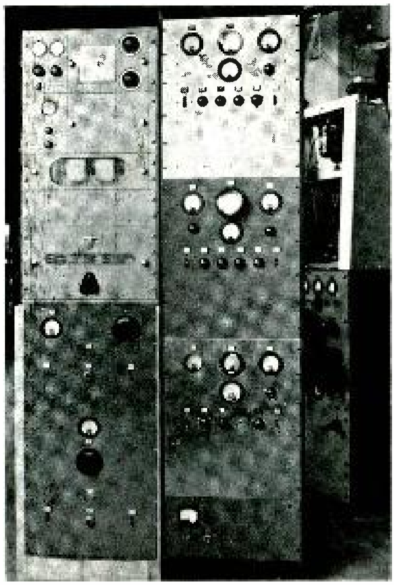

Fig. 8—500-Megacycle repeater amplifier.

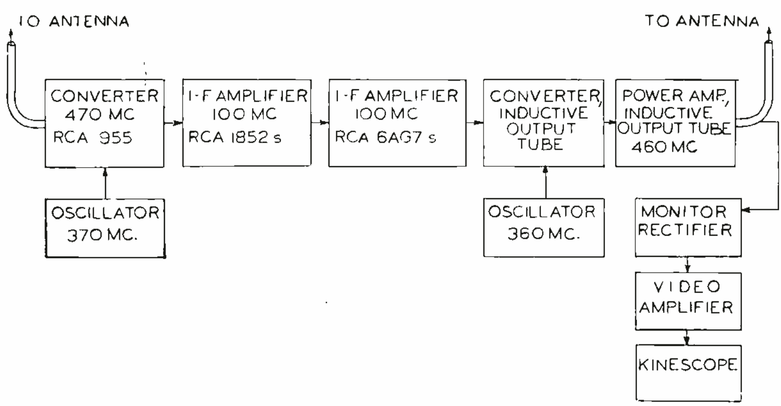

Fig. 9—Repeater-amplifier schematic diagram.

whose video output was fed to the 474-megacycle frequency-modulated terminal transmitter. These signals were relayed to Riverhead with good quality, a total distance of 70 miles from New York. Another source of signals at Hauppauge was supplied by a monoscope described elsewhere in this paper.

The terminal transmitter is shown in Figure 6 and consisted of a 52.7-megacycle oscillator coupled with a reactance tube which was fed from a video amplifier carrying the picture signal. Following the oscillator, a wide-band tripler stage brought the carrier frequency to 158 megacycles after which a power-amplifier stage served to drive another wide-band tripler stage giving an output frequency of 474 megacycles

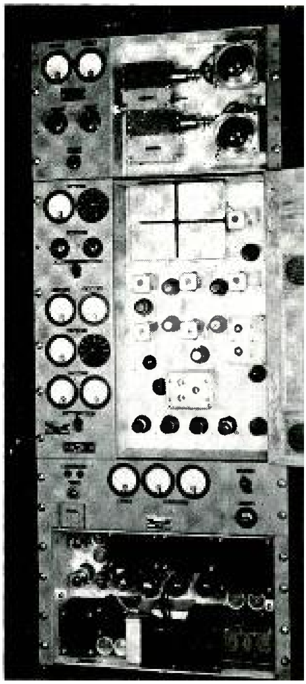

Fig. 10—Terminal receiver.

at a power level of one watt. Figure 7 shows a block diagram of the transmitter elements. A monitor rectifier at the transmitter output gave video signals to allow the output picture to be observed at all times.

The terminal transmitter was coupled to the cylindrical parabolic-reflecting antenna by means of a single 75-ohm coaxial feed line having a loss of one decibel per 100 feet. The parabolic reflector was excited by four folded doublets located along the focal axis. The antenna aperture of 110 square feet gave a measured power gain of 20 decibels over that of a half-wave dipole in free space. A 100-foot tower supported the antenna house which was fabricated from ¼-inch waterproof plywood treated with boiled linseed oil. The repeater amplifiers and antennas at Rocky Point were housed in another cylindrical plywood structure at the top of a 115-foot tower. The antennas were similar to the one used at the Hauppauge terminal.

The repeater amplifier is shown in Figure 8 and the block diagram of elements is shown in Figure 9. The input signal of 474 megacycles was fed to a triode converter, 8 stages of intermediate-frequency amplification, an inductive output tube operated as a high-level converter, followed by an inductive output tube operating as a power amplifier. The input converter made use of an RCA 955 triode by feeding the signal to the grid and supplying a local oscillator excitation of 374 megacycles to the cathode. The 100-megacycle intermediate frequency

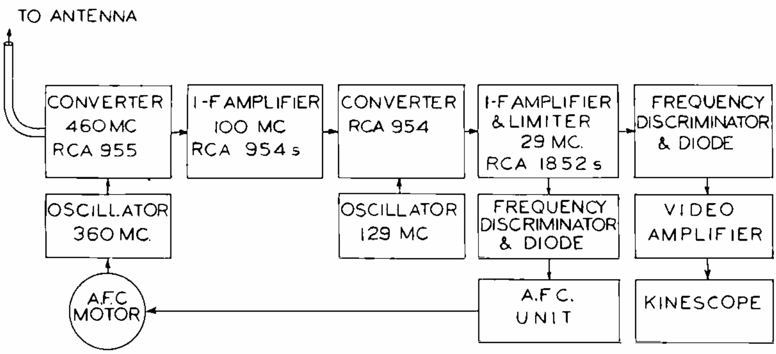

Fig. 11—Terminal-receiver schematic diagram.

was taken from the plate circuit and amplified by six wide-band transformer-coupled stages using RCA 1852 pentodes. Following this were two wide-band stages using RCA 6AG7 tubes which brought the level up to 1 watt. This level was necessary to drive the inductive output tube as a converter. A local oscillator of 360 megacycles was also supplied to the input of this tube in order to obtain the final output frequency of 460 megacycles at a level of about 0.7 watts. The final inductive output-tube power amplifier provided sufficient gain to feed 2 watts to the antenna. The overall gain of the repeater amplifier was measured to be 80 decibels under actual operating conditions.

Monitor equipment was provided in order that the picture quality could be continuously observed. At the repeater station the monitor Kinescope was located on the ground and was fed by a coaxial cable from the monitor rectifier at the antenna.

At the Riverhead terminal a cylindrical parabolic antenna situated 70 feet above ground fed the signal to the terminal receiver over a 75-ohm coaxial cable. The receiver used an RCA 955 converter identical to that in the repeater amplifier. Following this, a two-stage 100-megacycle i-f amplifier using RCA 954 tubes gave sufficient gain and selectivity to allow another conversion to 29 megacycles at which frequency the major portion of the gain was obtained with six wide-band transformer-coupled stages of RCA 1852 tubes. A frequency discriminator of the Conrad type in combination with an RCA 6H6 diode rectifier delivered push-pull video signals to an amplifier which in turn was connected to the monitor Kinescope. Figure 10 shows a photograph of the terminal receiver and Figure 11 shows the block diagram.

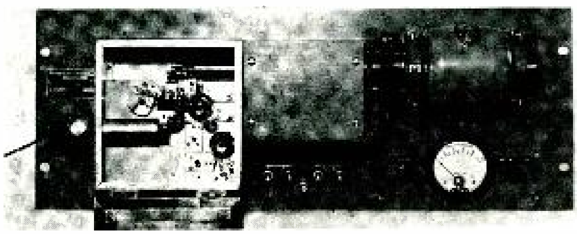

Fig. 12—100-Megacycle sweep oscillator.

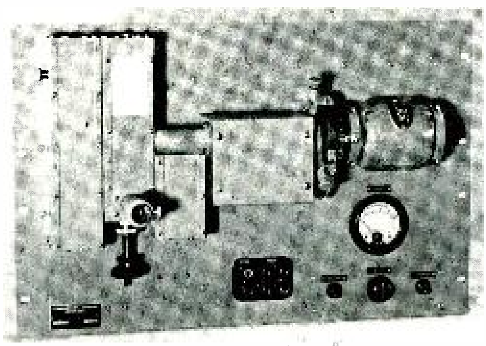

Fig. 13—500-Megacycle sweep oscillator.

In any new radio development, accurate test equipment is of the utmost importance. While such test equipment is rapidly becoming standardized for 50-Mc television broadcasting, the state of the art was such at the time the 500-Mc relay project was undertaken, that it was necessary to develop the required test tools during the original research. This need will be readily appreciated when it is understood that individual links of the relay must be capable of very high fidelity video transmission in order that cumulative distortion in an overall chain shall not exceed the television-broadcast requirements.

Figures 12 and 13 show two sweep-oscillator signal generators that were invaluable during the alignment and adjustment of the r-f and i-f bandpass stages. The former has a mean-carrier frequency which may be manually adjusted from 100 to 150 Mc and the latter covers the range from 450 to 550 Mc. Each unit is provided with a motor-driven variable capacitor which permits any amount of frequency deviation up to 16 Mc at a 50-cycle rate. A sweep rate other than 60 cycles was chosen to avoid confusion with hum patterns. Both units have automatic volume-control circuits to eliminate any amplitude modulation in the output.

Figure 14 is a photograph of a typical television oscilloscope used throughout the relay. Its prime function is to monitor the composite

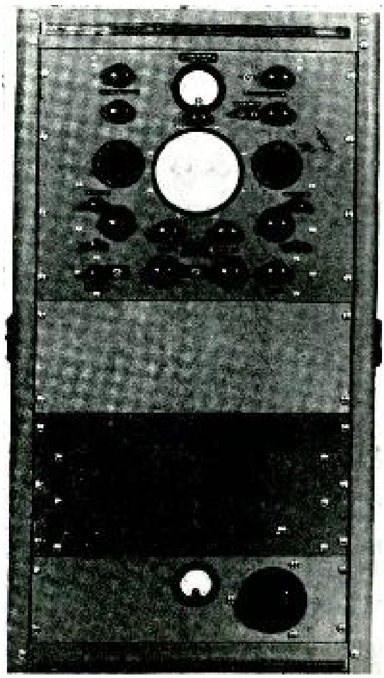

Fig. 14—High-fidelity oscilloscope.

video signal and, particularly, to show the formation of the synchronizing and blanking impulses. For this purpose, 60-cycle and 13,230-cycle horizontal sweep oscillators were provided. In addition, horizontal and vertical amplifiers having a flat response from 60 cycles to 7 megacycles with identical phase characteristics over this range were incorporated to allow precise study of the individual relay repeaters. A calibrated attenuator together with a fixed voltage source in each amplifier facilitates its use as a peak voltmeter. To simplify schedule analyses of received square waves into their Fourier series components, a 20-megacycle keying circuit was inserted to "break up" the received wave envelope into accurately spaced dots. The overall sensitivity of the unit produces a one-inch deflection on the RCA 1802 tube with a 0.1-volt peak-to-peak signal applied at the amplifier input. Operation of the oscilloscope is conventional and requires no description. The calibrated peak-voltmeter feature was especially helpful since it permitted direct correlation of the r-f frequency deviation with the video-signal amplitude.

For low-frequency amplitude and phase correction, 60-cycle square waves were found to be the most satisfactory test signal from the standpoint of expediency as well as accuracy. This method has the

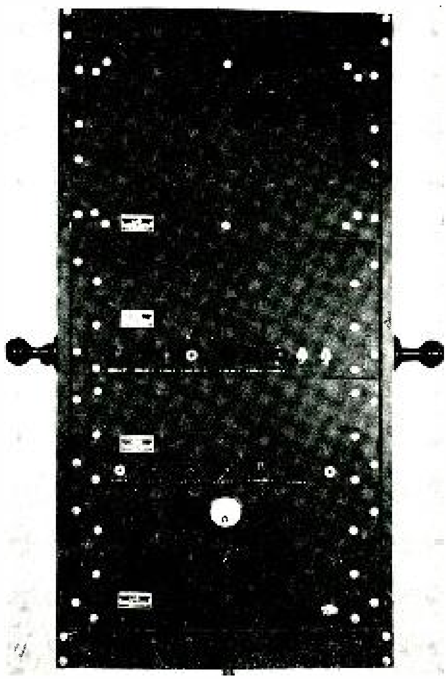

Fig. 15—Monoscope-signal generator.

advantage that any normal oscilloscope has sufficient fidelity to reproduce the received test wave. A particular variation, when used with the television oscilloscope, was to transmit the video-blanking impulses only. This test signal consists of short 60-cycle square waves (7 per cent duration) and 13,230-cycle square waves (15 per cent duration) transmitted simultaneously. When the received signal is viewed on a 60-cycle sweep in the scope, the two waves will have equal amplitudes only when the transmission medium is perfectly corrected. With slight errors in compensation, the line-frequency-pulse amplitude will be greater or less than the field-pulse amplitude due to the difference in gain or phase characteristics of the medium for the 13,230 and 60-cycle signal frequencies. To correct the high-frequency portion of the band, 100-Kc square waves were used with some success. In general, however, it was more expedient to utilize ordinary signal generators with ranges up to 10 Mc due to the wide pass bands involved.

The final and most conclusive test employed a high-fidelity monoscope—kinescope chain. Obviously, this signal checks the overall-circuit performance with respect to stability of synchronization, amplitude linearity, signal transients, picture definition, and noise effects simultaneously. The monoscope unit, containing the RCA 1899 tube,

Fig. 16—Synchronizing-signal generator.

the deflection circuits, mixing amplifier, and power supplies, is shown in Figure 15. The synchronizing signal generator is illustrated in Figure 16. The monitor kinescope, housing the 12-inch RCA 1803 tube, video amplifier, separators, and deflection circuits, is shown in Figure 17. All units were made portable for operation in the field. The monoscope chain transmits a standard composite video and "sync" signal of 500-line definition. Provisions were made to reverse the polarity of the video signal to obtain either a white or black background thus simulating the extreme conditions obtained with the average "movie". The 1899-tube pattern contains a wedge of half-tone steps between black and white for linearity checks. The usual horizontal and vertical line wedges show the picture definition, the presence of transients, and at what frequencies these transients occur.

Before discussing the overall performance of the relay and the various tests conducted, it will be helpful to consider briefly the nature of a television signal and, particularly, to compare frequency modulation with amplitude modulation for this type of modulating signal. Normally, amplitude modulation is studied in one of three forms; namely, the modulation envelope, the sideband configuration or the

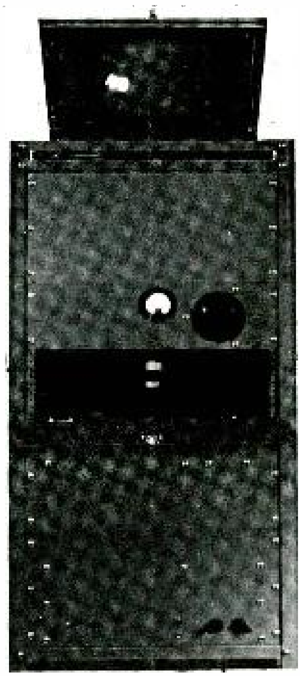

Fig. 17—Monitor Kinescope.

vector method of representation. For our purpose, the first two forms will be sufficient.

With amplitude modulation, it is well known that the sidebands are symmetrical about the carrier both as to amplitude and phase no matter how unsymmetrical the positive and negative polarities of the modulating wave may be. It is a curious fact that the first tendency of a student, during his early amplitude-modulation studies, is to associate one sideband with one polarity of the modulating wave. While this tendency is definitely incorrect when applied to amplitude modulation it is not so far wrong when the sideband distribution of a frequency modulated wave is considered. Actually, the amplitude of the sideband

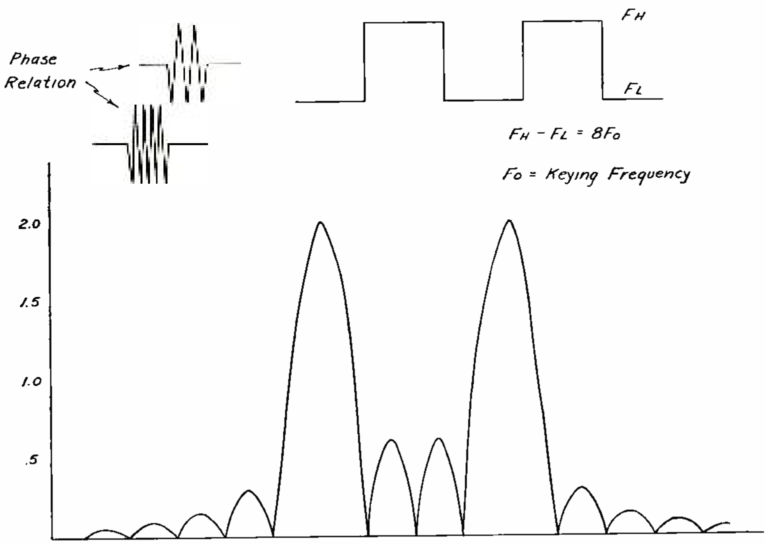

Fig. 18—F-m sidebands (symmetrical modulation).

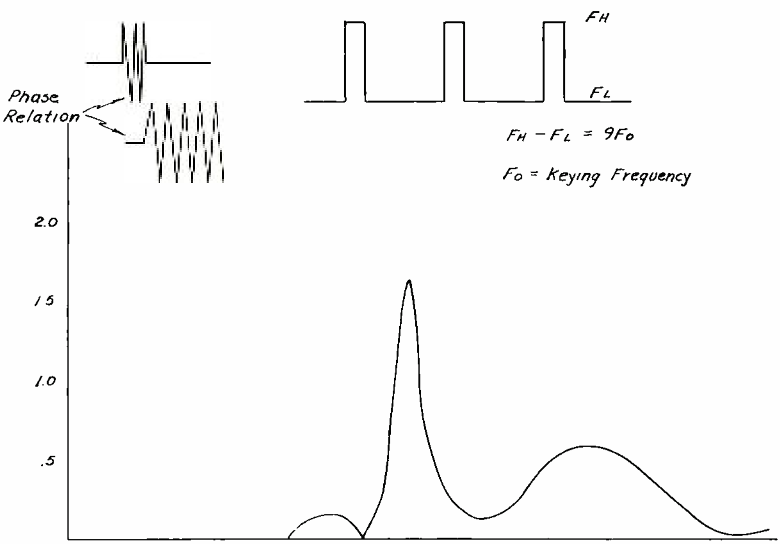

Fig. 19—F-m sidebands (unsymmetrical modulation).

components are symmetrical about the carrier in frequency modulation *only when the polarities of the modulating wave have symmetrical waveshapes.* The phases of the sidebands, on the other hand, are never symmetrical about the carrier.*

To illustrate this more clearly, consider Figure 18 which disregards the phase reversals of the even-order sideband components. It will be seen that the amplitudes of the components are symmetrical for the

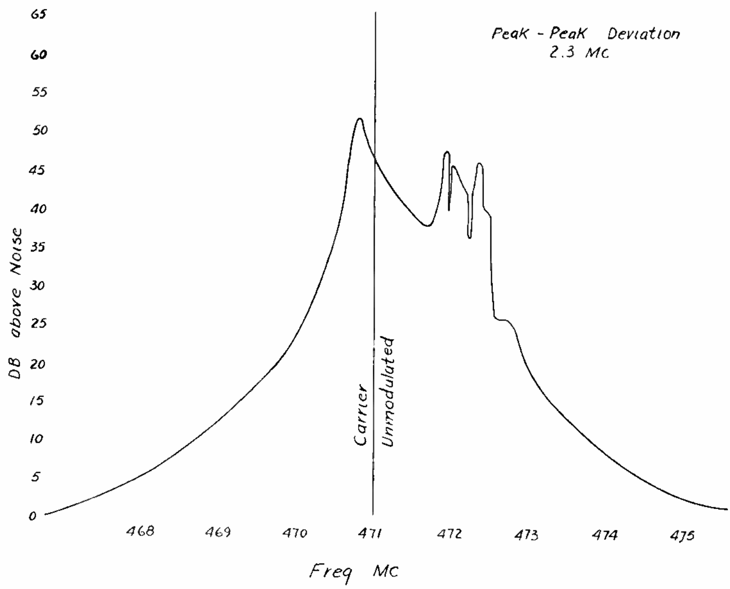

Fig. 20—F-m sidebands (monoscope composite signal).

50-50 square-wave envelope. The sidebands are calculated on the basis that the peak-to-peak frequency deviation is eight times the keying rate. A higher value of this ratio would merely move the peak sideband maxima farther apart without destroying symmetry. It will be understood that, theoretically, the sidebands extend from zero to infinity since it has been assumed that the modulating wave produces an instantaneous frequency change in the modulated wave. Now observe Figure 19 where the modulating wave is a 20-80 dot. The sideband configuration shows a very pronounced component amplitude corresponding to the frequency of longest duration of the instantaneous carrier. Further, there is no sideband symmetry about any frequency. In fact, the

* This statement is correct even if we include the very special case of sine-wave modulation with a small modulating index, since the even-order components having a 180-degree phase difference still exist although they are usually disregarded due to their small amplitudes.

sidebands are about as unsymmetrical as the polarities of the modulating envelop.

It will be understood that the present discussion is not an attempt to cover the theory of frequency modulation completely. The intention has been to show some fundamental concepts that were instrumental in determining the type of tests to be conducted. Heuristically, we may

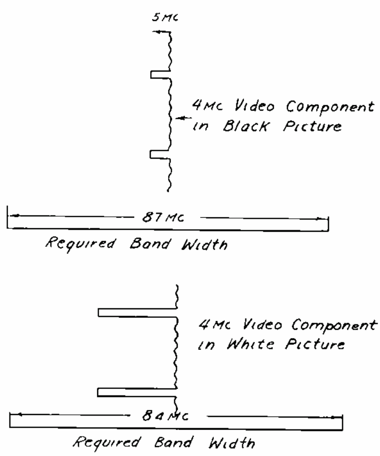

Fig. 21—A-C frequency modulation (carrier at mid-band)

reason from the two examples above, that the blanking and synchronizing pulses will give sideband dissymmetry. However, the higher frequency components of the video signal, considered as small amplitude sine waves, would tend to submerge this result. Figure 20 showing the measured sidebands of the 1899-monoscope composite signal indicates that no great amount of dissymmetry is obtained for a complete picture signal.

A distinction should be made between a-c transmission and d-c transmission of the video signal as applied to a frequency-modulated system. Since the standard video composite signal inherently contains all necessary information as to the picture background level (by maintaining a fixed-peak amplitude and fixed super-sync amplitude), the d-c component may be restored in the video circuits at the receiving terminal. Insofar as the relay is concerned, the signal to be transmitted may be considered as an a-c wave only. If this is done the r-f carrier component will not be shifted when modulation is applied. On the other hand, if the d-c component is transmitted over the relay, the r-f carrier component will vary with the picture background and the frequencies corresponding to the supersync pulses will remain unchanged. Comparison of Figures 21 and 22 indicates the range of r-f or i-f frequency deviation with respect to the pass band for the two cases.

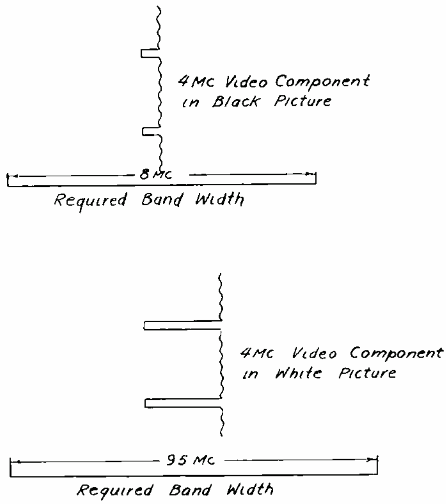

Fig. 22—D-C frequency modulation (carrier variable)

The results obtained on the overall experimental relay will now be described. With the carrier set at midband and employing a-c transmission with a peak deviation from the carrier of 2 Mc, picture definition of 375 lines was obtained with negligible transients. Synchronizing and linearity were entirely satisfactory. The r-m-s signal-to-noise ratio was 31 db.*

To determine if more effective use of the pass band could be obtained with d-c transmission, the carrier was manually adjusted in steps from the center towards the edge of the pass band. This was permissible since the d-c component of the monoscope signal is constant. At each point the video-signal polarity was reversed to compare the transmissions for two extremes of background level. In effect, this test also simulated partial sideband transmission as the carrier approached the edge of the pass band. The integrated opinion of several observers was that the fidelity of transmission remains unchanged until definite sideband clipping of the synchronizing pulses resulted in poor syn-

* R-m-s S/N ratio is here defined as the ratio of the r-m-s value of a sine-wave having the peak-to-peak amplitude of the received video signal as compared to the r-m-s noise.

chronization. With the carrier shifted toward the edge opposite to the excursions of the supersync pulses, however, it was found that a somewhat larger peak-to-peak frequency deviation could be employed. Because of this and the possible theoretical advantage of maintaining the synchronizing pulses in a fixed portion of the band, automatic d-c insertion in the frequency modulators may be employed. As a matter of interest, the simultaneous transmission of video and sound on the same carrier was satisfactorily demonstrated in an additional test.

In conclusion, it can be said that radio relaying of television signals in the ultra-high-frequency spectrum above 400 Mc has been successfully accomplished and that a system consisting of radio relays would be technically adequate and feasible for television program distribution.

A PRECISION TELEVISION SYNCHRONIZING-SIGNAL GENERATOR

BY

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Summary.—*The R.M.A. standard television synchronizing signal consists of three types of rectangular pulses in a single wave. The timing of all pulses is such that the leading edge of each pulse is either 1/26460-second or 2/26460-second from the leading edge of its adjacent pulses. The three types of pulses differ only in the duration (or width) of the individual pulses, which is not extremely critical.*

In the synchronizing-signal generator described a frequency-regulated master oscillator and multivibrator generate a master wave of uniform pulses which occur at 1/26460-second intervals. This master wave as produced contains only pulses which have the width specified for the narrowest type of pulses of the R.M.A. synchronizing wave, but of course, it has many extra pulses which are not required in the R.M.A. signal. Also many of the pulses should be made wider as the R.M.A. signal contains two other types of pulses of greater widths. These two other types of pulses are produced by "lap-joining" other suitable auxiliary pulses to the trailing ends of the narrow pulses of the master wave, without altering the critical leading edges of the pulses. The particular pulses of the master wave which are undesired in the final wave are "blanked" out by other auxiliary-keying pulses.

Since some of the auxiliary-keying waves are produced by 60-cycle pulse waves, it is necessary to lock the 26,460-c.p.s. and the 60-cycle pulse waves together. This is done by a chain of three pulse-counters (acting as frequency dividers) which derive the 60-cycle wave from the master 26,460-cycle wave. Stability is obtained since the tubes involved are employed substantially as low-resistance keys.

The advantage of frequency dividing over frequency-multiplication is discussed.

The entire chain of frequency-dividers is kept in synchronism with the 60-cycle power system by comparing the 60-cycle pulse wave to the power wave to obtain a control voltage for regulating the 26,460-cycle oscillator. The circuit used is such that the derived-control voltage is made free of 60-cycle components without sacrifice of quick response.

For economic reasons involving reliability of the transmitting system, maximum attention was given to stability and inherent accuracy of performance in all critical respects.

INTRODUCTION

MANY of the standards which are necessary to specify a television system pertain directly or indirectly to the characteristics of the synchronizing-signal generator. The shapes of the synchronizing pulses, the type of interlacing, number of lines, return periods, and picture-repetition rate, all affect the design of this generator. The Radio Manufacturers Association has adopted certain standards covering these points, but has not set tolerances on many of the standards. In order to obtain acceptable performance extremely high and stable accuracy is necessary in certain respects, as indicated in the description below. An entirely-electronic generator has been developed which employs circuits for obtaining high reliability and inherent precision of the output waves in all critical respects.

Principles of Methods Used

In odd-line interlaced scanning as adopted by R.M.A. the horizontal-deflecting frequency, 13,230 c.p.s., is precisely a whole number plus one-half times the field frequency, 60 c.p.s. This results in interlacing since each field scan then contains a whole number plus one-half scanning lines. (This number is 220.5 lines for the R.M.A. standard 441-line television.) To meet these conditions it is essential that the generator of the 60 c.p.s. synchronizing pulses be rigidly interlocked with the 13,230 c.p.s. pulses. Since the two frequencies differ by such a large ratio and also do not have a whole number relation, a single-stage stable direct interlock is not feasible. The exact number of lines was chosen such as to permit interlocking by several stages where each step differs in frequency by a small whole-number ratio. A regulated master oscillator produces 26,460 cycles-per-second signal for driving a frequency-divider circuit producing half that frequency, i.e., 13,230 c.p.s. Similar frequency-divider circuits operating in cascade also divide the frequency of the 26,460-c.p.s. oscillator in whole-number odd steps of 7, 9, and 7 producing frequencies of 3,780, 420, and 60 cycles per second, respectively.

For several reasons it is desirable (though not essential) that the nominal 60-cycle-per-second output of the synchronizing generator be accurately synchronized with the main power system of the community being served by the transmitter.* Due to the cost of filtering and shielding, television receivers will generally have some residual 60-cycle and 120-cycle ripple in their deflection systems and beam-modulating amplifiers. If these spurious influences are synchronous with the picture deflection they are much less annoying as the small power-frequency waves of displacement of the picture subject and the modulation shadows will not move vertically over the screen. Also when motion picture films are used as program material, the film projector should be synchronous and phased with the iconoscope-deflecting system within approximately 4 degrees. This condition is conveniently

* R. D. Kell, A. V. Bedford, and M. A. Trainer, "Scanning Sequence and Repetition Rate of Television Images," *Proc. I.R.E.*, Vol. 24, pp. 559-576; April 1936.

obtained by driving the projector with a synchronous motor on the 60-cycle power system. Then in order to keep the "60-cycle" signal produced synchronous with respect to the local 60-cycle power supply, the two are compared in a special improved circuit which produces a controlling voltage for regulating the frequency of the 26,460-c.p.s. master oscillator.

An inspection of the R.M.A. standard synchonizing-signal in Figure 1 shows that it consists of various time-mixtures of 26,460 and 13,230 cycle-per-second pulses. The 26,460-cycle pulses are of two kinds and occur in small groups at regular sixtieth-second intervals. Each group consists of six narrow "equalizing" pulses, six much wider pulses and six more narrow "equalizing" pulses, all occurring in the order named. (The six "wider" pulses mentioned, acting as a unit, comprise a single serrated "vertical" or field synchronizing pulse.)* Each such group occupies only about 4 per cent of the one-sixtieth second.

The remaining 96 per cent of the time is occupied by the normal 13,230-cycle-per-second horizontal or "line" synchronizing pulses. Their width is $0.08H$ (where H is 1/13230 second) which is twice as wide as the equalizing pulses.

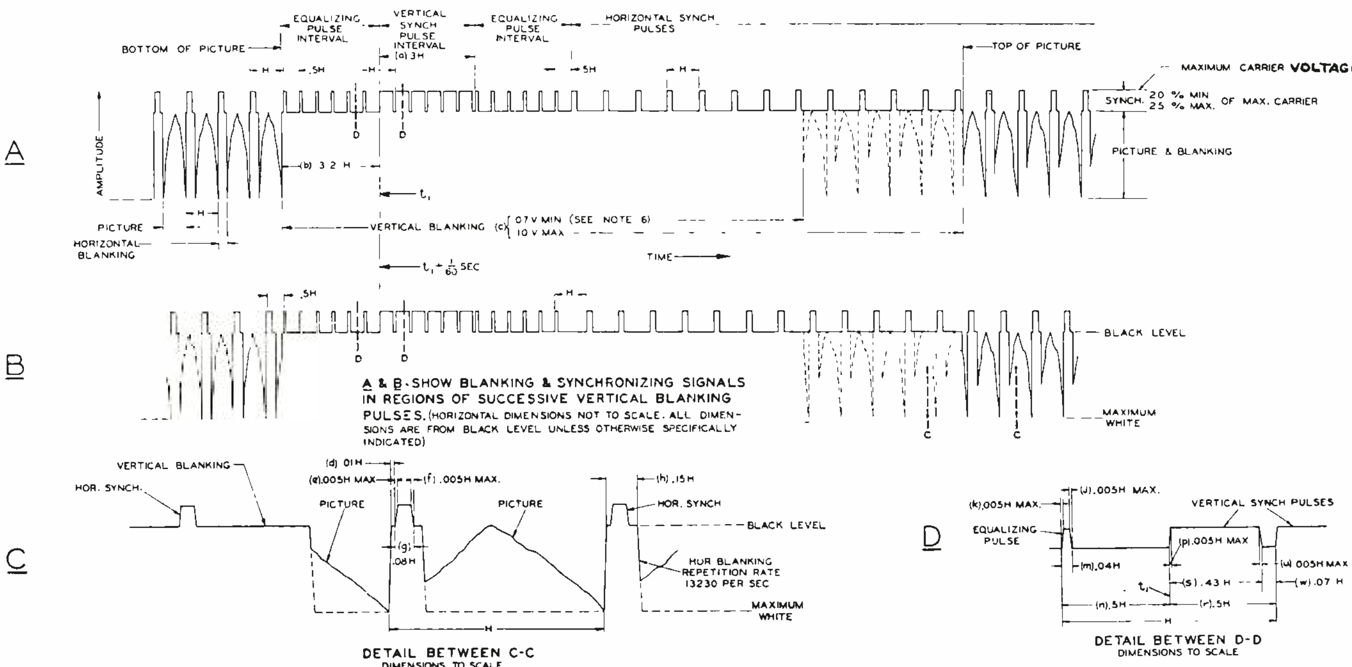
When used in the television receiver the leading edges of these pulses by their abrupt rise cause "firing" of the horizontal-deflecting oscillator. The duration of the pulses or shape of the trailing edge of the pulses do not appreciably affect the horizontal-deflecting circuit. According to the standard the 26,460-per-second equalizing pulses and the serrated-vertical pulses have alternate rising leading edges which are timed with the 13,230-per-second pulses such as to provide continuous uniform rising edges at intervals of 1/13,230 second. These rising edges provide horizontal synchronization in the receivers which is uninterrupted by the vertical synchronizing pulses.

In one of the methods tested while developing the synchronizing generator, the three different kinds of pulses which comprise the entire synchonizing wave were generated continuously in separate circuits. The output of each circuit was then keyed by an amplifier which was intermittently driven to cut-off by certain "keying" waves. The final wave was obtained by adding the several keyed outputs. This simple method had a serious fault in that permanently accurate relative timing of the three types of pulses was not obtained except by frequent adjustment. The three circuits which generated the three kinds of

* Receivers can be made to operate on a synchronizing signal which is somewhat less complicated than the R.M.A. standard signal. However, to do so requires either some sacrifice in performance or the use of additional complication in the receiver. Since the number of receivers will be much greater than the number of transmitters, the generation of the R.M.A. signal is economically preferable.

RMA STANDARD T-111 TELEVISION SIGNAL

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



1- DIAGRAM C SHOWS ENLARGED DETAIL VIEW OF SIGNAL IN VIEW B BETWEEN LINES C-C

2- DIAGRAM D SHOWS ENLARGED DETAIL VIEW OF SYNCH. SIGNAL IN VIEW A BETWEEN LINES D-D

3- H=TIME FROM START OF ONE LINE TO START OF NEXT LINE = 1/13230 SEC.

4- V=TIME FROM START OF ONE FIELD TO START OF NEXT FIELD = 1/60 SEC. = $220\frac{1}{2}$ H

5- LEADING AND TRAILING EDGES OF BOTH HOR. AND VERT. BLANKING PULSES HAVE SLOPES (NOT INDICATED IN A & B) WHICH SHOULD BE KEPT AS STEEP AS POSSIBLE

6- RECEIVER VERTICAL RETRACE SHALL BE COMPLETE AT END OF .07 V

Fig. 1—*A* and *B* are portions of the same wave. The Synchronizing-signal Generator produces the portion (25 per cent) of the R.M.A. Standard Television Signal above the "black level". It also produces the blanking portion of the signal shown below "black level".

pulses were synchronized by the common 26,460-cycle pulses, but due to variable degrees of "firing resistance" and finite slope of the synchronizing pulses, timing errors, between the several kinds of pulses in the output wave, of several per cent of H frequently occurred. Further difficulty was experienced due to the three kinds of pulses having different wave shapes of their leading edges and different rates of rise due to different constants in the three generating and keying circuits. These errors sometimes resulted in loss of horizontal synchronism in the receiver at those points of non-uniformity which would require an abrupt increase in oscillation frequency and at other times caused a slight displacement of a few scanning lines at the top of the picture screen. For operation beyond reproach it seems that the error between any two adjacent horizontal-synchronizing leading edges should be considerably less than that corresponding to one picture element, which is of the order of 0.002 H.

In the present synchronizing generator, uniformity of both timing and wave shape of all leading pulses is inherent due to a single 26,460-c.p.s. multivibrator (which is driven by rectangular waves obtained from a tuned oscillator and limiter) producing the leading edges of all pulses in the finished R.M.A. synchronizing wave. These pulses as produced have a width of 0.04 H and without alteration become the "equalizing pulses" in the final wave. A section of the 26,460-c.p.s. pulses are made wider by having other pulses added to their trailing ends in order to widen them to the 0.43 H as specified for the vertical synchronizing pulses. During the region which is to contain only the normal 13,230-per-second horizontal pulses the alternate unwanted pulses are keyed out and the remaining pulses are widened to 0.08 H by adding other suitable pulses to their trailing ends. In each case the leading edges of the original pulses are not altered by the additions. Since any slight change in the shape or slope of the leading edges or delay in their transmission through the circuits will be the same for all types of pulses, no relative errors will be introduced. The widths of the various composite pulses will vary somewhat with unavoidable changes in the timing of the added pulses since they are produced in separate circuits, but considerable tolerance is permissible in this respect. The widening pulses are always added to the 26,460-c.p.s. pulses with an appreciable overlap in time so that no gap in the completed pulses is ever present after a final step of limiting amplification. The principles involved will become more evident when the specific apparatus is described.

Frequency Division vs. Frequency Multiplication

Considered casually it would seem better to begin the frequency chain with the 60-cycle power supply and use frequency multipliers in the various steps to obtain the 26,460-c.p.s. signal, since it would avoid the indirect method of obtaining 60-cycle synchronism and the relatively complex frequency dividers as were used in the later signal generator to be described below. The objection to this simpler method resides in an inherent weakness of the frequency multipliers themselves, namely that the instantaneous-output frequency of a frequency multiplier is not necessarily correct and in close agreement with the average frequency at all times as will be explained. In this type of device the lower frequency sine-wave signal is greatly distorted by an amplifier tube to produce harmonics, and the desired harmonic is then presumably selected and isolated in the plate circuit by a tuned circuit. The operation is imperfect because the tuned circuit can not readily be made to have sufficiently low loss to be adequately selective to isolate the desired harmonic completely. Furthermore even if it were adequately selective it would not be capable of following the slight frequency changes in the 60-cycle power system. (If a flat-top band-pass filter were used the change in phase with frequency might be objectionable if sharp cut-off is obtained.)

From a physical point of view the lower-frequency input signal merely shock-excites the tuned-plate circuit once for each lower-frequency cycle and leaves the tuned circuit to generate say 7 or more cycles by free damped oscillation. Between shocks the frequency may be slightly off the ideal since it is determined only by the tuned circuit. Figuratively speaking it may be said that a 60-cycle per second wave, for example, measures or divides time into 1/60-second units. Then in attempting to produce by frequency multiplication a frequency of say 420 cycles per second we have the difficult task of measuring 1/420-second intervals by a "ruler" calibrated only in units seven times as large. The experimental efforts of others seem to support these conclusions.

FREQUENCY DIVIDERS

Two very different types of frequency-dividing circuits are available: the multivibrator* which, when driven by a higher frequency, may have its natural frequency adjusted so that firing occurs only on say every seventh pulse, and the pulse-counter circuit† which accumulates the effect of several consecutive cycles of pulses without regard to their frequency and fires producing a single pulse output when the accumulated effect is adequate.

* The blocking oscillator or other forms of self-running relaxation oscillator may be used instead of the multivibrator. The multivibrator is however, the preferred form of this general type.

† The Electric Music Industries, Limited of Great Britain is credited with developing the pulse-counter circuit for synchronizing-signal generators.

Certain advantages in stability lead to the adoption of the "counter" type of circuits for the synchronizing generator in spite of its greater complexity,

With reference to Figure 2(a) the "counter" circuit for frequency dividing may be explained as follows: Since the first frequency divider in the chain has been chosen for the explanation, V_1 is shown amplifying rectangular waves derived from the master 26,460-c.p.s.

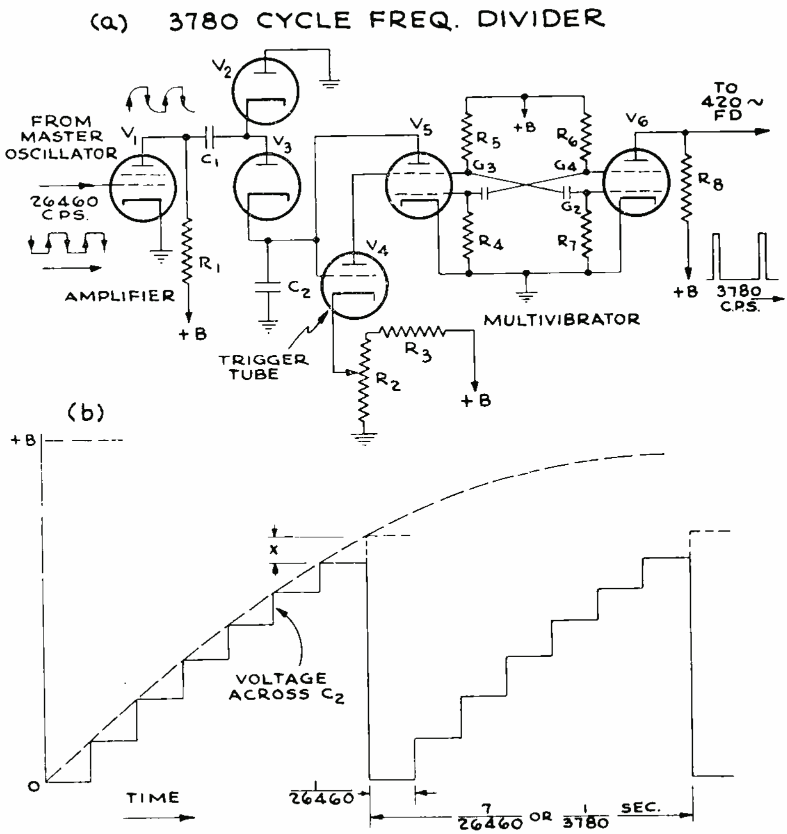


Fig. 2—(a) Frequency Divider. (b) Shows how C_2 is charged in steps through V_3 and C_1 and discharged at intervals by V_5.

oscillator by limiting amplifiers (as shown further in Figures 3 and 4). The plate resistance R_1 is high and the grid swing of V_1 is of saturating magnitude so that the output-rectangular wave has amplitude limited by and approximately equal to the voltage of the power supply + B. Starting with no charges on C_1 and C_2, the plate voltage swings to + B and the capacitors C_1 and C_2 charge in series through diode V_3 to substantially the entire voltage. The + B voltage of say 250 volts is divided between the two condensers inversely as their respective capacitances. C_1 is small and C_2 large so that C_2 has the lower voltage, say 1/20 of + B which is 12.5 volts.

On the negative stroke diode V_2 conducts, discharging C_1 to ground, but not altering the charge on C_2. On the next positive stroke the conditions are repeated except that this time only approximately 225 net volts are available for adding *new* charges to the two condensers. Condenser C_2 then will obtain a second incremental charge of only about 10 volts. On the next negative stroke of the plate, diode V_2 will again discharge C_1. The "stair-step" rise of voltage across C_2 is shown at (b) in the figure. Note that the voltage across C_2 would asymptotically approach the +B voltage as the voltage increments decrease in amplitude if not interrupted. During the "build-up" time trigger tube V_4 is biased beyond "cut-off" due to a definite portion of the positive voltage from the +B supply applied to its cathode. When the "stair-step" voltage across C_2 reaches slightly higher than the cut-off condition for the trigger tube, depending upon the setting of the potentiometer R_2, the trigger tube conducts, driving the control grid of the multivibrator tube V_6 negatively. Tubes V_5 and V_6 are connected in a conventional multivibrator circuit using the screen grids as anodes for the multivibrator action so that the plates are available for other purposes to be explained. The multivibrator constants are such that tube V_5 would remain in the cut-off portion of the cycle for extremely long intervals if it were not for the pulses received from the trigger tube V_4. When the trigger tube conducts, the multivibrator is triggered, tube V_5 conducts and its plate circuit quickly discharges condenser C_2 substantially to zero potential. Then the multivibrator re-sets to its initial condition where V_5 is cut-off and V_6 conducts. The "stair-step" charging cycle starts again and the charge due to the next seven cycles is metered by the trigger tube, and so on. The plate output of tube V_6 is a 3,780-c.p.s. rectangular wave of suitable amplitude to charge a similar "C_1 and C_2" of the next "counter" circuit (or frequency divider).

This frequency-dividing circuit is entitled to be called a "pulse-counter" only on the grounds that it produces one output pulse for every certain number of input pulses over a wide-frequency range. In the present application the ability of the output to "follow" the input for a large frequency change is of no great value, except that it simplifies changes for experimental purposes. The real advantage of using the "counter" circuits instead of the multivibrators alone is their very great stability in counting accurately with changes in tube characteristics and +B voltage. The magnitude of each step in the stair-step charging of condenser C_2 tends to vary in proportion to the +B voltage. The bias voltage on the cathode of the trigger tube, which responds to the accumulated voltage across C_2, also varies in proportion to the +B voltage so that approximate cancellation of the effects of B—voltage changes is obtained. Some tendency to error in "counting" occurs due to the variable resistances of the diodes, which prevent complete charging of C_2 through V_3 and the complete alternate discharging of C_1 through V_2. Also tube V_5 does not discharge C_2 to completion due to tube resistance. However, these effects may be made negligible

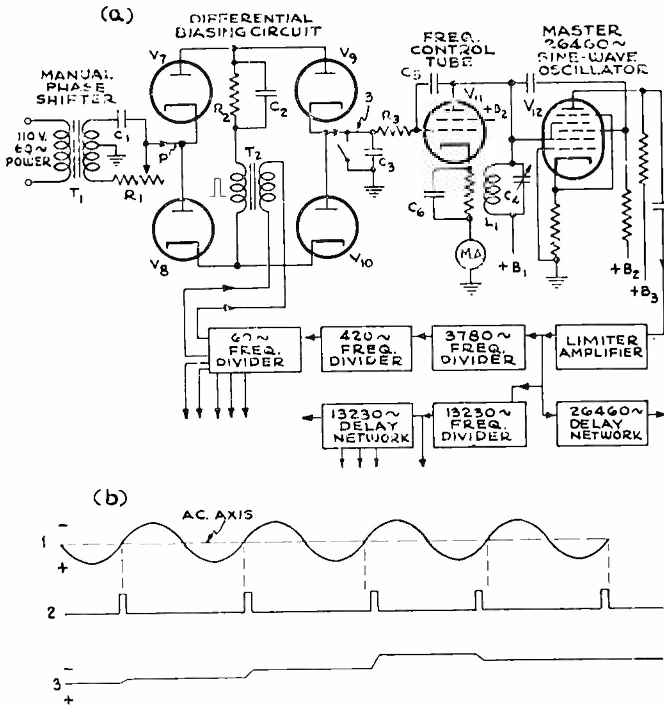


Fig. 3—(a) 60-cycle Locking Circuit. (b) Wave 1 from the Phase Shifter and wave 2 from the 60-cycle Frequency Divider produce frequency control wave 3 at 3 in the circuit. Note wave 3 is uniform except when correction is required.

by using circuit constants of relatively high impedance. Also the cut-off voltage of tube V_4 may vary incorrectly with plate voltage and from tube to tube, but this voltage is relatively small, especially in the high-mu-type tubes. Another and more-serious source of irregularity was found to be due to gas current and leakage in tubes V_4 and V_5 when certain unfavorable types were tried for these circuit positions. The merit of this type circuit is due to the fact that to a certain extent the tubes act only as switches so that stability is largely determined by the relatively-stable condensers and resistors.

Sixty-Cycle Locking Circuit

The frequency-regulating circuit for maintaining the entire synchronizing generator in synchronism with the 60-cycle power system is shown in Figure 3(a), in which each frequency-dividing circuit is represented by a block having the lower or output frequency indicated. The 26,460-c.p.s. oscillator is of the negative-transconductance type and electronic coupling to the plate circuit is used for output. The frequency is determined by the tank circuit which includes the constants L_1 and C_4 and the automatically adjustable impedance due to the plate current of the frequency-control tube V_{11}. Since the grid of this tube is excited from the tank circuit, through a small condenser C_5, which provides phase shift, the plate current is largely in quadrature so the mutual conductance of this tube will affect the resonance frequency. The bias of tube V_{11} and, hence, its mutual conductance is controlled by the "differential biasing circuit" in accordance with the relative phases of the 60-cycle power circuit and the "60-cycle" pulse output of the last frequency-divider of the chain.

The differential-biasing circuit can best be understood by considering the entire bridge circuit comprised by the four diodes, V_7, V_8, V_9 and V_{10}, the condenser C_2, the resistor R_2 and transformer T_2, as merely a key or switch that momentarily connects the 60-cycle power supply through the manual phase shifter to the condenser C_3. The narrow 60-cycle pulses as shown at 2 in Figure 3(b) introduced by transformer T_2 into the bridge cause the four diodes to conduct briefly for each pulse. This conduction charges condenser C_2 so as to retain the diodes entirely biased off between pulses while resistor R_2 continually discharges the condenser slightly so that the diodes will continue to conduct during the pulses. The pulses from the frequency divider as shown at 2 in Figure 3(b) occur near the time the sine-wave power-supply voltage (wave 1) impressed at P crosses the a-c axis from positive to negative and has a maximum rate of change. Hence, slight changes in the relative timing of the frequency-divider circuit and the 60-cycle power line will cause the voltage accumulated upon condenser C_3 to vary considerably as shown by wave 3. Since the sine-wave voltage at P is changing negatively during these pulses, a lagging condition of the frequency divider for example, will adjust the bias on condenser toward the negative, making the frequency-control tube less conducting. This in turn tends to cause the master oscillator to increase its frequency, thereby decreasing the lag of the entire chain of frequency-dividers. Conversely an advancement of the phase of the pulses produces a retarding influence upon the master oscillator. Hence, when the tank, L_1, C_4, is set manually such that its natural frequency would be approximately 26,460-cycles per second, the synchronizing generator will be automatically adjusted to synchronism with the 60-cycle power supply. Also if the manual setting is made such that balance is obtained with the 60-cycle pulses occurring very nearly the time the sine-wave passes through zero, the locking phase relation will not vary appreciably with the amplitude of the power wave. The adjustment necessary for this condition may be determined practically by the milliammeter in the cathode circuit of the frequency-control tube.

The time required for a frequency adjustment to occur after it is needed is very short, but no appreciable 60-cycle changes in voltage occur across C_3 since C_3 can neither lose nor gain charge between the narrow 60-cycle pulses as indicated by Wave 3 being uniform between pulses of Wave 2. Note also that if no frequency correction is needed the control-voltage Wave 3 is uniform d.c.

These advantages were not present in an early form of regulating circuit used experimentally in which the 60-cycle power wave was beat with the 60-cycle pulse wave and rectified to produce variable height controlling pulses. In that circuit the pulses were filtered by R-C circuits to change the pulses to a variable d-c control voltage. Difficulty was experienced in that when the filtering was made adequate to avoid excessive 60-cycle frequency modulation, the control action was sluggish and subject to hunting or over-swing.

Wave-Shaping Circuits for Auxiliary Output Waves

Figure 4 shows the entire synchronizing generator in block diagram. In the left-hand portion of the figure the chains of frequency dividers are indicated by blocks containing the letters FD and a number corresponding to the cycles-per-second output. The block marked "L-26,-460" is the limiter amplifier shown in Figure 3 and supplies the delay network DN-26,460 with 26,460 rectangular pulses per second. Similarly the frequency divider FD-13,230 is a source of synchronized 13,230-cycle rectangular pulses which supplies another delay network DN-13,230 having several different output taps. The frequency divider FD 60 (shown in Figure 3 also) is a 60-cycle synchronous pulse source for synchronizing five separate multivibrators shown as blocks "MV". These 60, 13,230, and 26,460-cycle-per-second sources are used to produce all the synchronizing-generator output signals.

For example the output signal known as "video blanking" is a mixture of 60-cycle-per-second pulses and 13,230-cycle-per-second pulses with the 13,230-cycle pulses eliminated during the occurrence of each 60-cycle pulse. The letter "*d*" in Figure 4 indicates the conductors for this wave and the wave shape is shown at "*d*" in Figure 5. The component waves *a* and *b* are mixed in the mixer-limiter, block ML-1 of Figure 4, to provide the sum wave *c*. Wave *c* is limited in the same block at the level of the broken line to produce the wave d which is supplied and transmitted at 75-ohm impedance by the line amplifier, block LA-1. In each case the small letter adjacent the conductors in Figure 4 corresponds to the wave shape present in that portion of the circuit as shown in Figure 5. Figure 6 shows the essential circuit elements (which are

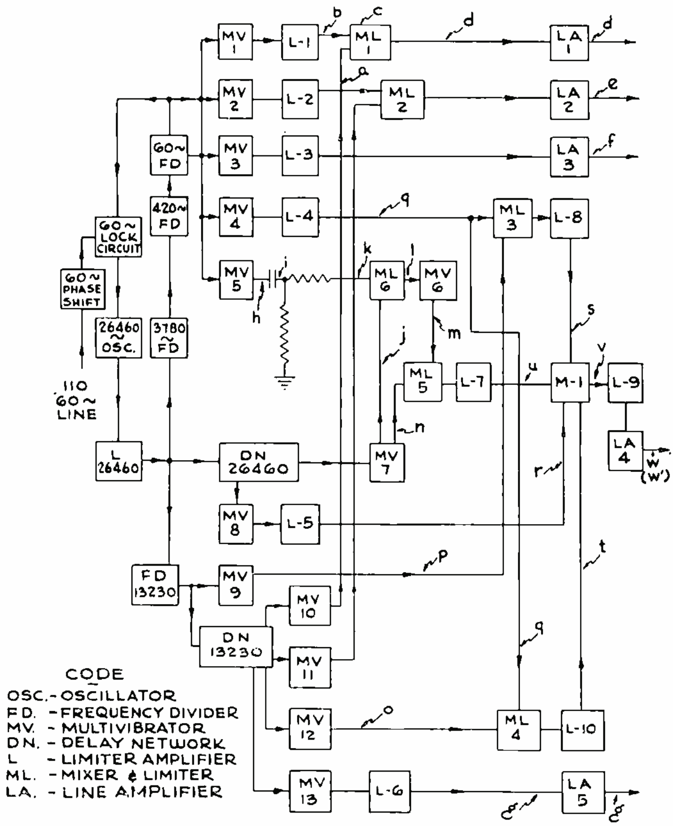


Fig. 4—Block diagram of entire Synchronizing-signal Generator. Letters a, b, c, etc. refer to wave shapes in Figure 5. Wave w (w') represents R.M.A. Synchronized output wave while d, e, f, and g are used for auxiliary equipment.

represented by the blocks in Figure 4) for producing the "video blanking" wave d as explained. Tube V_{13}, the output stage of the 60-cycle frequency divider (FD 60) synchronizes the multivibrator (MV-1) via the amplifier-buffer stage V_{14}. The time constant (RC) in the grid of the first triode of Tube V_{15} is very long compared to that of the second triode of the tube. Hence, the first triode remains cut-off for longer periods and is caused to conduct by negative pulses applied upon the grid circuit of the second triode. The brief conduction period of the first triode of V_{15} can be adjusted by the potentiometers which controls the positive bias on the second triode so that the duration of the positive pulses impressed upon tube V_{16} can be made as long as desired in the wave *b* of Figure 5. Tube V_{16} serves as a mixer since its plate is in parallel with the plate of the second triode in tube V_{18}. Tube V_{16} also limits its positive plate swing by cut-off and its negative plate swing by drawing grid current.

The multivibrator MV-10 which includes Tube V_{17} is synchronized by 13,230-cycle-per-second pulses from the frequency-divider, FD 13,-230, which have been properly delayed by the delay-network DN-13,230. The multivibrator output-wave *a* is combined with wave *b* by means of the second triode of V_{18} to form voltage wave *c*. (Actually waves *a* and *b* are present as current waves in the plate leads while the voltage on the plate of either Tube V_{16} or the second triode of V_{18} is wave *c*.) The first triode of Tube V_{18} and the line-amplifier Tube V_{19} serve as limiting amplifiers to reduce wave *c* to wave *d* by saturating-off the portion of wave *c* above the broken line. The line-amplifier LA-1 is connected for output from its cathode circuit as this provides a lower impedance for operation into a 75-ohm distribution cable.

The "Iconoscope blanking" output wave and the apparatus for producing it is essentially the same as for the "video blanking" wave except that the pulses are shorter and are delayed different amounts in the synchronizing generator due to the Iconoscope blanking being subjected to additional delay by transmission through the television camera cables. The wave is shown at *e* in Figure 5 without the additional delay and the circuit apparatus involved in its generation may be determined by following backward along the lines from the letter *e* in Figure 4.

The "vertical-driving signal for the Iconoscope" is a simple 60-cycle-pulse wave as shown by wave *f* in Figure 5. Its generation involves only one multivibrator MV-3 (synchronized by the 60-cycle frequency-divider), a limiter and a line amplifier as shown in Figure 4.

The "horizontal driving signal for the Iconoscope" is a similar kind of wave involving a similar type of apparatus except that the frequency of the pulses is 13,230 per second as shown by the wave *g* in Figure 5 and the letter "*g*" in Figure 4.

Circuits for Generating the R.M.A. Synchronizing Wave

The last signal output of the synchronizing generator is the "R.M.A. synchronizing wave" as indicated at *w* and *w′* in Figure 5 for the intervals near the even and odd vertical pulses respectively. (*W* and *w′* are views of the same voltage wave taken 1/60-second apart.) For clearness the number of equalizing pulses and the duration of each vertical pulse shown in w and w' has been halved as can be seen by comparing with the R.M.A. standard drawing T-111 of Figure 1. Since several unusual methods are employed to insure a very high degree of accuracy in the wave at its critical points, a brief review of the steps is of interest.

All of the waves from h to v in Figure 5 are generated and used in various combinations to produce the final wave w. In the last step, wave w is obtained from wave v alone by simply limiting or "clipping" at the positive and negative levels indicated by the broken lines on wave v. Wave v on the other hand is derived by adding the four waves

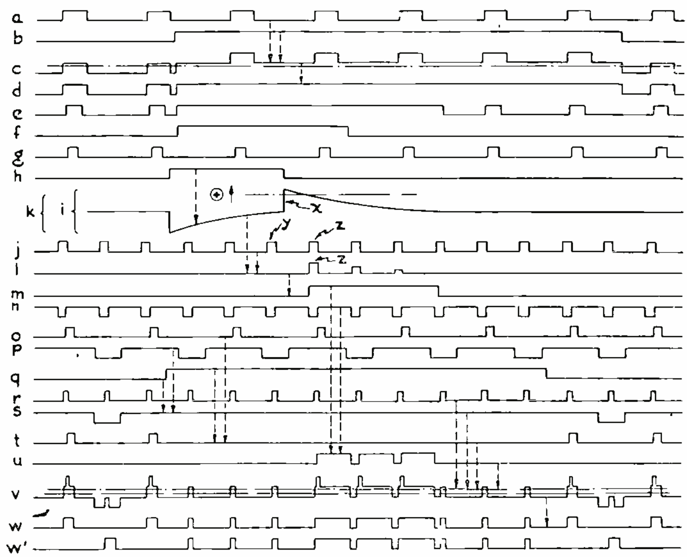


Fig. 5—Wave shapes present in circuit of Figure 4 at points indicated. Wave v, from which R.M.A. synch wave w is derived, is the sum of r, s, t, and u as shown by dotted arrows.

r, s, t, and u in mixer M-1 as indicated in Figure 5 by the vertical dotted arrows connecting the several waves. Wave r is a simple 26,460-cycle pulse wave produced by a delay-synchronized multivibrator, MV-8, and a limiter, L-5. Wave s is the 13,230-cycle pulse wave p after having a group of the pulses keyed-out by each pulse of the 60-cycle wave q. The keying is accomplished in the mixer-limiter, ML-3, of Figure 4 by applying the two waves p and q respectively to the first and third grids of a tube of the type commonly known as a pentagrid converter. When wave p allows electrons to pass the first grid, the wave q modulates their flow to the plate output circuit whereby one wave modulates the other. Wave t is produced in a similar manner by allowing wave q to key wave o in the mixer-limiter ML-4. The waves p, q, and o originate in multivibrators synchronized by suitably-delayed pulses.

Wave *u* is obtained by the 60-cycle pulse wave *m* keying the 26,460-cycle wave *n* in the mixer-limiter, ML-5, with such polarity that the high-frequency pulses are passed only *during* each 60-cycle pulse.

The pulse of wave *m* must be delayed with considerable accuracy with respect to the output of the 60-cycle frequency-divider in order to insure the leading pulse of each 60-cycle group of pulses in wave *u* being a whole pulse. The main delay is obtained indirectly by using the back or trailing edge of the pulses of wave *h* instead of a delay network as will be explained. Wave *h* is distorted to a shape such as shown at *i* by transmission through a small condenser with a resistance

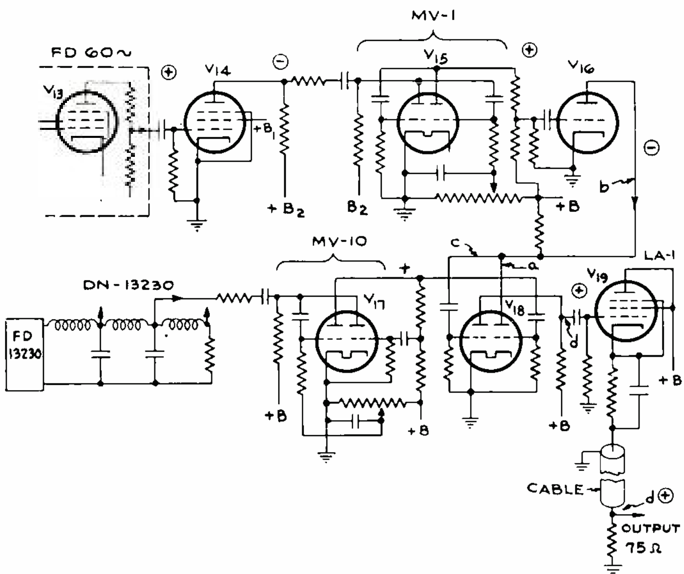


Fig. 6—Multivibrators, mixer, limiter, and line amplifier for combining waves *a* and *b* to form wave *c*, which is limited to produce *d*.

output load. It is then applied to the first grid of a pentagrid tube through a series resistor which limits wave *i* at the broken line due to grid current, producing wave *k*. Wave *j* is applied to the third grid of the tube and the limited wave *k* holds the tube cut-off except when pulse *x* of wave *i* occurs. Then the plate-output circuit contains small isolated 60-cycle groups of the 26,460-cycle pulses as shown in wave *l*. The leading pulse of each 60-cycle group of wave *l* then accurately triggers the multivibrator MV-6. This multivibrator produces the 60-cycle-pulse wave *m* which is used as described above. Wave *m* is, therefore, accurately timed with respect to the 26,460-cycle-pulses of wave *n* since waves *n* and *j* are outputs of the same multivibrator. It should be noted that the trailing end of the pulse of wave *h* could occur any time be-

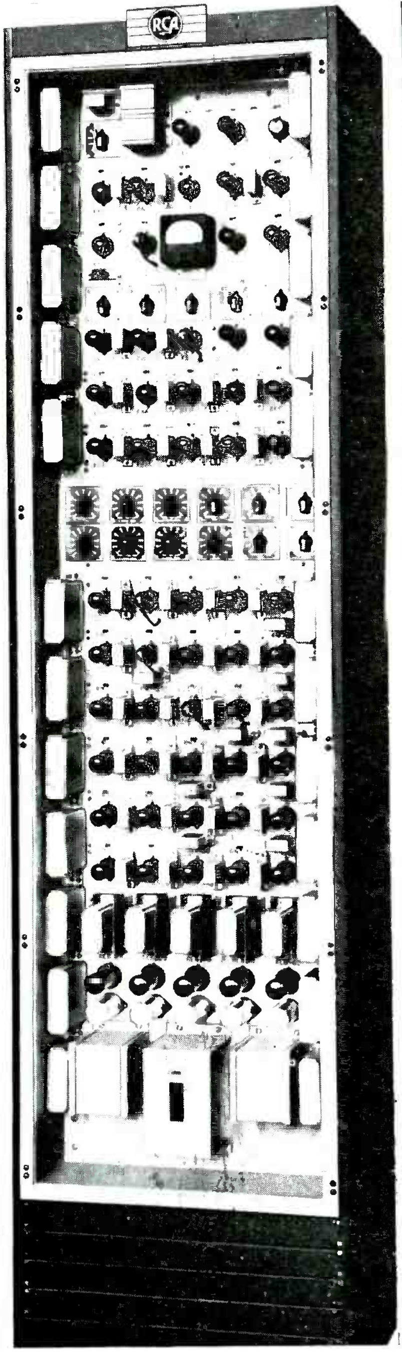


Fig. 7—Front View of Complete Synchronizing-signal Generator.

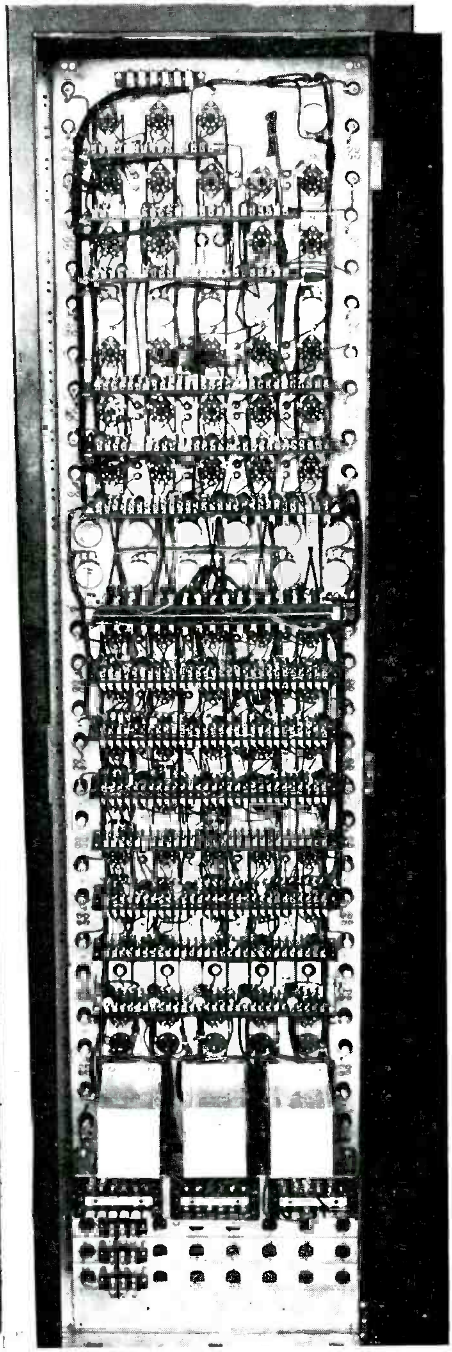


Fig. 8—Rear View of Synchronizing-signal Generator.

tween the pulses y and z of wave j and still select the same pulse z as the leading pulse in the group in wave l. Such tolerance in operation allows reliable accuracy in performance without critical adjustments.

As mentioned before, the extreme uniformity of all leading edges in the finished synchronizing wave w is due to the fact that the same oscillator produces all pulses of wave r, which provide the leading edges of all the pulses of various widths in the finished wave. The unwanted pulses of wave r are extinguished by wave s and certain other pulses are made wider by adding to their trailing end the pulses of waves t and u. If it had been attempted to add these pulses so that no over-lap was caused, i.e., as a sort of "butt point" the relative timing and wave shapes would have to be adjusted to extreme accuracy to avoid notches and gaps in the sum pulses. Study of Figure 5 shows that by using a "lap-joint" between adjoining pulses and by subsequent clipping, considerable tolerance in the timing and shape of the various component waves may be allowed.

Practical Features

Figures 7 and 8 respectively show front and rear views of the synchronizing generator mounted in a cabinet rack. All of the 62 tubes (most of which are double tubes) and the controls are on the front of the single chassis, and are made accessible by opening the hinged door. Most of the smaller components are mounted on bakelite terminal boards and are readily accessible as seen in the rear view. The wiring has been greatly simplified by carefully grouping the apparatus so that leads are very short. This also avoids the need of shielded wires except in a very few connections. Adequate electric isolation of the various parts from one another is obtained by locating the parts in suitable groups on the chassis. Of course all circuits which contain pulse signals having steep wave-fronts, corresponding to several megacycles, are relatively low-impedance circuits in order to provide fidelity and prevent cross-talk.

The tube heaters are supplied by the five 60-cycle transformers near the bottom of the chassis, two in front and three in the rear. The plate supply required is 770 milliamperes at 250 volts, which is usually supplied from two external regulated power rectifiers operating on 110 volts a.c. (320 milliamperes of this current is used in the five line amplifiers for distribution about the studios.)

Since a failure of the synchronizing generator in a commercial television installation might cause a serious interruption or impairment of service, great attention was given to attain reliability and *inherent* accuracy (rather than accuracy which is dependent upon critical adjustments of controls). The use of relatively complicated circuits in the signal generator which require the use of many tubes (with the resulting increased number of chances for tube failure) might seem to decrease the reliability. However, the circuits were chosen to permit the operation to be unaltered by very large changes in tube characteristic without readjustment of controls. Therefore, routine checking of tubes should avoid failure by the normal deterioration of tube characteristics. Furthermore, abrupt structural failure of tubes is relatively rare when, as in this case, all tubes are operated conservatively within their rating.

Similarly, a number of variable controls are provided as an aid to reliability as they allow easy periodic adjustment to optimum mean positions using an oscilloscope for an indicator. This insures that gradual changes in the circuit elements will not likely cause failure. They also permit some changes for experimental purposes.

Considerable experience with the several factory-built synchronizing generators of the type described, indicates that excellent results may be expected if reasonable care is used in routine maintenance.

ACKNOWLEDGMENT

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FIELD-STRENGTH MEASURING EQUIPMENT AT 500 MEGACYCLES

By

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Summary—*The problems encountered in ultra-high-frequency field-strength measurement are discussed. Equipment for making measurements at frequencies on the order of 500 megacycles is described which is adapted to the basic method involving the use of a half-wave dipole-receiving antenna having known constants, and a signal generator which can be substituted directly for the antenna to duplicate the signal voltage induced in the antenna. A method of making peak-signal or noise measurements is briefly described.*

METHODS used in making field-strength measurements at low frequencies can be successfully applied at ultra-high frequencies, but as may be expected, the difficulties increase rapidly with frequency. The methods and equipment to be described are believed to yield a reasonable degree of accuracy because of the fundamental considerations involved. The most important elements in such a measuring system, including the antenna system with transmission line and the standard signal generator with their related problems, will be discussed in some detail. Briefly, the method chosen uses a standard signal generator which is connected to the transmission line in place of a half-wave dipole and is adjusted to deliver the same signal strength to the transmission line as did the antenna, this equality relation being indicated on a suitable receiver connected to the output of the transmission line. The signal generator has an internal impedance substantially equal to the radiation resistance of the half-wave dipole which is known to be nearly 75 ohms. The calibration of the signal generator yields the induced voltage which is equal to the voltage induced in the antenna. The voltage induced in the antenna and its known effective height then yields the field strength in volts per meter which is the antenna voltage divided by the effective height of the antenna. A desirable feature of this system is that the receiver and transmission line are used only as a comparative voltmeter and, therefore, their characteristics are relatively unimportant. Modifications of this method are sometimes desirable in practice, with the result that some care must be given to the choice of the transmission line and its termination at the receiver.

Antenna Systems

For ultra-high-frequency measurements of field-strength, the half-wave dipole is most convenient since its free-space constants of resistance and effective height are known. Practically, of course, free-space conditions cannot be obtained, but the error due to use of the free-space constants of the dipole is usually inconsequential when reasonable care is used in selecting the location of the antenna. The problem then is to determine the voltage induced in the dipole. The most direct procedure is to replace the antenna with a signal generator having the same impedance as the dipole and introduce in series with this impedance the same voltage as was induced in the dipole to obtain the same receiver response. The accuracy of the voltage calibration of the signal generator then in a large measure determines the accuracy of the measurement.

Transmission-line characteristics and receiver gain must be stable during the period of measurement, since these factors affect the resultant accuracy. It is generally desirable that the antenna be matched to the transmission line and that the receiver properly terminate the transmission line, but it will be apparent that these factors can be imperfect without appreciably impairing the accuracy of field-strength measurements by this method. It is only necessary that the signal generator have the same internal resistance as the antenna and that the transmission line be properly coupled to the receiver so that voltage picked up on the line cannot combine with the voltage delivered by the antenna. Measurements can be facilitated by measuring the line loss with the signal generator at the desired frequencies and subsequently duplicating the received signal by connecting the signal generator directly to the receiver. In this case the voltage at the antenna is obtained by considering the voltage at the receiver input and the line loss.

Low-impedance, twisted-pair transmission lines up to about 100 feet long are usually satisfactory for use at frequencies as high as 150 megacycles. This type of line is unshielded, requires the use of a well-balanced receiver-input coupling or an electrostatic shield placed between the coupling coil and the first tuned circuit. The use of coaxial transmission line is almost a necessity at higher frequencies because of its perfect shielding and reasonably low loss. It is somewhat fortunate that satisfactory coaxial lines having about 75 ohms characteristic impedance are available as this impedance line can be terminated readily by a coaxial resonant-input circuit in a receiver, and requires no critical-matching transformer to match to a half-wave doublet antenna. A single coaxial line is preferred for practical reasons. The two halves of the dipole are connected to the inner and outer conductors of the line. In this case it is desirable to assure the antenna balance by some means. Figure 1 shows one method which by virtue of the quarter-wave-long sleeve placed around the coaxial line, makes the end of the line have relatively high impedance to ground. Thus, the dipole antenna connected as shown is at high impedance to ground and each half is practically balanced. This balance also prevents unbalanced currents from flowing in the halves of the antenna as a result of currents induced in the outer conductor of the transmission line.

A certain accepted type coaxial line seems to withstand reasonable handling and has been used even in short lengths with suitable end-

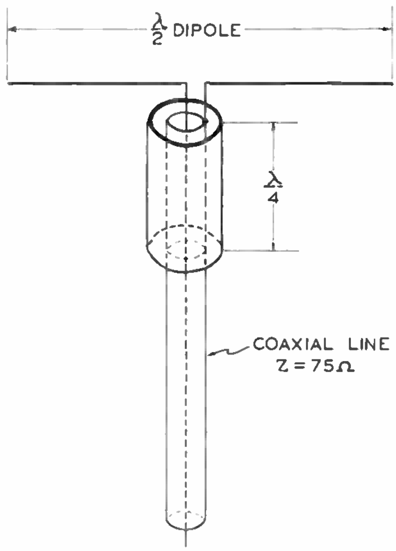

Fig. 1—A standard dipole connected to a coaxial transmission line.

fittings for "patch cords". In this way fairly uniform coaxial connections can be had which are "smooth" electrically at frequencies over 500 megacycles. This line has a loss of about 3.5 decibels per 100 feet at 450 megacycles.

Signal Generators

The importance of electrically "smooth" connections in measuring work cannot be overstressed especially at frequencies above 100 megacycles. Using a small commercial 75-ohm terminating resistor, a satisfactory termination of a 75-ohm line is obtained. In the case of the signal-generator-output attenuator shown in Figure 2, the 75-ohm coaxial line is carried up to the 75-ohm resistor which with about ¼ inch of lead, forms the coupling loop. This added lead in the coupling loop tends to mis-match the line termination appreciably at 500 megacycles and in some cases may make it desirable to calibrate the instrument in terms of voltage delivered to a 75-ohm load, or the characteristic impedance of the desired transmission line. In this case, the equivalent voltage induced in the antenna is equal to twice the voltage delivered to a 75-ohm transmission line.

In Figure 2 are also shown the essentials of a signal generator including the output-line termination, which are incorporated in such an instrument operating up to nearly 600 megacycles. A shielded oscillator supplies a calibrated reference voltage E to the primary inductance L_1. Near the low-voltage end of L_1 is placed the outer cylinder of the attenuator. The inside diameter of this cylinder is small compared with the shortest wavelength used; hence, the voltage induced in L_2 inside the cylinder varies as a logarithmic function of

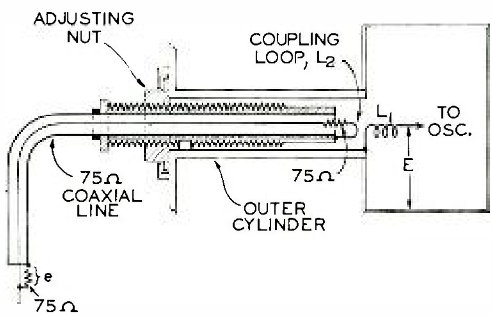

Fig. 2—Attenuator detail and elements of a signal generator.

the distance between L_2 and the end of the cylinder near L_1. A very useful feature of this type of attenuator is that the attenuation law is substantially constant regardless of the applied radio frequency. The chosen diameter of the outer cylinder is about 11/16 inch which in practice gives an attenuation of about 47 decibels per inch along the axis of the cylinder. By threading the movable element carrying L_2, and driving it with an indicating nut, this attenuator can be set with an error of less than 0.1 decibel. Thus, it will be seen that the practical accuracy of this signal generator is dependent on the adjustment and measurement of E across L_1, to the values for which the instrument is calibrated. As will be shown later, the absolute value of E is unimportant, but it is desirable to use a means of measuring E which is reliably constant and preferably has no frequency error. Up to about 200 or 300 megacycles, the vacuum-tube voltmeter is highly satisfactory and requires but one calibration good for this and lower frequencies. This is true because the voltage induced in L_2 is proportional to the voltage E across L_1 for all frequencies where L_1 is electrically short compared with the wavelength. Diode voltmeters are still stable and fast-acting at the higher frequencies and therefore useful, but require that the instrument be calibrated at a number of frequencies. The reference voltmeter measuring E is placed in a separate shielded compartment and connected with L_1 by the lowest possible reactance path.

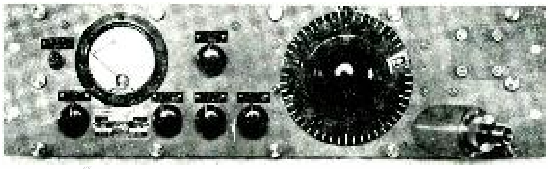

Fig. 3—U-H-F signal generator, 400 to 580 megacycles. Output continuously variable from about 0.1 volt to less than 1 microvolt.
Output impedance 75 ohms.

Absolute calibration of this signal generator is had by comparing the output voltage e, with a known voltage. Then if desired, the voltage in L_2 can be calculated if the impedance of L_2 and its series resistance are known. The known or standard voltage can be measured by several means; an ultra-high-frequency thermocouple is probably one of the best available at the present time.

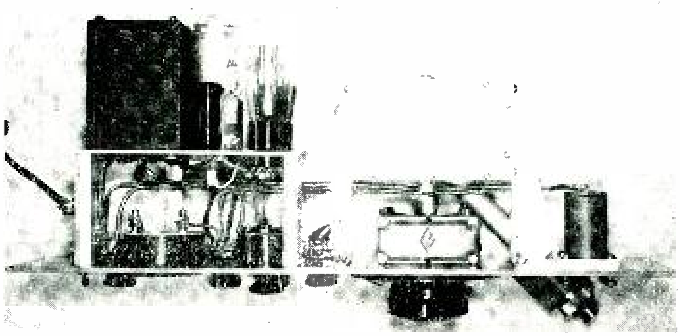

Fig. 4—U-H-F signal generator shown in Figure 3 with cover removed.

In Figure 3 is shown an example of a signal generator covering a continuous frequency range of 400 to 580 megacycles and conforming with the principles discussed. The maximum output voltage e, is on the order of 1/10 volt. A replaceable attenuator element to give balanced output to a low-impedance two-wire line is also provided for this instrument. This of course requires a separate set of calibration data. The frequency can be set by the calibration to within 1 megacycle of the desired value and is relatively easily adjusted to give an audible beat note with the local heterodyne oscillator in a receiver. A trace of frequency modulation is present when using a.c. on the oscillator heater. The frequency is changed by means of a conventional condenser which can be rotated through 180 degrees by a suitable speed-reducing mechanism. Either a-c or battery-power supply is contained within the unit.

In Figure 4 is shown the signal generator with the outside cover removed. The output of the attenuator projects from the casting at an angle through the panel on the right. The casting houses the oscillator and diode voltmeter, both acorn-type tubes, and filter elements for the supply leads. The microammeter for the diode voltmeter is on

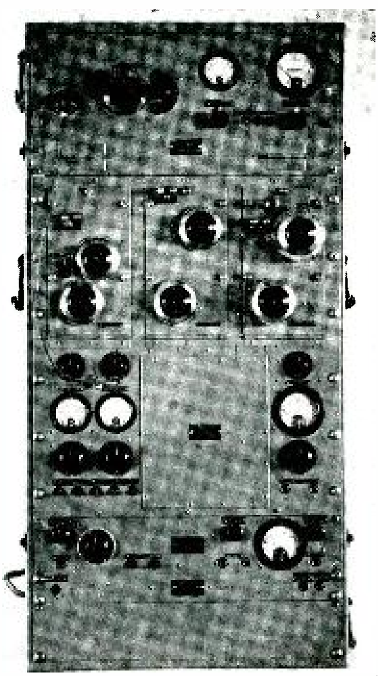

Fig. 5—60 to 500-megacycle receiver with auxiliary measuring equipment.

the left side and behind this is an a-c power supply including a voltage-regulator tube for the oscillator-plate supply.

Receiving Equipment and Methods of Measurement

In Figure 5 is shown an ultra-high-frequency receiver with other equipment useful in measuring work. The top unit is a signal generator similar in design to the one described, but covering a frequency range of from 30 megacycles to 200 megacycles. The second and largest unit is the receiver covering a frequency range of from 60 megacycles to a little over 500 megacycles. This frequency range is obtained by the use of a 40-megacycle intermediate-frequency amplifier in the lower half of the unit which can be connected to one of the three u-h-f converters in the upper half of the unit. The converter on the right covering from 250 to 500 megacycles is provided for either two-wire or coaxial-line input. The input to the i-f amplifier can be brought out and connected with the signal generator as shown, thus providing a convenient means for intermediate-frequency-amplifier measurements.

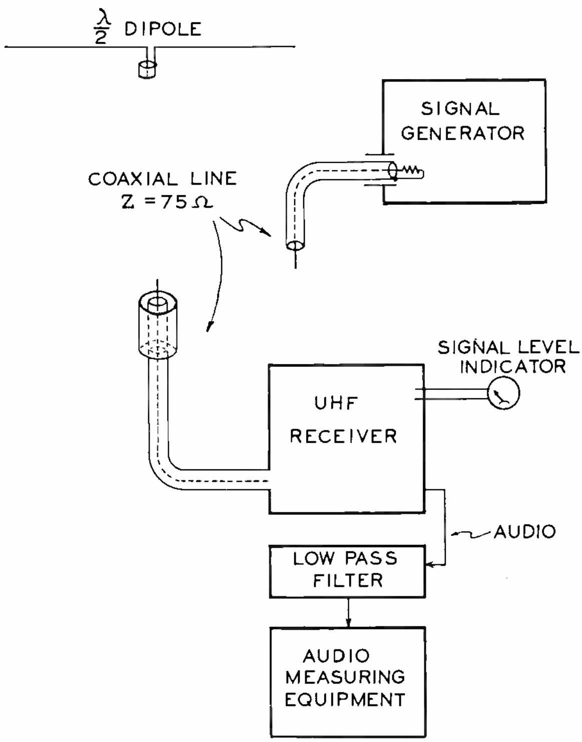

Fig. 6—Elements of an u-h-f field-strength measuring system.

In this intermediate-frequency amplifier the 40 megacycles is converted to a second intermediate frequency of 4.1 megacycles and then rectified with a diode, the diode current being indicated on a microammeter. A beat oscillator at 4.1 megacycles is also provided which is used in making peak measurements as will be explained. The next lower unit is an audio amplifier with an adjustable 80-decibel attenuator on the input and a rectifier-type meter on the output. This amplifier is also equipped with a tube operating at cut-off bias, the output of which, in headphones, is nil until the peak voltage just over-rides the bias, thus providing a means of measuring peak voltage. A 5-kilocycle low-pass filter is incorporated in the narrow strip panel below the amplifier.

A convenient method of making peak radio-frequency measurements is as follows with reference to the block diagram of measuring equipment shown in Figure 6. A known peak carrier is supplied to the receiver having preferably a strength somewhat greater than the peak signals or noise to be measured. The receiver gain is adjusted so that the strength of the signal at the diode is about 20 per cent or less of the value of the beat oscillator signal as indicated on the diode meter. The beat-note output of the receiver is fed to the audio-amplifier through the low-pass filter and the audio-amplifier gain is adjusted to give a peak voltage just over-riding the bias of the last tube as indicated in the headphones. This setting of the attenuator on the audio amplifier is the calibration for the peak value of the carrier supplied. Subsequent peak measurements of weaker signals can then be readily referred to this calibration point. For instance, with the known signal input to the receiver, the audio attenuator may indicate a 60-decibel loss inserted in the input to the audio amplifier in order to obtain a peak voltage which will barely over-ride the bias on the peak voltage measuring tube. With the unknown signal input to the receiver, the same output may be obtained with the audio attenuator having, say, 42 decibels loss. It will be apparent that, in this case, the unknown signal strength is 18 decibels lower than the signal strength used for calibration. If desired, a peak recording system can be used instead of the manual measuring system. The low-pass filter is especially important for noise measurements to provide a known band width for the measurement.

The high degree of flexibility in the various elements described enhances the usefulness of the equipment for functions other than the making of field-strength measurements. The measuring equipment and methods described were developed in the field laboratories of R.C.A. Communications, Inc., at Riverhead, New York, under the direction of Mr. H. O. Peterson.

U-H-F OSCILLATOR FREQUENCY-STABILITY CONSIDERATIONS

BY

S. W. SEELEY AND E. I. ANDERSON

RCA License Laboratory

Summary—Each of the circuit components in an oscillator may change its characteristics with change in temperature. Some of them are subject to change with changes in humidity and operating parameters. These effects are discussed with particular emphasis on change in inductance with change in temperature, since less attention has generally been given this cause of oscillator instability. An oscillator designed to have very small drift is described.

ONE of the principal problems in the design of receivers for operation on the ultra-high frequencies is that of oscillator stability. The magnitude of the problem is perhaps best indicated by the fact that on a percentage basis, 5,000 cycles drift at 40 Mc is the equivalent of only 125 cycles at 1 Mc. Because the most important services expected to occupy the u-h-f region, namely television and aural broadcasting, utilize wider channels than do lower frequency broadcast stations, it is probable that the absolute value of tolerable drift in cycles may be somewhat greater than is now considered satisfactory for the broadcast band. Nevertheless, greater stability in u-h-f receivers is needed than is necessary in the broadcast band.

A study of the problem has been made, and the practical results have indicated that any desired degree of oscillator stability may be achieved. Furthermore the increased stability may be obtained by straightforward engineering practice involving no unusual circuits or effects, and with readily available commercial components. Model oscillators and a complete receiver have been built for 40 to 42-Mc reception in which the oscillator drift has been limited to 1 kc. There is no indication that the results achieved in this laboratory work cannot be duplicated in production.

FACTORS CONTRIBUTING TO FREQUENCY INSTABILITY

The factors influencing oscillator frequency drift can be divided into three broad classifications, namely: heat, humidity, and operating parameters.

The effects of temperature variation on oscillator frequency are well known. Because most commonly used materials have a positive temperature coefficient, the result of an increase in temperature is an increase in the amount of L and C in the circuit, with a consequent decrease in frequency. The only components which may be expected to have a zero or negative temperature coefficient are some of the newer types of fixed capacitors.

The effect of an increase in humidity is also to lower the oscillator frequency, since water vapor has a higher dielectric constant than the air it displaces, resulting in an increase in the capacities of the system.

The operating parameters include the potentials applied to tube elements. These are subject to variation as the line voltage varies or as audio output causes fluctuation in the B supply.

Heat

The effects of heat are probably of greatest importance, since they are present under any circumstances. Even if it were possible to arrange the other circuit elements so that they would not be subject to temperature change, some effects of the change in temperature within the tube must inevitably be present. In a typical receiver of practical construction the temperature rise of carefully located components may be as much as 30° C. Any greater increase is probably the result of inadequate ventilation or improperly located components. Because fixed condensers, which are impervious to temperature changes, are readily available, their contribution to frequency change need not be discussed. All other components however, play a part in the total drift of the oscillator.

Inductance Variations

A simple expression for the inductance of a coil is as follows:*

$$L = \frac{r^2 n^2}{9r + 10l} \qquad \begin{aligned} L &= \text{inductance in microhenries} \\ r &= \text{radius of coil in inches} \\ l &= \text{length of coil in inches} \\ n &= \text{number of turns} \end{aligned} \tag{1}$$

From this formula we may derive various other useful formulas to evaluate the effect of temperature changes on an inductance. It is highly improbable that the radius and length of a coil will expand at the same rate if its temperature is changed. For example, if a coil is wound with large wire on a thin form or if the form has a coefficient smaller than that of the wire, the coefficient of expansion of the radius is likely to be that of the wire while the coefficient of expansion of the length is likely to be that of the form. Under other conditions both the wire and the form may be instrumental in determining the change in radius of the coil with temperature changes. If these effects can

* H. A. Wheeler, *Proc. of I.R.E.*, Vol. 16, Oct. 1928.

be evaluated, a formula for the change in inductance with change in temperature is useful.

Let A = coefficient of thermal expansion of radius in coil
B = coefficient of thermal expansion of radius of coil
t = temperature change in degrees C.

Then $L = \dfrac{r^2n^2(1+Bt)^2}{9r(1+Bt)+10l(1+At)} = \dfrac{r^2n^2+2r^2n^2Bt+r^2n^2B^2t^2}{9r+9rBt+10l+10lAt}$

Differentiate and set $t = o$

$$\frac{dL}{dt} = \frac{20lr^2n^2B + 9r^3n^2B - 10lr^2n^2A}{(9r+10l)^2} = L\,\frac{20lB + 9rB - 10lA}{9r+10l}$$

$$\frac{\triangle L}{\triangle tL} = B + \frac{10l(B-A)}{9r+10l} = \text{Coefficient of inductance vs. temperature} \qquad (2)$$

If the radius of the coil and length of the coil both expand at the same rate $(B = A)$ then:

$$\frac{\triangle L}{\triangle tL} = B \qquad (3)$$

or the coefficient of inductance vs. temperature is equal to the coefficient of expansion vs. temperature.

If the expression is solved for zero inductance change with temperature, it becomes

$$\frac{\triangle L}{\triangle tL} = o$$

$$\text{or } B + \frac{10Bl - 10Al}{9r+10l} = o.$$

$$\text{or } \frac{A}{B} = 2 + \frac{0.9r}{l}. \qquad (4)$$

In other words if we are to have a coil with zero change of inductance with temperature, the coefficient of linear expansion of the coil must be greater than twice its coefficient of radial expansion by an amount equal to nine-tenths of its radius divided by its length.

The coefficient of expansion of copper wire is 16×10^{-6} parts per degree C. Equation (3) indicates that the inductance of a copper-wire coil would increase linearly at this rate if both the radius and length of the coil were increased at the rate of expansion of copper. The change in frequency for small changes in inductance may be expressed as

$$-\left(\frac{\Delta f}{f}\right)=\frac{\Delta L}{2L} \tag{5}$$

Thus if the inductance of an oscillator coil is changing at the rate of 16×10^{-6} parts per degree C, the resultant change in oscillator frequency would be 8×10^{-6} parts per degree C. *Therefore at 40 Mc, this coil would produce a frequency shift of 320 cycles per degree or 9.6 kc for a 30° change.*

According to available data, the coefficient of expansion of paper base phenolic tubing as commonly used by radio manufacturers is from 17 to 25×10^{-6}. If a coil were wound on material having a coefficient of 17×10^{-6} the above results would be very nearly correct. If we assume a coefficient of expansion of 22×10^{-6} for the coil form, and a coil loose about the form so that expansion of the form will not increase the radius of the coil beyond that due to the expansion of the copper, but with the ends securely fastened so that the length of the coil will be determined by the form, solution of equations (2) and (5) indicates that a change in frequency of 6 parts per million per degree centigrade may be expected if $l = 2r$.

Equation (4) indicates that for equalization of inductance variations with temperature of a coil with a diameter equal to its length ($l = 2r$), B should be 39.2×10^{-6} if $A = 16 \times 10^{-6}$. No such material was available, but an experimental coil was wound on hard rubber which has a coefficient of approximately 80×10^{-6}. Using this material r/l should have been 3.33, an impractical shape, but a coil was wound with an r/l ratio of unity. This coil should have exhibited a negative-temperature coefficient of inductance of approximately minus 18×10^{-6} per degree C. The measured-temperature coefficient was minus 8×10^{-6}. This may have been due to the expansion of the form increasing the radius of the coil beyond that due to expansion of the copper, or because the coefficient of expansion of the form was less than the assumed value.

Data as furnished by the various manufacturers of phenolic tubing vary widely, so that information regarding the particular tubing to be used in any case must be obtained. In addition, there may be a considerable difference between the rates of axial and radial expansion.

The use of a type of wire with a very small coefficient of expansion was investigated. Invar and Nilvar, trade names for an approximately 36 per cent nickel and 64 per cent iron alloy, have a coefficient of expansion of less than one part per million per degree centigrade. A coil wound with this wire has an inductance change of less than 1 part per million and a frequency shift of less than ½ cycle per million cycles per degree centigrade. But a typical coil of this material has a Q factor of only 10 or so, as compared with a Q of 180 for a mechanically identical copper-wire coil. However, the Nilvar coil can be plated with some low-resistance metal such as copper. It will then have the physical thermal characteristics of Nilvar and the electrical characteristics of copper because of skin effect. For example, Nilvar wire with a diameter of 0.058″ was plated with copper to a diameter of 0.063″, providing a copper coating 0.0025″ thick. A coil of this composite wire had the same Q as a copper-wire coil of the same physical characteristics at frequencies as low as 10 Mc (the lowest frequency at which tests were made); thus at the higher frequencies a thinner coating of copper might be used.

A coil wound with this composite wire fully justified the calculated results. The change in inductance was so small as to be difficult to measure—substantially less than 1×10^{-6} per degree C.

Although the Nilvar wire is rather stiff, much more so than even hard-drawn copper, it may be necessary to mount the coil on a form to obtain the required mechanical rigidity. For this purpose a ceramic is indicated. Ceramics have suitable properties as u-h-f insulators, and in addition have the lowest coefficient of expansion (approx. 4×10^{-6}) of any economical, readily available material. Application of Equation (4) assuming $A = 4 \times 10^{-6}$ indicates that the Nilvar-Copper coil on a ceramic form should have a negative temperature coefficient if r/l is less than 2.2, but this has not been experienced with tightly wound coils on a ceramic form with r/l as small as ⅓, the temperature coefficient being positive and less than 1×10^{-6}. This may be explained by the fact that the wire is tightly wound on the form and the radius is changed more rapidly than the assumed coefficient or because the coefficient of the ceramics used were smaller than available data would indicate. In any event, experience has shown that an almost negligible temperature coefficient of inductance can be obtained with a coil of copper-plated iron-nickel alloy wound on a ceramic form. With zero-temperature coefficient condensers and a coil of this type, the tube and the socket, switches and other components will account for practically all residual frequency drift.

Tube Capacitance Variations

The input capacity of a tube (grid to all other electrodes) will increase by a substantial amount during warm-up. Measurements of typical tubes indicate that the increase is likely to be of the order of

0.02 to 0.04 $\mu\mu$f. At 40 Mc an oscillator which has a total of 40 $\mu\mu$f capacitance across the tuned circuit would have a frequency change of 5 kc per 0.01 $\mu\mu$f change in capacitance, hence, as much as 20-kc drift from this cause alone may be anticipated under the specified conditions. If the design and layout of the circuit is such as to permit a reasonable amount of lumped inductance, and thus high-tuned impedance, there are two methods of reducing the effect of changes in the tube input capacitance. The grid can be tapped down on the coil, or the L/C ratio can be reduced. The net effect is probably the same, except that switching problems may become more involved if the grid is tapped down. In any event the circuit must be arranged to minimize the effect of the tube-capacitance variation. This effect decreases as the real or apparent circuit capacitance is increased. Thus, in the example mentioned above, if the circuit capacitance were 80 $\mu\mu$f the change in frequency would be 2.5 kc per 0.01 $\mu\mu$f change in capacitance instead of 5 kc. At frequencies of the order of 40 Mc or higher, there is a limit to the amount of capacitance which can be used across the oscillator tank if suitable oscillator strength and reliability are to be maintained. It is for this reason that operation of the oscillator at a submultiple of the normal oscillator frequency may be advantageous. If the oscillator is operated at half frequency, something over twice as much capacitance may be used for the same tuned impedance. It is more than twice as much, because a larger proportion of the increased inductance is in the coil rather than the leads which tends to increase the circuit Q and, therefore, its tuned impedance. Thus, the frequency shift due to variation in tube capacitance can be reduced to less than half on a percentage basis by operating at half frequency. (The total frequency variation at half normal oscillator frequency will be less than ¼ the number of cycles which would be experienced at the normal oscillator frequency.)

This type of operation imposes a greater burden on the r-f circuits, since at the converter grid the receiver will be sensitive for signals at approximately 0.5 and 1.5 times signal frequency. These responses will be further off frequency than normal images. So if a receiver has a reasonable image ratio, it will have sufficient rejection for these spurious responses.

In a receiver operated in this manner with one r-f stage the rejection at these frequencies was found to be greater than 10,000 to 1.

Conversion gain does not necessarily suffer by virtue of operation of the oscillator at half frequency. Measurements indicate that if adequate injection voltage is applied at half frequency the conversion conductance of pentagrid converters is nearly equal to that obtained with normal operation. A further advantage of half-frequency injection is that oscillator frequency is less affected by tuning of the r-f circuits, and that less of the oscillator voltage appears on the converter signal grid by virtue of direct and space-charge couplings.

Sockets, Switches, etc.

Ceramic material rather than phenolic compounds should be used for insulation of "hot" points of the oscillator circuit. As a typical example, a 6-channel push-button switch of conventional construction using phenolic insulation had a change in capacity of 0.2 $\mu\mu$f for a 30°C temperature change. If it were connected across a 40 Mc oscillator with 40 $\mu\mu$f total capacitance a frequency shift of 100 kc would result. At this writing ceramic rotary switches are available, and their use is almost imperative for the oscillator band switching, even though phenolic sections may be used for r-f circuits, to reduce cost. Ceramic push-button switches are not available as yet. If a push-button switch or a tuning condenser having a substantial temperature coefficient must be used for station selection, some method of reducing the capacitance change effect is necessary. Where the band to be covered is a small increment of the oscillator frequency, it becomes possible to connect the switch or condenser across only a part of the coil. For example, connecting the switch across ⅓ of the coil results in 1/9 of the effect for a given capacitance change. The switch or condenser might alternatively be coupled to the oscillator coil through a mutual inductance to obtain the same effect.

Several types of variable air dielectric trimmer condensers have been tested. All possessed positive temperature coefficients which were small compared with that of other components, and, in general, the use of air-dielectric, ceramic-insulated trimmers is to be recommended for changing oscillator frequency. The constant portion of the oscillator-tuning condensers may be of the fixed zero-drift type, for reasons of economy and stability.

Compensation

Condensers with negative temperature coefficients may be used to compensate for changes in capacitance of components having positive temperature coefficients. It is possible to compensate in the same manner for inductance variations at only one frequency, although for a small tuning range and a small inductance variation with temperature, a close approximation to compensation may be realized. It should also be noted that it is impossible to compensate for a variable tuning condenser having a temperature coefficient except at one setting of the condenser.

However, best design requires that the uncompensated oscillator drift be at a minimum so that a minimum of compensation is required.

In the first place, it is virtually impossible to make all components change temperature at the same rate, hence, although an oscillator may reach equilibrium at the starting frequency after temperature stabilization has been reached, it probably has gone through some large frequency changes during the warm-up period. In a typical receiver under average conditions, temperature stabilization may require an hour or more, so that during this period the oscillator frequency is changing constantly. The tube temperature usually approaches temperature stabilization within about 15 minutes, with the balance of the components requiring the longer time. Rapid heating of the compensating condenser is sometimes accomplished by a heater unit. If the tube were the only cause of frequency drift, rapid heating of the compensator would be necessary to offset the rapid heating of the tube, but with other slower heating components also contributing to the drift, this method can at best be an approximation.

Better results might be obtained by using two compensators, one rapidly heated to offset tube effects, and the other heated by the surrounding air temperature to offset the drift of components which heat more slowly.

If a large amount of compensation is necessary, production variations may be expected to be much greater. For example, if an oscillator had an uncompensated drift of 60 kc as designed and the uncompensated drift and compensation were each to vary 10 per cent in opposite directions, the compensation would be off 12 kc. If, on the other hand, it is only necessary to compensate for a 6-kc drift the same production tolerances would yield a maximum discrepancy of 1.2 kc.

Mechanical Construction

Mechanical construction of the oscillator is of great importance in oscillator-frequency stability in addition to the well-recognized necessity for proper construction to minimize microphonic feedback effects. It is essential that the construction of the oscillator be such as to render unimportant any flexing, expansion, or other movement of the chassis so that oscillator frequency will not be influenced from this cause. It is further helpful to have leads bent to fit their position so that none of the leads or parts is under tension when secured in place. Experience has indicated that several cycles of alternate heating and cooling over a temperature range somewhat greater than is to be experienced in service is a further help if the uncompensated drift is very low. Obviously, proper mechanical construction of switches is necessary if good repeat accuracy is to be realized.

Humidity

Because there is no simple method of compensating for frequency changes caused by variations in humidity, it is essential that each component be immune to moisture effects insofar as it is possible to make it so. All components should have substantially zero porosity. This again indicates the desirability for ceramic insulation for humidity as well as temperature change reasons.

If suitable insulating materials are not used, the change in oscillator frequency due to humidity changes may well be considerably greater than those due to temperature variations.

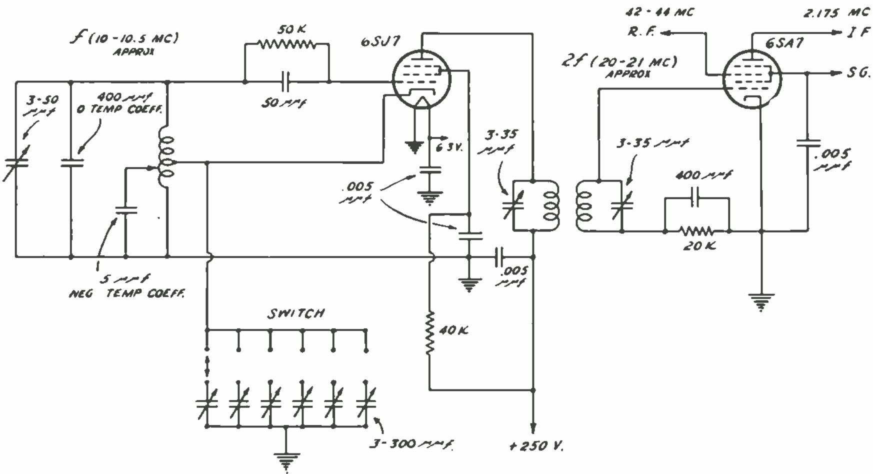

Fig. 1—Oscillator circuit.

Operating Parameters

It is important that the oscillator frequency be not greatly affected by changes in plate and heater potentials. A change in heater potential may have two effects; first, it may change the operating characteristics of the tube (g_m, r_p, etc.), and second, it may change the heater-cathode capacitance which will be of importance if that capacitance forms part of the oscillating circuit. In general, a reasonable change in heater voltage about the design value (as might be anticipated from line-voltage variations) is not likely to cause sufficient change in the operating characteristics to alter the oscillator frequency substantially, if the oscillator is otherwise stable and if the tube is not emission-limited.

The type of heater construction is apparently an important factor in cathode-heater capacitance variations, both during warm-up and heater-voltage change periods. No attempt has been made to evaluate this effect in terms of types of construction, but certain tubes seem to be very good in this respect, while others are definitely inferior. As might be expected, those types which have least change in capacitance during warm-up are least susceptible to changes in heater voltage, and the effect seems to be of secondary importance in such tubes. The exact value of cathode-heater capacitance variation has not been determined, since it is difficult to evaluate separately because of the fact that in the type of oscillator with the cathode above ground, the grid is also across the tuned circuit, and its capacitance variations exert greater influence on oscillator frequency. It seems safe to conclude that in a typical oscillator of this type where the cathode may be tapped about ⅓ of the way up the coil, the effect of grid-cathode

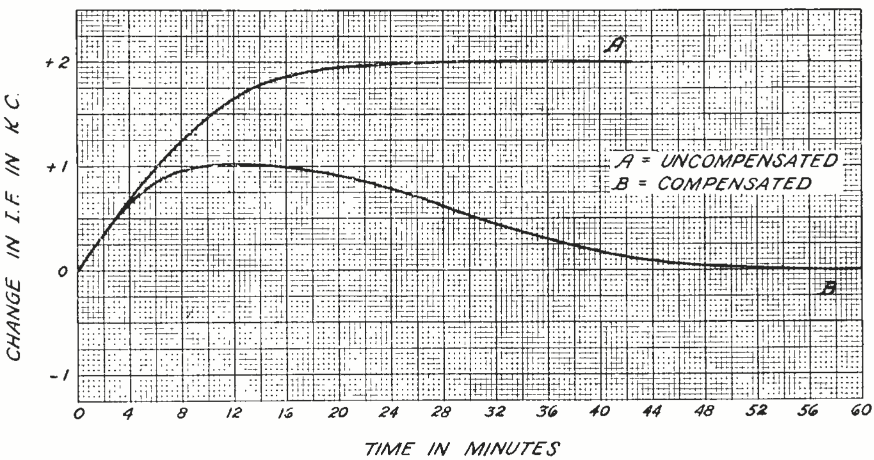

Fig. 2.

variation is substantially greater than cathode-heater variations with suitable tubes.

Changes in frequency with changes in plate voltage are principally due to resultant changes in effective grid-cathode electronic capacitance. Any expedient which reduces the influence of change in grid-cathode capacitance (as for example tapping the grid down on the coil or using large shunt capacitance) effects an improvement in frequency vs. plate-voltage stability as well as in frequency vs. temperature stability. Obviously, the plate voltage supplied to the oscillator may be regulated (as by a gas tube), but it is reasonable to conclude that any oscillator which uses a large tank capacitance to swamp out tube-capacitance variations will be sufficiently stable with respect to plate-voltage variations to eliminate any necessity for unusual precautions in this respect.

A measure of compensation for plate-voltage variations may be obtained by suitable selection of the position of the cathode tap if the oscillator uses a Hartley circuit. As the amount of inductance between grid and cathode is reduced, the change in frequency with change in plate voltage becomes less. If the oscillator circuit is of the type wherein the cathode-heater capacitance is part of the oscillating circuit, use of this expedient to reduce plate voltage change effects increases the influence of cathode-heater changes in capacitance.

An oscillator-converter system designed to have very small drift and used to evaluate many of the effects previously discussed was built in this laboratory. The circuit diagram is illustrated in Figure 1. This unit is not intended to typify the ultimate in oscillator design, nor is it intended as a standard of performance. Rather, it is included as an example of an oscillator incorporating the design considerations brought out herein.

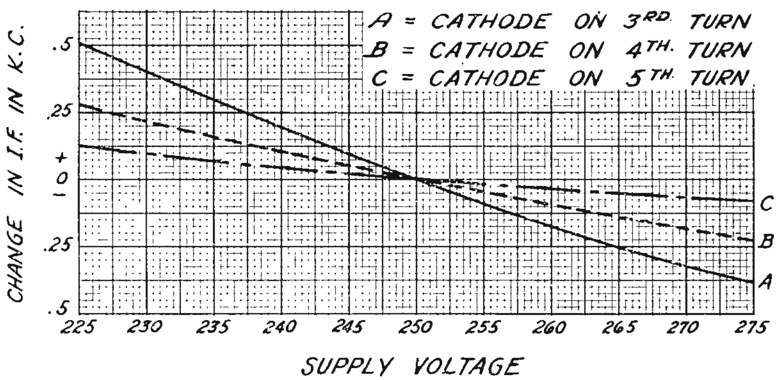

Fig. 3—Change in frequency with change in oscillator supply voltage.

The oscillator was used in a receiver to cover 42 to 44-Mc signal frequency with an intermediate frequency of 2.175 Mc and with the oscillator on the low-frequency side. The injection was at half-normal oscillator frequency, and the oscillator was at quarter frequency. The oscillator output circuit and the 6SA7 injection grid tuning circuit form a band-pass filter covering the desired frequency range (approximately 20 to 21 Mc). This circuit also eliminates the oscillator fundamental frequency from the mixer grid. The coil is tightly wound with copper-plated Nilvar wire on a ¾″ ceramic form. It has 7 turns and the winding length is approximately 1 inch. The cathode tap is 3 turns from the ground end of the coil. The push-button switch indicated in the diagram is of conventional construction and has phenolic insulation. The coil and condensers do not contribute appreciably to the observed drift, since both have substantially zero temperature coefficients. The tube and switch are responsible for the major portion of the drift and to approximately the same extent. The performance of the oscillator is indicated in Figure 2. Here the change in intermediate frequency is plotted against time. Zero time is taken as 45 seconds after the power is applied. It should be noted that because the oscillator is on the low-frequency side, a decrease in oscillator frequency results in an increase in i.f. Curve *A* is that of the oscillator without compensation. Curve *B* is with the amount of compensation necessary when the receiver reaches temperature stabilization.

Figure 3 indicates the change in frequency vs. supply voltage over the range from 225 to 275 volts for 3 different positions of the cathode tap. Curve *A* is with the cathode on the 3rd turn, Curve *B* 4th turn, and Curve *C* 5th turn from ground. As mentioned previously the coil has 7 turns. Again the measurement is at i.f.

TEST TECHNIQUE

It is worthy of note that stability of the order considered here is such as to require an extremely accurate frequency standard if reliable measurements are to be obtained. Ordinary laboratory crystal oscillators are often insufficiently stable for this work unless they are in temperature-control ovens. Standard-frequency transmissions from WWV (The Bureau of Standards Station in Washington, D. C.) on 5 Mc and 10 Mc were used in making these measurements.

As the work on a stable oscillator nears completion, the only accurate method of measuring frequency variations against time is in the chassis and cabinet in which it will be used so that heat will reach the components at the proper time and in proper amounts, but, for preliminary work where individual components are being tested a small hot-air blower of the hair-dryer type is invaluable. A large fan of 12 to 16-inch diameter will return a chassis to room temperature in a matter of a few minutes.

REACTANCE-TUBE FREQUENCY MODULATORS*

By

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Summary—*An improved reactance-tube modulator is described which utilizes a push-pull circuit to balance out the carrier frequency instability due to power supply variations.*

Design considerations are given for the type of reactance-tube modulator which uses automatic frequency control to stabilize the mean frequency.

When a simple oscillator is frequency modulated by means of a reactance tube, the problem of frequency stability immediately manifests itself. The reason for this becomes apparent when it is realized that the reactance tube is a device which makes the frequency of oscillation dependent upon the element voltages supplied to the reactance tube. Normally this condition is produced so that frequency modulation may be accomplished by applying the modulating voltage to one of the reactance-tube elements. However, at the same time the frequency stability of the oscillator, which normally depended on the oscillator circuit itself, now becomes dependent on the voltages supplied to the reactance tube. This results in a rather poor frequency stability unless steps are taken in addition to the mere connection of a single reactance tube across the oscillator to produce frequency modulation.

An obvious method of improving the stability of this simple combination is by the use of a regulated power supply to supply the element voltages to the reactance tube. It is the writer's experience that such an expedient is practically a necessity where this simple combination is used.

Another method of eliminating the high susceptibility to power-supply variations is by means of the push-pull reactance tube circuit as shown in Figure 1. In this circuit two reactance tubes are used which have opposite reactance variations so that they must be modulated in push-pull to cause their reactive effects to aid. Any push-pull modulation or power-supply variation is then canceled out in the same manner that push-pull circuits cancel hum from the power supply of an audio amplifier.

Tubes A and B of Figure 1 are the reactance tubes and the 6J5 is the oscillator. The Type 6SA7 tubes were used merely for the convenience effected by the extra grid for applying modulation. The radio-

* Reprinted from *QST*, June, 1940.

frequency feedback which converts the tube into a reactance tube is fed to the control grid. Tube A is connected in the same type of reactance-tube circuit as is conventionally used for AFC in broadcast receivers. Reactance R_1 and the grid-to-cathode capacity of the tube, C_{gk}, form a phase shifter which feeds phase-shifted voltage from the plate to the grid of the tube. R_1 is made large compared to the reactance of C_{gk} so that the phase of the current is determined by the resistance and is in phase with the voltage applied from the plate

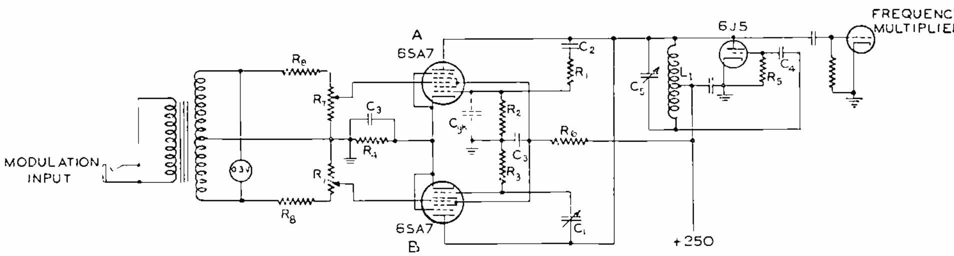

Fig. 1—Circuit of push-pull reactance-tube modulator. Tube A is fed phase-shifted voltage to its grid so that its plate circuit acts like a shunt inductance dependent on the gain of the tube. Tube B is fed by a different phase shifter which causes the plate circuit to act like a shunt capacity dependent upon the gain. Push-pull modulation of the two tubes thus produces additive frequency modulation from them both.

$R_1 = 50{,}000$ ohms	$R_5 = 100{,}000$ ohms
$R_2 = 0.5$ megohm	$R_6 = 9{,}000$ ohms
$R_3 = 1{,}000$ ohms	$R_7 = 100{,}000$ ohms, ganged
$R_4 = 175$ ohms	$R_8 =$ see text

$C_1 = 0\text{-}2\ \mu\mu$fd. Two-plate midget.

C_5 (for 5 megacycle oscillator frequency) $= 0\text{-}25\ \mu\mu$fd. Midget variable.

L_1 (for 5 megacycle oscillator frequency) $= 32$ turns No. 22 enameled, 1 inch diameter. Tap at 8 turns.

circuit. This current flows through the grid-to-cathode capacity and causes a voltage drop which lags the current (usually we speak of the current leading the voltage across a condenser, but in this case we are talking about the voltage drop with respect to the current so it is lagging). When this lagging voltage is amplified, the plate circuit is caused to appear as an inductance since its current lags the applied voltage.

Tube B uses a phase shifter which feeds a leading voltage to the grid and therefore produces a capacitive reactive effect instead of an inductive effect in the plate circuit. Instead of there being a low-reactance condenser in shunt to the grid, which is fed by a high resistance, there is a low resistance in shunt to the grid which is fed by a condenser of high reactance. The series condenser, C_1, is small enough to have a reactance which is high in comparison to the resistance, R_3, so that the current through the phase shifter is determined by the condenser and is therefore a leading current. This leading current also flows through the resistor in the grid of the reactance tube B so that a voltage drop is caused which is in phase with the current and is therefore leading. When this leading grid voltage is amplified by the tube, a capacitive effect appears in the plate circuit since the current flowing in the plate circuit is caused to lead the applied voltage.

The effective susceptance of a reactance tube has a magnitude which is proportional to the gain of the tube. Thus an increase in the gain of the inductive tube A decreases the effective inductance which is in shunt to a portion of the oscillator tuned circuit. This causes the oscillation frequency to increase. Tube B, which acts like a capacity in shunt to a portion of the oscillator circuit, produces a decrease in the effective capacity when the gain of the tube is decreased. Hence the tendency is to increase the frequency for a change in gain which is in the opposite direction to that which caused the same direction of frequency change on the inductive tube. Consequently, the application of push-pull modulation causes additive frequency modulating effects by the two tubes.

It will be noted that when the Hartley oscillator circuit is used as shown in Figure 1, the reactance tubes are connected across only the plate portion of the oscillator circuit. This somewhat reduces the effectiveness of the reactance tubes as compared to an arrangement in which the reactive effect is connected across the whole tuned circuit as might be the case if an oscillator circuit were chosen in which one end of the circuit were grounded. However, in the oscillator circuits with one end of the tuned circuit grounded, the cathode is usually at a radio-frequency potential from ground. Such an arrangement invariably produces hum in the form of frequency modulation. The remedy is to choose a circuit with the cathode grounded as was done in this case. Another alternative is to raise the heater to the same radio-frequency potential as the cathode by means of choke coils.

This push-pull circuit has the important advantage that it may be adjusted so as to neutralize all frequency instability due to power-supply variations. That is, if the oscillator taken alone happens to have reaction between the frequency and power supply, this reaction may be neutralized by a proper adjustment of the reactance tubes. In such circumstances the reactance tubes are slightly off-balanced to produce a residual push-push reactance characteristic which is equal and opposite to the reactance characteristic inherent to the oscillator. The adjustment of this balance is made by means of condenser C_1. For the normally balanced condition of the reactance tubes, the ratio of the reactance of C_1 to the resistance of R_3 would be equal to the ratio of the resistance R_1 to the grid-to-cathode capacity of tube A. In practice, C_1 may be adjusted by observing the beat note of the carrier and adjusting C_1 so that a variation of the power-supply voltage (obtained by cutting in a series resistance or in some similar manner) does not vary the frequency. For each setting of C_1, a resetting of the main-oscillator tuning is required in order to compensate for the reaction on the oscillator frequency effected by C_1.

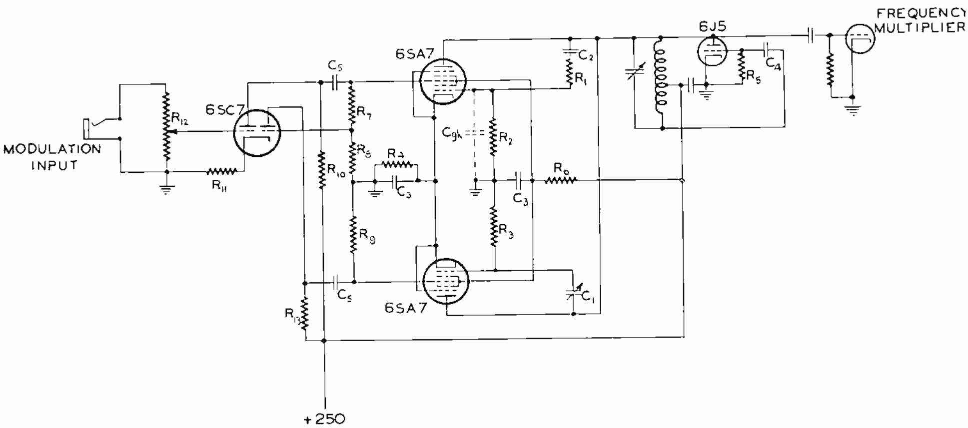

Fig. 2—Suggested push-pull reactance-tube modulator for a 14-megacycle oscillator frequency.

$R_1 = 50{,}000$ ohms
$R_2 = 0.5$ megohms
$R_3 = 300$ ohms
$R_4 = 175$ ohms
$R_5 = 100{,}000$ ohms
$R_6 = 9{,}000$ ohms
$R_7 = 475{,}000$ ohms
$R_8 = 16{,}000$ ohms
$R_9 = 500{,}000$ ohms
$R_{10} = R_{13} = 250{,}000$ ohms
$R_{11} = 4{,}000$ ohms
$R_{12} = 1$ megohm
$C_1 = 0\text{-}2$ $\mu\mu$fd two-plate midget
$C_2 = 0.001$ μfd
$C_3 = 0.005$ μfd
$C_4 = 0.0001$ μfd
$C_5 = 0.01$ μfd

Another method of balancing the reactance tubes may be employed when it is only desired to balance their reactive effects without neutralizing the power-supply reaction which may be inherent to the oscillator. In this method the frequency multiplied output of the modulator is observed on a frequency-modulation receiver and the modulator grids of the reactance tubes are temporarily tied together so that they are modulated in push-push. Modulation is then applied and C_1 is tuned for a minimum of the frequency-modulation output. Thus the reactive effects of the two tubes are arranged to oppose each other so that the adjustment for the minimum of frequency modulation indicates their equality. After the balance has been obtained the modulator grids are connected back in push-pull for normal operation.

In the circuit of Figure 1, the push-pull modulation transformer will depend upon the user's requirements. In the writer's usage a 600-ohm input was desired so that this transformer consisted of a line-to-line transformer with the secondary damped with 600 ohms.

This low-impedance secondary allowed the connection of the rather low-impedance rectifier-type voltmeter across the secondary without upsetting the impedance matching (the voltmeter had a resistance of 1000 ohms per volt and a full-scale reading of three volts making a total resistance of 3000 ohms). The potentiometers, R_7, are ganged. Resistors R_8 are chosen to produce full modulation for the maximum position of the potentiometers for a comfortable reading of the meter. Thus, they might be set at a value which gives a 25-kilocycle frequency deviation for the maximum setting of the potentiometer and a 2-volt reading on the meter. This arrangement allows the use of a readable deflection on the meter for normal modulation. Without such a potentiometer arrangement, a 25-kilocycle frequency deviation at 112 megacycles would be produced at about 0.05 volts on each grid (0.1 volts across the transformer secondary). This value is obviously too low to read on an ordinary meter.

The circuit of Figure 1 used an oscillator frequency of about 5.0 megacycles. While this low frequency makes possible a high degree of modulation and a sensitive modulator, it is felt that for amateur usage a higher master-oscillator frequency can be used. The circuit of Figure 2 is suggested for a master-oscillator frequency of 14 megacycles. The use of a phase inverter in place of the push-pull transformer is suggested since the added gain of the phase inverter will make possible the operation of the modulator directly from a crystal microphone.

A STABILIZED FREQUENCY MODULATOR

Where the utmost in frequency stability is desired, the circuit of Figure 3 may be used. This circuit utilizes automatic-frequency control to eliminate frequency instabilities regardless of their source. While a circuit employing the same principle has been described in *QST* before,[1] the circuit of Figure 3 is somewhat different and a few design considerations and a discussion will be given here which will aid the user of this type of frequency modulator. It will be seen that this circuit uses the simple oscillator V_2 and single reactance tube V_1. Normally this arrangement would be quite unstable if it were not for the automatic-frequency control system consisting of converter V_3, frequency-discriminator D and detectors V_4 and V_5. This automatic-frequency control system operates on the heterodyned output from the converter which is a relatively low frequency so that a small frequency change is a large percentage of the resulting heterodyned intermediate frequency. The crystal oscillator which is used as the

[1] D. E. Noble, "Frequency-Modulation Fundamentals," *QST*, August, 1939.

beating oscillator for the heterodyning process must have a stability which is at least equal to that desired for the final carrier frequency.

For maximum stability, the intermediate frequency at which discriminator D operates must be as low as possible. For broadcast frequencies where a high frequency deviation is used, an intermediate frequency in the vicinity of one or two megacycles is used, but for

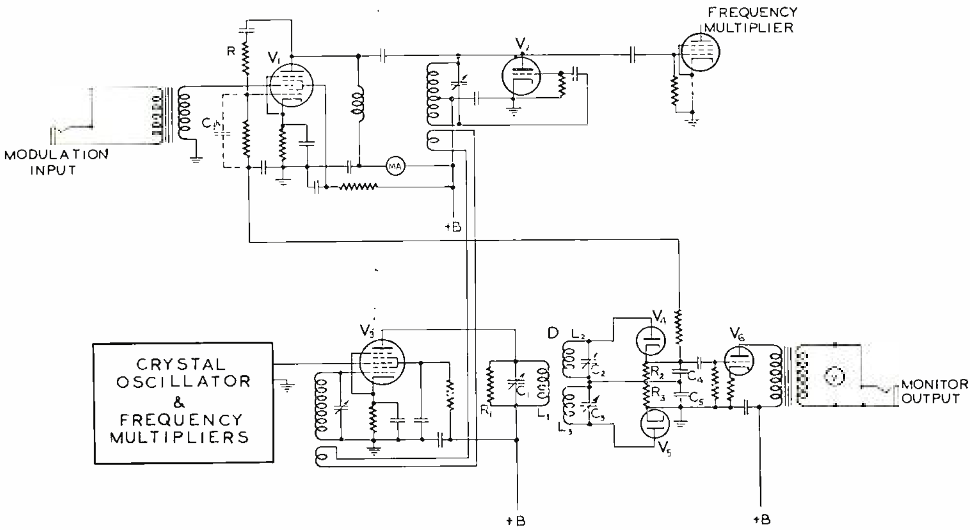

Fig. 3—Reactance-tube frequency modulator with automatic frequency control. The AFC circuits serve the dual function of maintaining frequency stability and providing a monitoring circuit.

I.F. = 460 kc
$C_1 = C_2 = C_3 =$ 0-75 $\mu\mu$fd midget variable
$L_1 = L_2 = L_3 =$ 2.5 M henry
$R_1 =$ 50,000 ohms
$R_2 = R_3 =$ 200,000 ohms
$C_4 = C_5 =$ 200 $\mu\mu$fd
Coupling between L_1 and $L_2 =$ 14%
Coupling between L_2 and $L_3 =$ 2.5%
Spacing between the primary and either secondary = ½″.
The input impedances of the diodes damp the two secondaries.

amateur usage, the conventional 450-kilocycle broadcast intermediate frequency has more appeal.

In Figure 3 it will be noticed that discriminator D is somewhat different than the usual Seeley type of a-f-c discriminator. It consists of a primary circuit which is tuned to the carrier frequency, and two secondaries which are off-tuned to opposite sides of the carrier frequency. The following are approximate empirically-determined rules which may be used in designing the tuned circuits:

$$Q_p = \frac{X_{C1}}{R_1} = \frac{X_{L1}}{R_1} = \frac{F_c}{3F_d}$$

where Q_p is the Q of the primary circuit, F_r is the discriminator mid-band frequency, and F_d is the maximum frequency deviation (one-half the total "swing") of the frequency-modulated wave applied to the discriminator. The Q of each of the secondary circuits is adjusted to be twice as great as that of the primary. This discriminator is merely an alternative to the Seeley discriminator and may be replaced by a Seeley discriminator with equal effectiveness. It may likewise be used as the discriminator in a frequency-modulation receiver using the same empirical design rules. The circuit is easy to align and is exceptionally linear.

The width of the discriminator must be adjusted to be able to receive the full frequency deviation of the frequency modulation present on the intermediate frequency applied to it. This means that if the converter is fed directly by the master oscillator, the discriminator may be narrower than it would be if the converter were fed by a harmonic (assuming that it is desired to generate wide-band modulation where the frequency deviation is several times the maximum modulation frequency). The narrower discriminator is capable of maintaining a greater stability, but if a high harmonic is fed to the converter, the frequency variations fed to the discriminator are multiplied so that the control is more sensitive. These two effects tend to offset one another so that it makes little difference wheher the converter is fed by a harmonic or by the fundamental of the master oscillator. Since the latter arrangement is simpler from the standpoint of the amount of multiplication required on the crystal oscillator, it is the logical arrangement to use. On the other hand, where the desired frequency deviation is on the order of an amount equal to the maximum modulation frequency, the discriminator will be of the minimum width (twice the audio-modulation band) regardless of which harmonic is fed to the converter. For this case maximum stability is obtained when the highest harmonic is fed to the converter.

The band width of the discriminator shown in Figure 3 is such that it is capable of handling the full frequency deviation of 20 to 25 kilocycles which has been proposed by Grammer and Goodman[2]. With a discriminator this wide, converter V_3 may be fed by the frequency-multiplied radiated wave, instead of by the master oscillator, if desired. If the converter is fed by the master-oscillator frequency as shown, somewhat greater frequency stability may be obtained by using a narrower discriminator.

The fact that the discriminator and detectors are available in the

[2] George Grammer and Byron Goodman, "Wide-Band Frequency Modulation in Amateur Communication," *QST*, January 1940.

circuit of Figure 3 affords a convenient method of monitoring the quality of modulation and the modulation level. By coupling audio amplifier V_6 to the detected output of the discriminator, the detected frequency modulation may be amplified for monitoring and may be fed to a meter which may be calibrated in frequency deviation.

The meter in the plate circuit of the reactance tube V_1 serves the purpose of indicating how much automatic frequency control is being used to hold the carrier on frequency. This meter has a normal reading which occurs when the carrier is in proper tune and a deviation above or below that reading indicates that a drift or change of some kind has caused the control to operate and bring the frequency back to as near normal as the degree of control allows. Normally the operator would only correct the tuning for large deviations of this meter since the frequency would be quite close in spite of a deflection from the normal.

For improved efficiency of the control circuit, an amplifier stage may be interposed between the converter and the discriminator so that the level fed to the discriminator is as high as possible. This amplifier should have a band pass which is capable of passing the full frequency swing of the modulation present in the intermediate frequency.

ULTRA-HIGH-FREQUENCY PROPAGATION THROUGH WOODS AND UNDERBRUSH

BY

B. TREVOR

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Summary—*Measurements of the attenuation of field strength through 500 feet of woods and underbrush on a frequency of 500 Mc showed a loss of approximately 17 to 19 db in summer and 12 to 15 db in winter as compared with the propagation over level ground. No great difference was found between horizontal and vertical polarization.*

At 250 Mc the attenuation through the same section of woods in winter showed a 10 db loss with horizontal and a 14 db loss with vertical polarization.

Transmission of 500-Mc signals over low scrub pines compared with that over sand ground showed a reduction of signal due to vegetation which can be interpreted as showing reflection rather than absorption of the indirect ray from a level considerably above ground or near the top of the vegetation.

INTRODUCTION

SINCE the use of frequencies above three or four hundred megacycles is finding increased usefulness the question naturally arose concerning the effect of foliage on the propagation of these frequencies. We might ask two questions:

1. What is the attenuation introduced by woods and underbrush by transmission through such a mass?

2. What effect does the foliage have on the indirect ray reflected from ground when the direct ray is in the clear above the underbrush?

MEASUREMENTS

A few experiments have been made in an attempt to partially answer these questions. A square patch of woods 500 feet per side on level ground was found which allowed a small 500-megacycle oscillator to be set up at one corner of the woods. The radiating antenna was nearly six feet above ground. The receiver in a car was driven from the remote corner of the woods along its edge, passing the corner of the woods 500 feet from the transmitter, and emerging in the clear in order to observe the difference in signal intensity as propagated over flat ground as compared with propagation through the trees and undergrowth. The receiving antenna was about 7 feet above ground. Measurements were made in July with full foliage out and again in November with no foliage present. Figure 1 shows a photograph of this patch of woods under winter conditions.

The results of these tests showed no appreciable difference between vertically and horizontally polarized transmissions in summer and an

Fig. 1

attenuation of 17 to 19 db due to the trees as compared with transmission over plain ground.

In November somewhat lower attenuations were obtained since the foliage had dropped off. Using vertical antennas the attenuation was approximately 15 db and with horizontal antennas 12 db.

The growth of vegetation was sufficiently dense to obstruct the view of the transmitter even with no foliage present.

Also in November a similar test was made on a frequency of 250 Mc. In this case the attenuation was measured to be 14 db with vertical and 10 db with horizontal antennas. Summer measurements on this frequency have not been made.

It should be pointed out that the accuracy of these measurements is not very great due to the bad standing wave patterns in space which were observed on the far side of the woods. The values shown above represent the best average that could be obtained.

In July 500-Mc transmissions were observed over a 500-foot span of level ground and compared with transmissions of 500 feet over low scrub pines. The antennas were about 8½ feet above ground and the height of the undegrowth was approximately 5 or 6 feet. The ground at this location was nearly pure sand. With vertical antennas

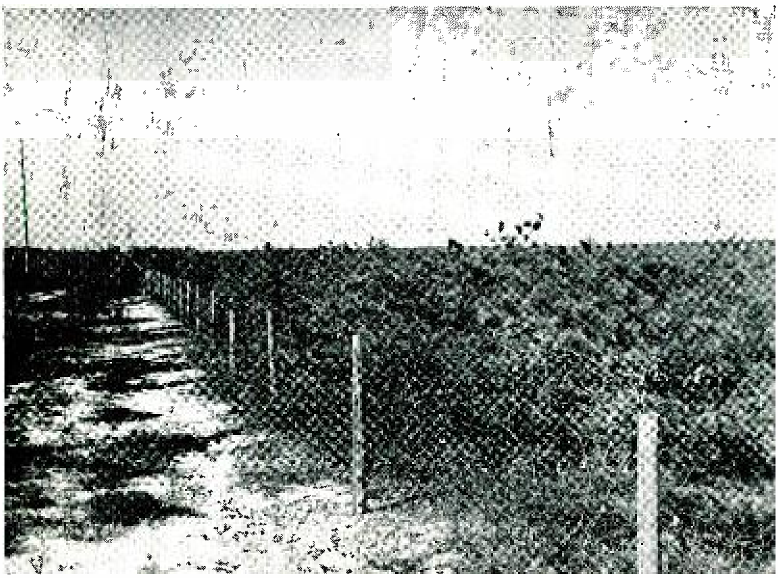

Fig. 2

transmission over the vegetation showed a loss of 8 db compared with transmission in the clear. With horizontal antennas the attenuation was 6 db. Figure 2 is a photograph showing a portion of the scrub pine area.

Conclusion

From these measurements we conclude that there is considerable attenuation of ultra-high-frequency waves in passing through woods and underbrush and that there is little difference between vertically and horizontally polarized waves. Also there is a noticeable difference in the attenuation between summer and winter conditions due to the absence of green foliage in winter. There is an indication that the horizontally polarized waves are attenuated somewhat less, particularly under winter conditions.

Since the signal was attenuated with transmission over low scrub pines we are led to conclude that the indirect ground ray was reflected from a level above that of the sand ground rather than absorbed by the vegetation. Under the conditions of measurement an absorption of the ground ray by the vegetation would have given an increase in signal.

REVERSED SPEECH*

By

E. W. KELLOGG

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Summary—*Recording sound and then reproducing it with the rotation of the record reversed is not only a means of amusement, but serves to throw some light on our habits of speech, particularly if one tries reversing the order of the original pronunciation, so that the reproduced sound may be actual words. Some of the writer's observations are given in the paper.*

ON FIRST thought it would appear that talking backward should be considered only as a sport or diversion, and it certainly has this feature. Those who have had at their disposal some ready means of recording speech and then playing the record either forward or backward have derived much amusement from the strange metamorphosis of the sounds. Experimenting with magnetic recording on tape or wire has inspired various people to try such reversals. Such recreation, however, is not without some scientific interest. Perhaps it does not bring out characteristics of our speech which might not be inferred from previously known facts and experimental data, but it brings home in a striking manner various characteristics of which for the most part we are unconscious, and it confirms a number of principles which might be predicted from purely theoretical considerations.

In order to draw conclusions, it is important that the recording and reproducing system have a minimum of distortion, otherwise many sounds are lost. The high fidelity recording and reproducing equipment for direct playback discs, which has recently been developed in the RCA laboratories makes available a medium for experiments of this kind which far surpasses what was to be had a few years ago.

The first impression one has on listening to reversed speech is that the talker is speaking a foreign language and talking very rapidly. The strange inflections suggest that it must be Chinese or at least some language with which we are very unfamiliar. Few of the individual word sounds are recognizable. Even though the record may be arranged to repeat a word or two over and over again, and we know what the word is, it is difficult to identify any sounds except a few of the prolonged vowels. This continues to be true even after consid-

* Reprinted from *The Journal of the Acoustical Society of America*, April, 1939.

erable experience. The difficulty is two-fold. In the first place when we talk, few of us make the sounds we think we do. Many of the sounds which ought to be present are slighted, or inexact substitutions are used. This is particularly true of short vowels. Just as a mother can identify her baby in spite of a very dirty face, so we recognize familiar words provided a reasonable fraction of thin characteristics are preserved. In the next place, we understand speech by recognizing whole words and not by a laborious synthesis of the word from a succession of sounds. A speed of speech which is entirely satisfactory when we can recognize the words, becomes many times too rapid when we are trying to identify the individual sounds in a meaningless sequence. We thus find that we get farther in learning about the reversibility of speech if we attempt to say the words backward and then listen to the result. There is then no danger of excessive speed, and if we are fortunate enough to succeed in making a series of sounds which when reversed bear a remote resemblance to the word intended, we find it much easier to determine whether the vowel and consonant sounds are present or are correct. If before making the speech, we analyze the sounds required with sufficient care, we will not carelessly make substitutions.

As an illustration of how we may give incorrect value to some sounds without appreciable effect on intelligibility, we may make use of the traditional difficulty in the spelling of Schenectady. Most of you know of the story of the man who planned to meet a business associate in Schenectady and after many attempts to think what was the correct spelling, told his secretary to wire his friend that he would meet him in Albany. Out of curiosity I once attempted to figure out how many ways Schenectady might be spelled without noticeably altering the pronunciation. My answer was 1125. This large number is reached partly by the various substitutions that can be made among the consonants such as "*ch*", "*k*" and "*ck*", all of which call for the same sound. There are also numerous accepted spellings for the identical sound, as in the case of the last syllable, but for the vowels in the syllables "*Schen*" and "*tad*", any of the five vowels (with their short sounds) might be substituted and the difference scarcely noticed, although each calls for a different sound.

Once we have acquired the technique of recording in such a way that we can actually recognize the reversed sounds, we are in a position to observe a number of factors. The first question which one might ask is whether a given sustained sound is to be expected to sound the same when reversed. Obviously the wave shape is not the same. For example, a saw tooth wave with slow rise and quick fall of pressure would, upon reversal, became a wave with quick rise and slow fall. The new wave would have the same component frequencies, but in different phase relation. There has been enough work done on the effect of phrase changes to warrant the prediction that this would not alter the character of the vowel sounds. The experience with reversed reproduction entirely confirms this. All sustained sounds seem to be produced without alteration.

We are on less certain ground in predicting what would happen to the explosive consonants such as *p, b, t, d, k* and *g*, but here again the evidence is that an initial and final consonant are the same, although with some qualifications. We can at least state that a given consonant does not become another when transferred from the beginning to the end of a syllable. It may, however, have a different value, or the same series of sound waves may produce much more impression in one position than another. There are obviously some conditions in the production of the sound, which cannot be very successfully reversed. For example, before the letter "*t*", a considerable pressure is built up behind the tongue and this is suddenly released. A similar rush of air just before closing off the passage in making a final "*t*" is probably nowhere near so violent. Putting the matter differently, we may start and end with the tongue against the roof of the mouth, but at that time we are not making any noise. There is no assurance that in the intermediate positions (when we *are* making a noise), the conditions are duplicated. When we come to several adjacent consonants, the reversal of the sequence often seems very awkward and we have no assurance that we are doing it correctly. For example, try pronouncing "world" backward, "dlrw". In this case, I have omitted the "*o*" in the reversed word. "*r*" has the quality of a vowel as well as a consonant and in pronouncing "world", the "*w*" and "*r*" sounds are separated by little if any vowel. "*w*" is a vowel sound which we often treat as a consonant, but it does not differ appreciably from "*oo*". In spite of the difficulties of forming sounds in reverse order, the results of the efforts to do so seem to be successful, and I do not believe that there are any exceptions to the general proposition that any of our words may be pronounced backward, and on reversal will produce the desired word, correctly if not altogether naturally.

One of the first things which is forced on one's attention is that a number of our vowels which we represent by a single letter and usually think of as a single sound are very distinctly a sequence of two sounds. For example, "*ī*" must be reversed to "*ē-ă*"; "*j*" as usually pronounced consists in "*d*" followed by a soft "*j*", and "*ch*" must be reversed to "*sh-t*". Overlooking any of these details produces an unexpectedly striking fault in the reversed sound.

In general final consonants are much less conspicuous than initial consonants. The final consonants *p, t, k* are often lost, and their absence does not seriously impair our understanding of the spoken words. Sometimes when a person wishes to enunciate with special clarity, he may release his tongue after a final "*t*", thereby speaking a short extra syllable in which the "*t*" appears as an initial consonant. The same is often done for "*k*" and "*p*". We are so used to this that it does not sound particularly strange and we are merely conscious of the fact that the person is putting effort into his pronunciation. We never do the corresponding thing at the beginning of a word. I see no reason why the consonants *l, m, n, r* should be less conspicuous at the end than at the beginning, but for some reason or other we have the habit of greatly prolonging them when they are used as final consonants. The last syllable of a word is also usually prolonged more than we are aware. In order to make a word sound anything like natural when spoken backward and reversed, it is necessary to drag out the initial sound and drawl the starting syllable to an unbelievable degree and correspondingly to shorten the final sounds. Otherwise the word, on reversed reproduction, shows the very characteristic which we should have put into it when speaking it backward.

The letter "*h*" is very elusive, and experience with attempting to talk backward makes one feel less justified in criticising the "cockney". We finish a word with a vowel, with no consciousness of having followed it by aspiration, and it comes back with a distinct "*h*" preceding it. This is merely another sample of the proposition already stated, that final sounds are very inconspicuous and this is more true of the "*h*" than of any other. In our written language we follow some vowels by "*h*" to indicate a certain vowel sound, such as "*ah*" or "*oh*", the actual "*h*" sound being either entirely absent or of no importance.

Inflection and accent have much to do with making our language what it is and making it understood. Accent is not difficult to figure out and in all probability inflection could be conquered with sufficient effort and practice, especially if one makes use of the expedient of first recording words as spoken normally and listening to and imitating the reversed speech as reproduced mechanically. Without such practice, however, about the best which can be done is to record in a sing-song manner which will playback as a monotonous mouthing of words with occasional very unnatural inflections. Singing is somewhat more successful. Once the new tune has been mastered, the melody takes the place of inflection.

Although singing is by no means synonymous with music, one topic suggests the other. It is of interest to play various types of music backward. One's ability to recognize the instrument producing the tones is cut to about 20 per cent of normal. Reversed piano music has a curious ending for all notes, but to most ears it is essentially organ music, or perhaps accordion music. This and the sounds of other kinds of instrumental music when reversed, shows that the attack and starting transients are probably an even more important factor in our recognition of the instruments than the quality of the sustained tones themselves.

It is often possible to recognize tunes when hearing them backward for the first time. The preservation of the general tempo is, of course, some help, but the factor that usually makes it possible to recognize a tune is that there are many phrases which are essentially symmetrical, and this is common enough in many melodies to give a clue.

As an illustration of talking backwards, I am giving a few lines spelled phonetically to the best of my ability. If you wish to learn what the corresponding English is, start from the bottom right-hand end and pronounce the sounds as you come to them from right to left.

"Tiónoot rawf tsujd[1] négyĕ dliósht ye eem kiém dnă."

"Tiólf yŏth[2] ni miŏt oh drawkáb[3] nrŭt drawkáb."

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[1] Soft "*j*", like French.
[2] Voiced "*th*"
[3] Pronounced as word "draw" but with the "*w*" emphasized.

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

By

B. J. THOMPSON, D. O. NORTH AND W. A. HARRIS

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

PART II—DIODES AND NEGATIVE-GRID TRIODES

BY DWIGHT O. NORTH

(*Continued from April, 1940*, RCA REVIEW)

EXPERIMENT

APPARATUS AND METHODS

ALL measurements were made at a frequency of about one megacycle. This is low enough to avoid transit-time complications and high enough to escape the "flicker effect" which contaminates shot-effect studies at audio frequencies. In design and construction of the special four-stage tuned amplifier,[1] extraordinary care was given to shielding, so as to minimize errors attributable to regeneration. The outfit was housed in a tight copper box, the false bottom of which contained thoroughly filtered supply leads; the upper portion was partitioned so that each tuned circuit was isolated. The partitions were constructed so that the tubes, themselves surrounded by shields, projected obliquely through the walls. Feedback was thus reduced to coupling between grid and plate elements in the type 57 pentodes. And even this source of regeneration was minimized by coupling each output to its tuned circuit through a large inductance which presented to the plate a capacitive reactance, varying but little with frequency. Each compartment was made accessible by a tight-fitting sliding cover. In this way voltage gains of over 10^6 were obtained without appreciable regeneration, and controlled by grid-bias adjustment of a type 58 pentode in the third stage. A fifth stage was alternatively employed as a vacuum-tube detector for preliminary work such as alignment, or as an additional stage of amplification feeding a thermocouple for accurate work. In the latter instance, a power tube was used to drive the couple to avoid cutting off the peaks. For detection the thermocouple was

[1] The amplifier and most of its associated equipment were constructed by L. I. Potter of these laboratories. Its continuously reliable performance must be credited entirely to his careful workmanship.

preferred in order to avoid any question as to whether the detector was really "square-law". A frequency-calibrated signal generator with calibrated attenuator provided measured signal accurate to 0.1 microvolt for alignment and calibration of the amplifier. It also provided, in one procedure, a comparison signal for noise measurement. Finally, the whole outfit was operated in a doubly screened room equipped with filters for all power lines.

The tube under observation was housed, together with its tuned coupling circuit (Z), in the first compartment of the amplifier box. All components of this unit were mounted on a removable false bottom, so that changes in tube, circuit, or connections could be made rapidly and conveniently. A small hole in the lid of this and each amplifier compartment permitted insertion of a condenser-tuning rod to allow adjustments in tuning without upsetting the shielding. The pass-band of the amplifier was about 10 kilocycles, and the "Q" of the coupling impedance Z was lowered by adding shunt resistance until it could be assumed

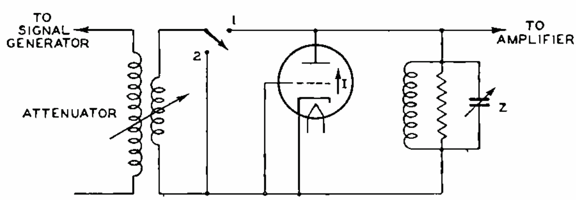

Fig. 8—Schematic shot-effect test circuit using signal generator.

that the over-all frequency-response curve was defined by the amplifier alone. This procedure simplified computation tremendously, since the pass-band could always be considered fixed, independent of whatever variations in the coupling circuit might occur in the course of measurement. The impedance Z was measured *in situ* by the familiar dynatron method, and fixed throughout the work at 45.5×10^3 ohms.

The early measurements were conducted in a manner explained schematically by Figure 8, which depicts only the essential radio-frequency connections. The signal-generator frequency was centered on the amplifier pass-band. Then, with the switch in positon 2, and the tube operating, the amplifier gain was set at a point which gave a suitable reading on the output ammeter. This reading was a measure of the sum of tube noise and thermal agitation in Z. Then the switch was thrown to position 1, and that signal (V) found which brought the ammeter reading back to its original value. (Inasmuch as the output impedance of the attenuator was negligible, it was immaterial whether or not the tube remained on.) The signal voltage was then a direct measure of the noise voltages integrated over the pass-band (Δf) of the amplifier. A measure of the noise per unit band width therefore required a determination of $\triangle f$, and this was accomplished in the usual way by running through a response curve. Thus, if V is the input voltage at frequency f which produces a fixed output meter reading, then

$$\triangle f = V^2_{\text{min.}} \int \frac{df}{V^2}.$$

Making a correction for thermal agitation in Z, one would finally evaluate Γ^2, for example, from

$$\Gamma_{\text{exp.}}{}^2 = \frac{\dfrac{V^2}{\triangle f}\left(\dfrac{1}{Z}+\dfrac{1}{r_p}\right)-\dfrac{4kT_o}{Z}}{2eI} \qquad (57)$$

In this expression r_p is the output resistance of the tube under test, in parallel with the circuit impedance Z, and I is the current, the fluctuations in which are under study. The second term in the numerator

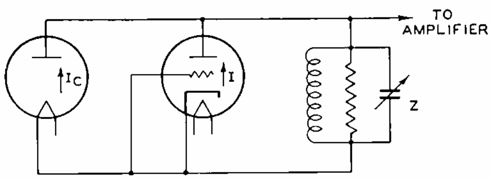

Fig. 9—Schematic shot-effect test circuit using comparator diode.

is the correction for thermal agitation in Z, and is almost always nearly negligible. No correction for background noise in the amplifier proper is in order since it affects both meter readings equally.

Straightforward variations in the circuit of Figure 8 permitted measurements of noise in not only the plate circuit, but grid and cathode circuits also. Yet for the latter, and for work with diodes, the method was not wholly satisfactory, for when the output resistance of the tube was small, the equivalent signal voltage V diminished in consequence, until inaccuracies in reading V led to adoption of a different procedure.

The second method replaced the signal generator by a temperature-limited diode, whose fluctuations could be calculated directly $(\Gamma = 1)$ from a measure of its anode current. One example of its use is shown in Figure 9 which, again, emphasizes only the radio-frequency connections. With input shorted, the amplifier had its gain adjusted at a point which gave an output reading R_1, a measure of the background produced by the amplifier proper. With the tube under test operating as desired, and with the comparator diode in the circuit but not operating (cold), a second reading R_2 was a measure of the sum of the shot effect under study plus amplifier noise. From a calibration curve of the output circuit, there was rapidly located a reading R_3 which would represent double the mean-square input to the amplifier. The filament of the comparator diode was then brought to that temperature at which the temperature-limited diode current I_c supplied additional noise sufficient to establish the reading R_3.

Just as in the first method, a correction for thermal agitation in Z will enter a determination of Γ from these data. And yet the working formula will be very simple because, first, all current fluctuations involved can legitimately be assumed to have a uniform frequency distribution; second, the input impedance remains unaltered when the comparator diode is in operation, the conductance of a temperature-limited diode being effectively nil. It turns out, then, that

$$\Gamma^2_{\text{exp.}} 2eI + \frac{4kT_o}{Z} = 2eI_c\,,$$

or, for the impedance used,

$$\Gamma^2_{\text{exp.}} = \frac{I_c - 1.14}{I}\,, \tag{58}$$

where I_c and I are expressed in microamperes.

The correction for thermal agitation is generally negligible. In comparison with the first method, this scheme has the disadvantage of necessitating a careful calibration of the output circuit, but this is not difficult and is permanent—until the thermocouple leaves in a puff. It has a distinct advantage in requiring no determination of the amplifier pass-band, a measurement which is always subject to correction as a result of aging, or other uncontrollable causes.

A third method, not employed in these studies, is worth mentioning inasmuch as it has proved very rapid and reliable whenever a measure of R_{eff} (see p. 471) was the chief objective. It is found directly by inserting resistance in the input of the tube until the mean-square input noise is doubled. Again taking into account a small correction for thermal agitation, it should be clear that this resistance is precisely R_{eff}. The method is probably unequalled as a swift, accurate scheme for rating tubes for signal-to-noise ratio. Its adaptation to studies of diodes has obvious drawbacks. For this reason, and because the present work is mostly concerned with testing theory, the other methods were preferred as more direct.

The comparator diode was essentially a copy of miniature diodes previously designed in this laboratory for short-wave work.[1] A nickel anode 6 millimeters long, and having a diameter of 15 mils, surrounded a 3-mil tungsten filament. With about 80 volts on the anode, currents of 2 or 3 milliamperes were unquestionably temperature-limited. For most measurements on radio-frequency receiving tubes, this proved to be ample current.

Although $\Gamma_{\text{exp.}}$ can be compared with $\Gamma_{\text{theor.}}$, it is somewhat simpler to find $\theta_{\text{exp.}}$ and compare this with $\theta_{\text{theor.}}$, which has been shown to be essentially a constant. At least it does not exhibit the wide variation that $\Gamma_{\text{theor.}}$ shows with change in operating conditions. The value of $\theta_{\text{exp.}}$ can be computed from, e.g., (58) and (43a) or (43b).

To make either test of theory, it is necessary to determine cathode temperature. Now this is one of the most awkward tasks if high accuracy is demanded. But if, as in the present studies, one is satisfied with experimental errors of 4 or 5 per cent, estimates of temperature accurate to well within these limits can be procured, at least for the more conventional structures which have a not-too-uneven temperature distribution. Filament temperatures were, therefore, computed from information found in "The Characteristics of Tungsten Filaments as Functions of Temperature".[2] A Leeds and Northrup optical pyrometer was used for early estimates of the temperature of sleeve-type cathodes. This was soon discarded; it was virtually impossible in many instances to get a direct line of sight. The temperature was subsequently computed from a measurement of heater power by means of the equation for temperature radiation. Extensive studies by E. G. Widell of these laboratories had shown that the fourth-power law was essentially valid for this type of cathode, and had permitted determination of emissivities. The working formula, developed from his studies and applicable to the present investigation, is

$$\frac{T}{1000} = \left[533 \frac{W}{dl} \right]^{1/4},$$

where W is heater power in watts, d is diameter in mils, and l is length in millimeters. This formula accurately applies only to the center "brightness" temperature of a type 6C6 cathode coated with a particular spray. Error due to cooling at the ends (50° to 100°) is largely offset by the fact that "brightness" temperature runs about 30° to 40° below true temperature. The formula is naturally somewhat altered

[1] L. S. Nergaard, "Electrical Measurements at Wave Lengths Less Than Two Meters", *Proc. I.R.E.*, Vol. 24, p. 1207, September, (1936).

[2] H. A. Jones and I. Langmuir, *Gen. Elec. Rev.*, June-August, 1927.

by changes in length, diameter, percentage of uncoated area at ends, type of base material, thickness and type of spray, and proximity of other bodies. For all sleeve-type cathodes discussed in the present paper such changes are minor, and the formula is believed to give their true average temperature with an error of less than 3 per cent.

Although the noise analysis is based upon a parallel-plane model, tests were conducted on cylindrical structures alone. Simplicity of con-

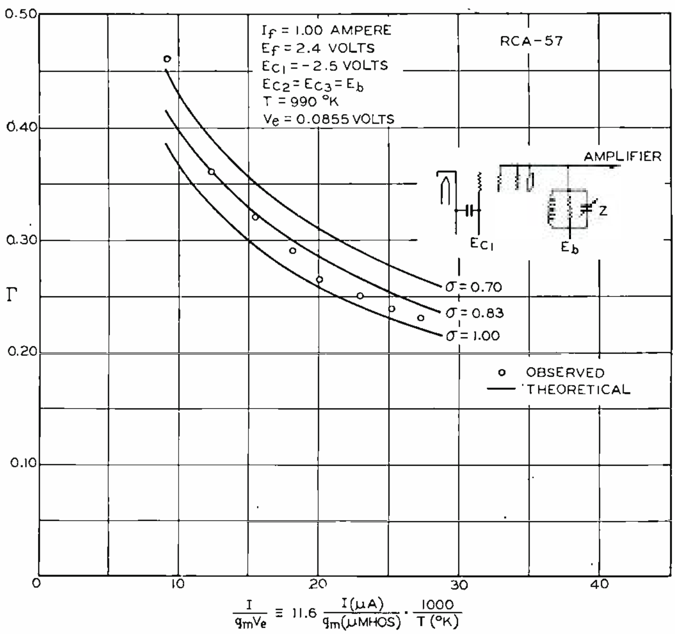

Fig. 10—Reduction of shot effect in cathode current of RCA-57 operated as a triode. The cathode current ranges from ½ to 4 milliamperes.

struction, minimum end effects, and ease of temperature determination all suggested the choice. Then, too, since commercial tubes are in the main cylindrical, it was hoped that the theory might be shown adequate for engineering purposes. Of course, many cylindrical tubes would be expected to behave essentially according to parallel-plane theory. The question becomes acute only in certain instances, e.g., filamentary cathodes, or especially large diameter ratios.

RESULTS

From a large group of observations on many commercial tube types, the following examples are chosen as typical. And although diodes would naturally be expected to receive first discussion, regular amplifying tubes will be given precedence, for reasons which will shortly be self-evident.

A precise test of theory against performance of amplifiers demands a determination of σ. It is unfortunate that there is no apparent way to measure this quantity directly; one is consequently forced to make estimates of σ from expressions such as (51), and tests of theory lose rigor thereby. However, it has been mentioned that σ will ordinarily lie between narrow limits, and estimates made for tubes of even very irregular design can be presumed accurate to well within 15 per cent.

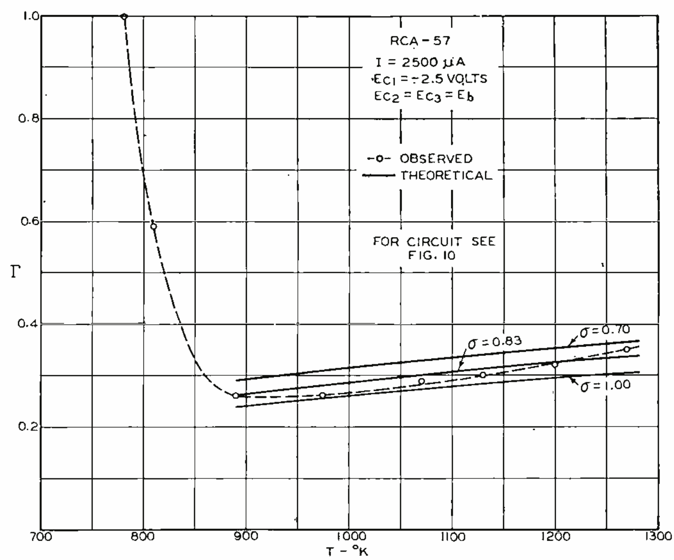

Fig. 11—Reduction of shot effect in cathode current of RCA-57 operated as a triode, for $I = 2.5$ ma and various cathode temperatures.

If, with this tolerance, there is seen to be good agreement with theory, there is much to recommend it, in view of the order-of-magnitude discrepancies typical of the past.

Tables IV and V and Figures 10, 11, and 12 all refer to the same tube, a regular RCA-57. This is a radio-frequency pentode amplifier which, for these measurements of cathode-current fluctuations, was operated in triode fashion, the screen, suppressor and plate being tied together as shown in Figure 10. Although the first two grids are elliptical, it is possible to make a good estimate of σ from (51). And because the μ is high (approximately 20 as a triode) and the transit-time ratio small, the estimated value ($\sigma \approx 0.83$) is not much below the ideal value of unity, and is virtually constant over the range of measurement.

TABLE IV

RCA-57 (Serial No. F-10)

$I_f = 1.00$ ampere $V_e = 0.0855$ volts
$E_f = 2.4$ volts $E_{c1} = -2.5$ volts
$T = 990°K$ $E_{c2} = E_{c3} = E_b$

E_b volts	I μa	g_m* μmhos	I_c μa	I/g_mV_e	$\Gamma_{exp.}$	$\Gamma_{theor.}$ $\sigma=0.70$	$\sigma=0.83$	$\sigma=1.00$	$\theta_{exp.}$† $\sigma=0.83$
75	500	640	108	9.14	0.46	0.45	0.415	0.38	0.81
85	1000	950	130	12.3	0.36	0.39	0.36	0.33	0.66
95	1500	1140	150	15.4	0.32	0.35	0.325	0.30	0.64
103	2000	1290	165	18.1	0.29	0.33	0.30	0.27	0.62
108	2500	1460	175	20.0	0.26	0.31	0.285	0.26	0.58
115	3000	1530	185	22.9	0.25	0.29	0.265	0.24	0.58
121	3500	1630	200	25.1	0.24	0.28	0.25	0.23	0.59
127	4000	1720	205	27.2	0.23	0.26	0.24	0.22	0.58

* As triode

$$\dagger\ \theta_{exp.} = \frac{\sigma}{2} \cdot \Gamma_{exp.}^2 \cdot \frac{I}{g_m V_e}, \text{ from (43b).}$$

TABLE V

RCA-57 (Serial No. 7-10)

$I = 2500\ \mu_a$
$E_{c1} = -2.5$ volts
$E_{c2} = E_{c3} = E_b$

T °K	E_b volts	g_m* $\mu mhos$	I_c μa	I/g_mV_c	$\Gamma_{exp.}$	$\Gamma_{theor.}$ $\sigma=0.70$	$\sigma=0.83$	$\sigma=1.00$	$\theta_{exp.}$ $\sigma=0.83$
780†	250	75	2500	430	1.0	1	1	1	‡
810	128	1050	880	34.1	0.59		‡		‡
890	114	1400	165	23.3	0.26	0.29	0.26	0.24	0.63
975	108	1460	175	20.4	0.26	0.31	0.28	0.255	0.59
1070	103	1500	205	18.1	0.29	0.325	0.30	0.27	0.62
1130	99	1530	230	16.8	0.30	0.34	0.31	0.28	0.64
1200	95	1540	250	15.7	0.32	0.35	0.325	0.295	0.65
1270	91	1590	300	14.4	0.35	0.365	0.335	0.305	0.72

* As triode.

† Lowest T which, for $E_b = 250$, gave $I = 2500\ \mu_a$. The tube was, therefore, essentially temperature-limited, as indicated by the low g_m.

‡ Not within the scope of the present analysis, since I/I_s is not sufficiently small. Following figures are evaluated under assumption that the analysis does apply, i.e., $I/I_s << 1$.

With fixed control-grid bias and constant cathode temperature, the currents shown in column 2 of Table IV were procured by adjusting anode voltage, column 1. The transconductance was measured, column 3, and the noise measurement is represented by the comparator-diode current in column 4. Column 6 was computed from (58), and column 5 from the data shown. If now a value is assigned to σ, and it is multiplied by values in the fifth column, $\Gamma_{\text{theor.}}$ can be read directly from Figure 6, since

$$\frac{\sigma I}{g_m V_e} = \frac{I}{g V_e}.$$

The next three columns were constructed in this manner and, together with $\Gamma_{\text{exp.}}$, are plotted in Figure 10. The curves for $\sigma = 1.00$ (ideal) and $\sigma = 0.70$ are included to indicate the spread occasioned by variations in σ, or the possible discrepancies arising from inaccuracies in making its estimate.

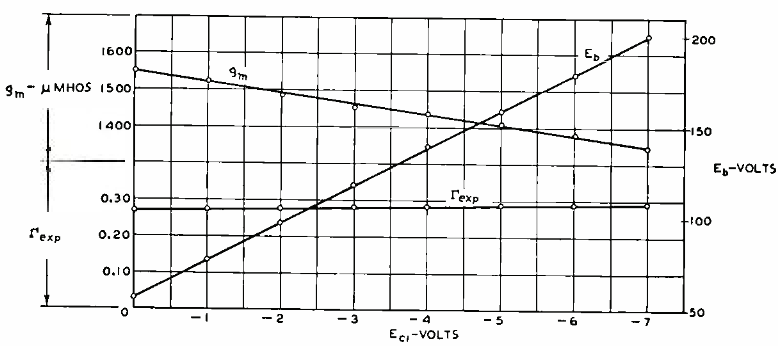

Fig. 12—Reduction of shot effect in cathode current of RCA-57 operated as a triode, for $I = 2.5$ ma, $T = 990°$K, and various control-grid biases.

Since I_s for this cathode temperature is well over 150 milliamperes, the current I could have been carried higher but was kept low on principle to avoid any doubt as to the observance of the stipulation, $I/I_s << 1$. The temperature, 990°K, is normal for this tube. Many similar runs of this kind at neighboring temperatures showed equally good agreement with theory.

In order to illustrate the course of Γ when T is varied, and also to show the transition from temperature-limited to fully space-charge-limited currents, the data of Table V and Figure 11 are presented as a typical example selected from many similar groups of data. The circuit was unchanged. But in this instance, with a fixed I (maintained by adjustment of E_b), the cathode temperature was varied, column 1. For the lowest temperature, the tube was virtually temperature limited.[1] For the two lowest temperatures no comparison with the theory of this paper is permissible, in view of the high value of I/I_s. The remaining data agree very well with theory, even exhibiting the predicted upward trend of Γ with increasing T. It is evident that, for this current, the tube should be operated with a cathode temperature of about 950°K, but that there is only a minor departure from the optimum Γ if T is permitted to deviate 100° to either side. The undesirability of too low a cathode temperature is impressively patent in the sharp rise to unity which characterizes the extreme left portion of this and all other Γ-curves of similar construction.

The plot is carried to only 1300°K for the simple reason that at higher temperatures measured noise exceeds the theoretical prediction by steadily increasing amounts. In itself this is hardly an excuse for discarding the phenomenon as non-pertinent, particularly after it is remarked that the behavior is characteristic of *all cathodes operated at temperatures several hundred degrees above normal;* this is true of tungsten, thoriated-tungsten, and coated cathodes alike. It has often been conjectured that the anomaly can be ascribed to emission of positive ions which, while pursuing a leisurely and long-lived course through the virtual cathode, control the destiny of multitudes of electrons. Evidence supports the view: (a) large-scale impulses appear simultaneously in the output meter; (b) the noise still approaches that predicted for true shot effect when the tube is made temperature-limited, i.e., when there is no virtual cathode upon which ions may react; (c) the change in excess noise with variation of temperature, and with time at a fixed temperature, is, in a qualitative manner, precisely that which would be expected on this hypothesis. Since the ion currents involved are much too small to detect by ordinary means, and since a quantitative analysis of the interaction of such ions with space-charge is much too laborious to promise immediately useful results, the theory remains in a semi-speculative state. The supporting evidence nevertheless appears to the author to justify exclusion of such data from tests of the theory of this paper. In Figure 11 it may be that part of the rise in $\Gamma_{\text{exp.}}$ with increasing temperature can, even here, be ascribed to this source.

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[1] The fact that g_m is not precisely zero does not contradict this assertion. A phenomenon of this nature, probably akin to Schottky effect is characteristic of coated cathodes. In contrast with the behavior of metallic emitters, coated cathodes exhibit no decisive saturation voltage, but only a gradual change in slope of the I vs. V curve. Measurements of I_s, which would perfect the picture presented by Tables IV and V, are for this reason doubtful and deliberately omitted.

Table V deserves final comment. The reader may already have wondered why, with I fixed, E_b continues to decrease, g_m to increase, even after $I_s >> I$. This behavior is best seen in terms of the equivalent diode which, stripped to essentials only, consists of the space between the virtual cathode and the control grid. Roughly speaking, then,

$$I \propto (E_a - E_m)^{3/2} \cdot d_{am}^{-2}$$

$$g_m \propto I^{1/3} \cdot d_{am}^{-4/3}.$$

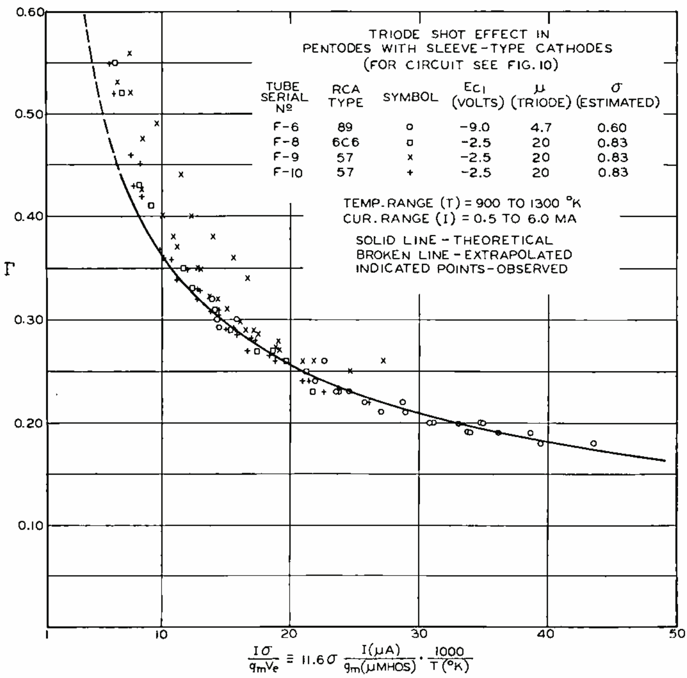

Fig. 13—Collected shot-effect data for a representative group of tubes with sleeve-type cathodes.

With constant I, an increase in T and I_s makes E_m more negative and decreases d_{am} (space between control grid and virtual cathode). The latter change accounts for an increase in transconductance. In the first expression both changes tend to increase I, and this necessitates a reduction in E_a, which, in turn, accounts for the decrease in E_b.[1] In order that the reader may have a correct appreciation of the magnitude and position of the virtual cathode, he may imagine a parallel-plane analogue to the equivalent diode of a type 57 under normal operating

[1] Unpublished work of H. R. Nelson of these laboratories suggests that the drop in potential may also be traced, in part, to thermo-electric emf's between cathode base metal and coating.

conditions. Assuming $I = 3$ milliamperes per square centimeter, $I/I_s = 1000$[1], $d = 0.05$ centimeters, and $T = 1000°$K, the effective anode potential will be about 0.6 volt ($E_a = 0.6$), and the virtual cathode will be approximately 0.6 volt deep ($E_m = -0.6$) and will be found at about 1/5 the distance from cathode to grid.

Figure 12 is included to demonstrate that, for fixed I, measured shot effect is independent of negative grid bias. This is in agreement with theory, for the quantity $I/g\,V_e$ which determines $\Gamma_{\text{theor.}}$ depends, not upon E_{c1} and E_b in themselves, but upon V_a which, if the current is to be held constant, must be invariant against alteration in electrode potentials. The course of E_b and of g_m is also shown. A small drop in the latter with increasing negative grid bias is normal, arising simply from non-uniformity of μ, i.e., effects due to grid support rods and grid ellipticity.

The agreement with theory which this tube exhibits is duplicated by tests on other tubes of the same general design. Figure 13 permits comparison of theory with approximately 100 observations of cathode-current fluctuations in four tubes. The 6C6 differs significantly from the 57 only in its heater resistance; the 89, on the other hand, is a low-μ power-amplifier pentode operated at well below its rated current, and is included to demonstrate the applicability of theory to tubes having a μ as low as even 5. Its grids are also elliptical, but a reasonably accurate estimate indicates that σ is about 0.6. This figure is lower than the 0.83 estimate given for the type 57 simply because the μ is much lower. The series of seven observations on F-9 showing the largest discrepancy on the left-hand side of the figure were all taken with $T = 1280°$K which is 70° higher than any other temperature for the data of this figure, and roughly 300° higher than the rated temperature for this tube. The discrepancy is, therefore, ascribed to incipient positive-ion emission. Although σ was assumed constant in constructing the plot, it should be remembered again that this is not exactly true. The common tendency of all points on the left-hand side to surmount their theoretical positions is possibly due to a lowering of σ for small currents as a result of the non-uniformities in μ mentioned above.

No such general accord with theory can be found in observations on these tubes operated as diodes. (In "diode" operation, all electrodes

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[1] That this is a reasonable choice of I_s, and representative of conventional oxide-coated cathodes, is supported by actual measurement; cf. B. J. Thompson, "High Efficiencies of Emission from Oxide-Coated Filaments", *Phys. Rev.*, 36, p. 1415; October, (1930). Although these measurements give too high a value of zero-field emission, on account of Schottky effect and surface inhomogeneities, they are probably high by a factor no greater than 10.

but the cathode were tied together; in other respects the circuit was unaltered.) Although the shot effect is still reduced by space-charge, it always exceeds the predicted amount. For very small currents, at the threshold of region (d) in Figure 7, the predicted shot effect is closely approached; with increasing current the mean-square noise exceeds the theoretical value by a steadily increasing factor. Figure 14 is representative of the relative behavior of "diode" as compared

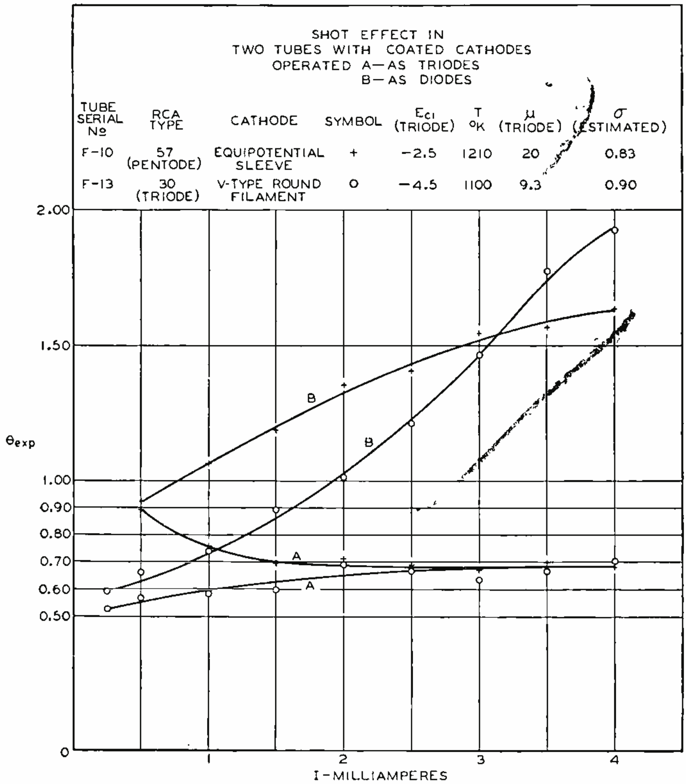

Fig. 14—Comparison of "diode" and "triode" shot effect in two regular tubes with coated cathodes.

with "triode" shot effect in amplifying tubes. Considering first the curves for F-10 (the same RCA-57 discussed above), note that the "triode" noise shows a $\theta_{exp.}$ which agrees closely with the theoretical value of approximately ⅔. The "diode" noise, on the other hand, shows a $\theta_{exp.}$ which steadily increases with current. It should be remarked that, although the same $\sigma = 0.83$ was assumed for plotting, the "diode" σ should properly have been chosen very slightly smaller on account of an increase in "h", (51); but this correction, while in the right direction, would by no means account for even a significant fraction of the excess noise. The second pair of curves belongs to a triode with a coated V-filament (RCA-30). The same extraordinarily high noise for large (but not abnormally high) currents is evident in the "diode" case. For both tubes the factor of disagreement is, at 4 milliamperes, not trivial, but rather an order of magnitude. The curve for "triode" noise of the RCA-30 is, incidentally, of interest in its own right, for it indicates that the parallel-plane formula applies to a structure even so remotely related as this one which, despite a flat grid and plate, has a cathode similar in no respect to the plane surface employed for analysis.

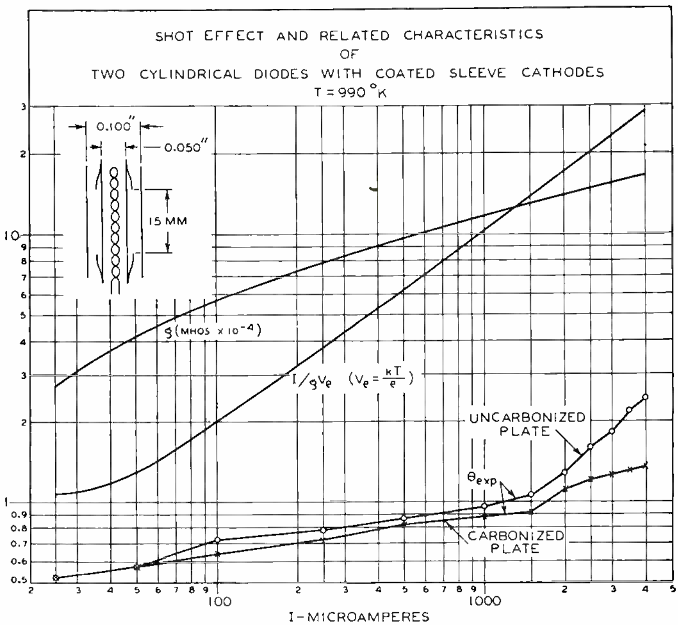

Fig. 15

Although one might conjecture that the anomalous "diode" noise is somehow obscurely connected with the presence of grids, such a notion is disqualified by measurements on actual diodes. Essentially the same results were observed in every instance. In an attempt to duplicate the theoretical model as closely as could be achieved with a cylindrical structure, two diodes were built as designed in Figure 15. A cylindrical coated cathode of 50 mils diameter was fitted at each end with short guard sleeves of thin nickel tubing, leaving a central exposed length of 15 millimeters. Both anodes were nickel of 100 mils diameter; one was hydrogen-fired at about 1100°C, the other carbonized. Both tubes gave ample emission at $T = 990°$K, and their current-voltage curves were so nearly alike that the plots of g and $I/g\,V_e$ in Figure 15 refer to either. The general course of these curves agrees very well with predictions based upon the steady-state equations for parallel-plane structures discussed above. No plot of voltage is shown since accurate absolute measurements were difficult in view of the small voltages required. The maximum voltage ($I = 4$ milliamperes) was estimated to be about 3.2 volts, corrected for contact potential. Calculation shows that currents below about 50 microamperes were collected in a retarding field. The curves for $I/g\, V_e$ and $\theta_{\text{exp.}}$ approach the theoretical limits 1 and 1/2 quite closely at the lowest current measured, 25 microamperes. There was, therefore, every reason to expect the shot effect to follow theory for larger currents. And yet $\theta_{\text{exp.}}$ is seen to rise steadily with current for both diodes just as it did for amplifying tubes in diode connections.

This strange behavior can, in the writer's opinion, be interpreted wholly in terms of a small amount of elastic reflection at collecting surfaces. Electrons returned to the vicinity of the virtual cathode as a result of either scattering or reflection will, of course, alter its potential and tend to decrease the steady-state flow of emission current across the barrier in a fashion strictly analogous to the process by which temperature-limited shot fluctuations are compensated and reduced by space-charge. The reflection hypothesis is chiefly concerned with those electrons which are elastically reflected or nearly so; an electron which loses more than a fraction of a volt on collision can not return sufficiently close to the virtual cathode to do damage at all comparable to the havoc promoted by those which actually pass through the virtual cathode a second time. It is impossible to handle this problem with quantitative rigor, but a crude notion of magnitude may be had. Consider a unit increase in the emission of electrons in a particular velocity class. Without space-charge compensation there would follow a unit increase in anode current. Our analysis has shown, however, that in the presence of a virtual cathode, there flows a compensating current $-(1-\gamma)$. The net increase in anode current is then $[1-(1-\gamma)] = \gamma$. But suppose, to exaggerate a little, that all of the original electrons are reflected back to the cathode to start out again and finally be captured on the second attempt. The net increase in anode current will be, in this event, something like

$$[1-(1-\gamma)-(1-\gamma)-(1-\gamma)] = -2+3\gamma\,.$$

If γ for this velocity class is, say 0.25, the net increase in anode current turns out to be -1.25 instead of $+0.25$. On a mean-square basis, the reflection of electrons has resulted in over-compensation to such an extent that the noise has been increased twenty-five fold. Now a brief calculation along these lines shows that if so little as 10 per cent of the *total* anode current is reflected as described, the *total* mean-square shot effect is increased by a factor of 3 or more in a diode comparable to those of Figure 15 and carrying three or four milliamperes current. It is logical that the factor should increase with I, for the quantity Γ *decreases* (see Figure 5), so that, supposing the coefficient of reflection to vary but little with an increase in anode potential, in the light of the example just discussed, the trend is obvious.

The precise amount by which noise will be increased by electron reflection depends upon too many factors to permit accurate prediction. Structure, nature of the electric field, electrode potentials, character of the reflecting surfaces, all are involved in a complex manner. Yet the chief features of the experimental difference between "diode" and "triode" noise are thus simply interpreted. For in the first place, a negative grid leaves smaller opportunity for anode-reflected electrons to return to the virtual cathode. But, more important, above a certain low critical potential a fairly high percentage of elastically reflected primaries drops to practically nothing. It is difficult to find much experimental evidence in point amongst the quantities of literature concerned with secondary emission, most of which is devoted to inelastic encounters and to emission of true secondaries, characterized by very low energies. But in one of Farnsworth's papers[1] the precise experimental information desired may be found. Of several conclusions which he draws from his measurements of the energy distribution of secondaries from *Cu, Fe, Ni,* and *Ag* the second is of present concern: "For primary voltages below a certain limiting value, which varies with the metal, most of the secondary electrons have energies approximately equal to the primary energy". Measurements of secondary energies for these four metals are much alike, and of the four *Ni* appears to reflect least. The table given here is constructed from his data for *Ni* and shows for each primary voltage V the per cent of the primary stream which appears in the secondary stream as electrons possessing within one-half volt of the primary energy.

V	6.2	10.4	18.6	33.5	50.0
%	12	11	9	6	5

It must be remembered that these data apply only to a specific surface put through a specific heat treatment, and cannot be considered

[1] H. E. Farnsworth, "Energy Distribution of Secondary Electrons from Copper, Iron, Nickel, and Silver", *Phys. Rev.*, Vol. 31, p. 405, March, (1928). In Compton and Langmuir's "Electrical Discharges in Gases", *Rev. Mod. Phys.*, Vol. 2, p. 171, April, (1930) will be found a brief factual background, but elastic and inelastic reflections are not clearly distinguished.

strictly representative of other samples. But the order of magnitude supports the hypothesis under discussion.

One might naturally expect to ascribe the difference between $\theta_{exp.}$ for the two tubes of Figure 15 to a reduction in elastic reflection as a result of carbonization. Subsequent experiments supported the notion. To separate out the possible alteration of reflection coefficient as a result of deposition of cathode material during activation, two diodes were constructed with dimensions similar to those of Figure 15, except that these had slidable nickel anodes. One anode was wholly carbonized; of the other, one-half was carbonized and the remainder was left "clean". The first showed substantially the same noise for

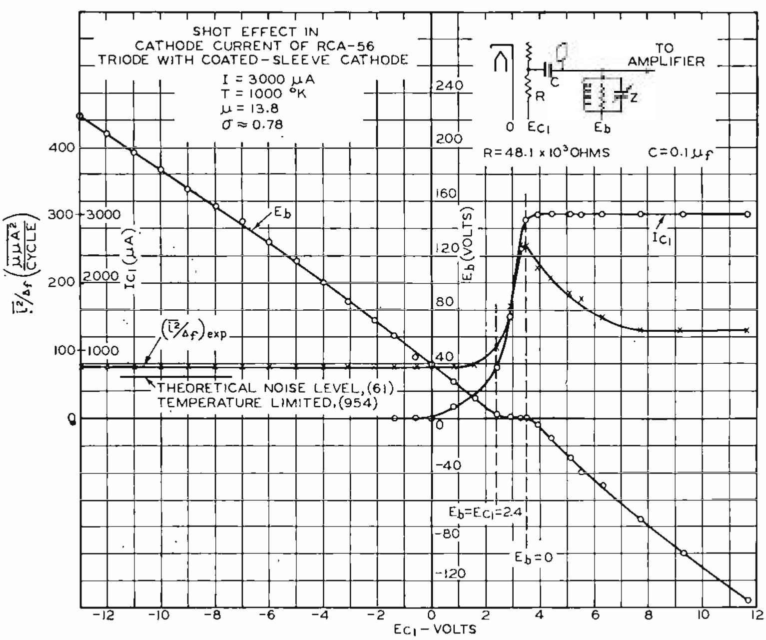

Fig. 16—Effect of electron reflection and scattering upon space-charge-reduced shot effect in a triode: constant cathode current.

either position of the anode, indicating that deposition of cathode material upon a carbonized anode during normal activation alters the reflection coefficient (of the kind in question) a negligible amount. The other, activated with the carbonized section exposed to the cathode, showed a slightly higher noise when the "clean" section was used. Even here, just as in Figure 15, the carbonized anodes showed noise far in excess of theoretical estimates; it is concluded that slow speed (2- or 3-volt) electrons are *reflected elastically* almost as profusely from a carbonized surface as from a clean one.

Further support for the hypothesis of noise augmented by reflected electrons is found in Figure 16. By operating a triode with grid positive and anode negative, one obtains an artificial electron reflection. *All* electrons that miss the grid on first passage are returned, and many of these may miss a second time and continue back to the cathode. The figure shows in panorama the entire course of shot effect in a constant cathode current for anode voltages running from high positive to high negative values. The construction of the RCA-56 differs little from that of the RCA-57 except that a carbonized anode replaces the 57's screen grid, and a smaller μ (13.8) brings σ down from an estimated 0.83 to 0.78. The circuit sketch shows how grid and anode could be given individual bias voltages and yet operated as a unitary electrode for radio frequencies so that, even though the cathode current was at times divided between the two, fluctuations in cathode current (I) were measured under all circumstances. Bias voltages were measured at the electrodes, were not corrected for contact potential, and for small values are therefore not precise. The trend of E_b vs. E_{c1} is normal; in particular the kink accompanying the approach of E_b to zero from the positive side is familiar and has long been correctly attributed to a sudden lowering of space potential due to the abrupt return of much of the anode-directed current to the vicinity of the virtual cathode. The noise, which for normal biases remains very constant and approximates the theoretical value, rises sharply to a peak at precisely the point at which it would be expected to do so. This maximum is considerably higher than the "diode" noise, at $E_b = E_{c1} = 2.4$ volts. The ultimate decline as more and more electrons are trapped at the grid on first passage, and the finally levelling off need little comment.

Finally, some observations on diodes with thoriated-tungsten cathodes should be mentioned. A 4-mil filament was used with a nickel anode of 100 mils diameter, either carbonized or not. At temperatures providing emission sufficient to remove suspicion from the ratio I/I_s, the behavior was much the same as that exhibited in Figure 15, including a decrease in noise with carbonization. Further, $\theta_{\text{exp.}}$ for these was, for the higher currents, always a little less than shown in Figure 15; with $I = 3.5$ milliamperes, $\theta_{\text{exp.}} = 1.95$ and 1.15 for these diodes, uncarbonized and carbonized, while $\theta_{\text{exp.}} = 2.19$ and 1.31 for their coated-cathode brothers. A reflected electron finds the virtual cathode less easily when the cathode has a diameter of 4 mils instead of 50 mils.

However unsatisfying it may be that the theory should agree but poorly with measurements on the simplest structures patterned closely after the model upon which the analysis was based, we conclude that there is generally excellent agreement between theory and observations of cathode-current shot effect in tubes operated with negative control grids. The formulas developed serve well as accurate engineering guides to construction of low-noise tubes for class A operation. And where experiment has shown the theory to be inapplicable, the hypothesis of elastic reflection appears adequate to account for discrepancies. The nature of cathode-current fluctuations, for small transit angles, is now thought to be well understood. The complexities which are added to the situation when the cathode current is divided and the noise measured in one portion only, for example in the plate current of a pentode, form the subject of the succeeding pages.

(To be continued)

OUR CONTRIBUTORS

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