RCA REVIEW

a technical journal

RADIO AND ELECTRONICS
RESEARCH • ENGINEERING

VOLUME XII MARCH 1951 NO. 1

RCA REVIEW

SUBSCRIPTIONS:

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Other Countries: One Year \$2.40, Two Years \$4.30, Three Years \$5.70

SINGLE COPIES:

United States: \$.75 each. Other Countries: \$.85 each

Published quarterly in March, June, September, and December by RCA Laboratories Division, Radio Corporation of America, Princeton, New Jersey

Entered as second class matter July 3, 1950, at the Post Office at Princeton, New Jersey, under the act of March 3, 1879

PRINTED IN U.S.A.

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Published quarterly by
RADIO CORPORATION OF AMERICA
RCA LABORATORIES DIVISION

in cooperation with

RCA VICTOR DIVISION
RADIOMARINE CORPORATION OF AMERICA
RCA INTERNATIONAL DIVISION
RCA COMMUNICATIONS, INC.
NATIONAL BROADCASTING COMPANY, INC.
RCA INSTITUTES, INC.

VOLUME XII — MARCH, 1951 — NUMBER 1

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RCA Review is regularly abstracted and indexed by ***Industrial Arts Index Science Abstracts*** (I.E.E.-Brit.), ***Engineering Index, Electronic Engineering Master Index, Abstracts and References (Wireless Engineer***-Brit. and ***Proc. I.R.E.)*** and ***Digest-Index Bulletin***.

RCA REVIEW

USE OF NEW LOW-NOISE TWIN TRIODE IN TELEVISION TUNERS*

By

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***Summary**—The sensitivity of television receivers can be substantially improved through the use of the 6BQ7, a new low-noise double triode, as a radio-frequency amplifier in "driven-grounded-grid" circuits devised specifically for this application. The merits of these circuits are discussed, with emphasis on the relationship between circuit performance and tube characteristics. Data are presented on noise figure, image rejection, gain, and standing-wave ratio for various frequencies in the very-high-frequency television bands. The attenuation of local-oscillator energy in the radio-frequency amplifier tube, an important factor in reducing total oscillator radiation, is greater with this tube and associated circuits than with comparable pentode circuits. The practical problems of applying the new circuits to a twelve-channel tuner are discussed. The use of the 6BQ7 in a low-noise intermediate-frequency preamplifier stage for ultra-high-frequency television receivers is also discussed, and pertinent data on noise figure, gain, and selectivity are provided.*

Introduction

SEVERAL years ago the writer investigated the performance of various receiving tubes in the radio-frequency positions of very-high-frequency television receivers.[1] The need for improvement in tuner performance evidenced then has resulted in the development of a new double triode, the 6BQ7, and of circuits for its use.

This development was based on an analysis of tuner requirements. The requirements for good tuner performance are determined by signal level, the type of antenna, the length of the transmission line, and the ambient interference levels. Therefore, if a television receiver is to work under a large variety of conditions, each of the following factors of tuner performance is highly significant: signal-to-noise ratio, selectivity and band-pass characteristics, voltage gain, amount of oscillator radiation, amount of antenna mismatch, and degree of cross-modulation in the radio-frequency tube, and mixer tube. None of the tubes measured in the previously mentioned investigation permit tuner operation that is adequate for all of the above factors. Some, as for

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* Decimal Classification: R583.5.

[1] R. M. Cohen, "Radio Frequency Performance of Some Receiving Tubes in Television Circuits", *RCA Review*, Vol. IX, No. 1, March, 1948.

example the 6J6 and 6J4 triodes, generate little noise but are unstable in neutralized circuits, or have objectionable antenna termination characteristics in grounded-grid operation. The pentodes, for example the 6AG5 and 6AU6, generate considerably more noise, but do not require neutralization and are more stable in tuned-input circuits. Thus improvement may be sought either with triode or pentode operation; a tube or tube and circuit combination which has the advantages of both is the desired objective.

Tuner Requirements and Tube Design

In considering the relationship of television tuner performance to tube design, the need for high sensitivity, i.e., low noise figure, dictates the use of a triode design having high transconductance, low input loading, low input and output capacitances, and low values of lead inductance.[2,3] Furthermore, for proper antenna termination the tube should have an input impedance that does not change with variation of the gain-control bias voltage which must be applied to the radio-frequency amplifier stage to avoid overloading with strong signals. To reduce cross-modulation in the radio-frequency amplifier tube, an extended cutoff characteristic is desirable. Unfortunately, this characteristic conflicts with the sharp-cutoff grid design desired for low input loading. The oscillator radiation attributable to the radio-frequency amplifier tube is a function of the capacitance from the radio-frequency amplifier output terminals to the antenna terminals, and of the circuit impedance at these terminals. The low-noise features of triodes have been recognized generally, but stability difficulties and other problems associated with the use of triode tubes in the conventional circuits have limited their extensive application in television tuners. Consequently, pentodes have been used in the radio-frequency stages in most receivers despite their higher noise. The development of the 6BQ7 and its associated circuits offers the possibility of a change in this situation.

The following discussion reviews various conventional triode circuits, outlines the features of the new circuits, and gives their advantages.

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[2] B. J. Thompson, D. O. North and W. A. Harris, "Fluctuations in Space-Charge-Limited Currents at Moderately High Frequencies", *RCA Review*, Vol. VI, No. 1, pp. 114-124, July, 1941.

[3] D. O. North and W. R. Ferris, "Fluctuations Induced in Vacuum Tube-Grids at High Frequencies", *Proc. I.R.E.*, Vol. 29, No. 2, pp. 49-50, February, 1941.

Grounded-Cathode and Grounded-Grid Circuits

Figures 1a and 1b show two popular radio-frequency amplifier circuits for triode tubes. The grounded-cathode circuit has the serious disadvantage of requiring a neutralization adjustment which is rather critical and unstable when a tuned input circuit is used. The grounded-grid circuit,[4] while it does not require neutralization, has a very low input impedance which varies inversely with transconductance. This variation makes it impossible to maintain correct antenna termination when gain-control voltage is applied to the radio-frequency stage. Also it is very difficult to produce the low-inductance input circuit required for good selectivity. It is necessary to provide gain control in the radio-frequency amplifier stage to avoid overloading the intermediate-frequency amplifier when strong signals are present. Variation in receiver input impedance with bias, experienced with grounded-grid

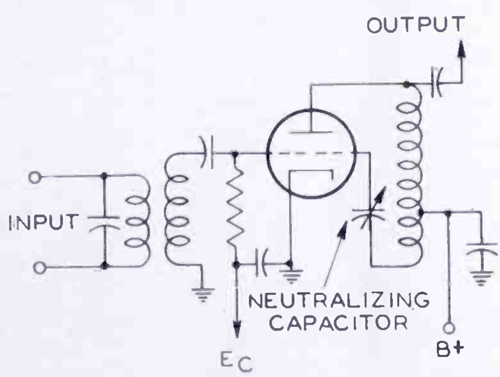

Fig. 1a—Grounded-cathode circuit.

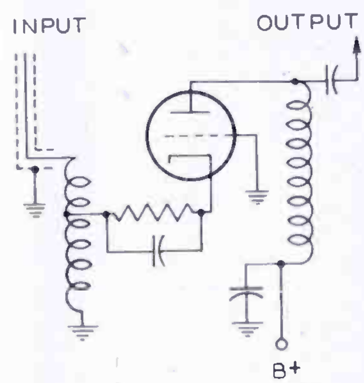

Fig. 1b—Grounded-grid circuit.

operation, causes improper antenna termination and resultant reflections which impair definition and may cause ghosts. These facts, plus the lack of a moderate-cost tube suitable for grounded-grid operation, may account for the rather limited use of the grounded-grid stage in the past.

Television boosters, however, do not require gain-control voltage since they are not used generally with strong input signals, and may, therefore, employ push-pull grounded-grid operation. The 6BQ7 has suitable characteristics for this application and data on push-pull grounded-grid operation will be presented.

Inverted-Amplifier Circuit

The circuits subsequently discussed in this paper are related to the basic inverted amplifier circuit, which is many years old. Figure 1c

[4] E. F. W. Alexanderson, U. S. Patent No. 1896534, filed May 13, 1927.

shows this circuit, a modification of Alexanderson's grounded-grid-amplifier circuit. This modification was described by C. E. Strong, used commercially and known in pre-war days as the inverted amplifier.[5] The circuit shown is an improvement of the earlier "Inverted Ultra Audion Amplifier"[6] of his associate Romander. Whereas Romander proposed eliminating all neutralization, Strong recognized the necessity for neutralization in both the driver stage and the grounded-grid amplifier, and discussed the various types of neutralization such as shunt inductance[7] and capacitance bridge methods. Strong worked with frequencies of about 20 megacycles in a transmitter, but recognized the utility of the amplifier for "higher frequencies as required for television and other purposes." He predicted its usefulness for low-power work at frequencies exceeding 300 megacycles.

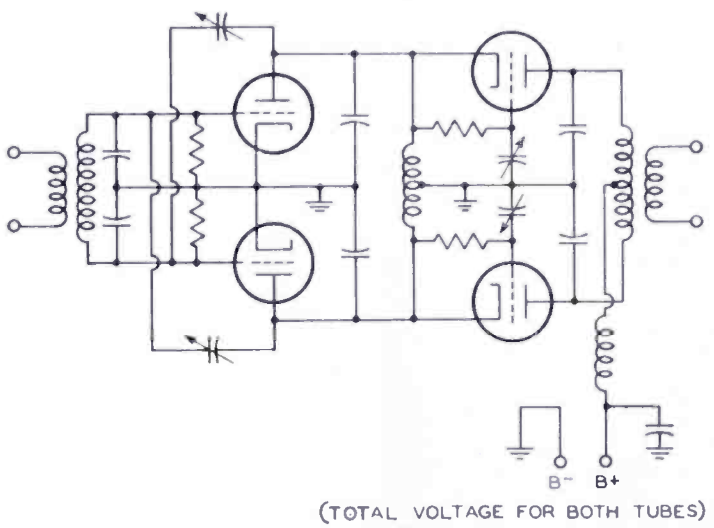

Fig. 1c — Inverted amplifier, push-pull grounded-cathode stages driving push-pull grounded-grid stages.

"Cascode" Circuits

Figure 2 is a "cascode" amplifier described and analyzed by Wallman and his associates and used in their article in a low-noise first intermediate-frequency stage.[8,9] This circuit arrangement combines the desirable features of a pentode, namely low output-to-input admittance and high input impedance, and the low noise quality of a triode. However, the circuit appears to have serious limitations when used in other than intermediate-frequency amplifiers or other single-

[5] C. E. Strong, "The Inverted Amplifier", *Electronics*, Vol. 13, No. 7, pp. 14-16, 55, July, 1940; and U. S. Patent 2241892, filed 1937.

[6] H. Romander, "The Inverted Ultra Audion Amplifier", *QST*, Vol. XVII, No. 9, p. 14, September, 1933.

[7] Nichols, U. S. Patent 1325879, December 23, 1919.

[8] F. V. Hunt and R. W. Hickman, "On Electronic Voltage Stabilizers", *Rev. Sci. Instr.*, Vol. 10, p. 16, January, 1939.

[9] H. Wallman, A. B. Macnee and C. P. Gadsden, "A Low-Noise Amplifier", *Proc. I.R.E.*, Vol. 36, No. 6, pp. 700-708, June, 1948.

frequency amplifiers, because neutralization of the input stage is required for optimum results. This neutralization is not extremely critical at any one frequency and can be accomplished with a tuning coil which is effectively in parallel with the grid-plate capacitance of the first unit. The neutralization coil also serves as a radio-frequency choke returning the cathode of the second unit to ground, thus eliminating the cathode choke otherwise required. This circuit, while well suited for intermediate-frequency amplifier use, is extremely difficult to apply to a multi-channel tuner because the neutralization is frequency-selective, and requires individual coil switching for each channel. Attempts to use this circuit without neutralization have been unsuccessful, except at the lower-frequency channels, because the degenerative feedback increases with frequency. The capacitance to ground from the plate of the input triode and from the cathode of the output triode, plus the distributed capacitance to ground of their connecting leads, also causes degeneration in the higher-frequency chan-

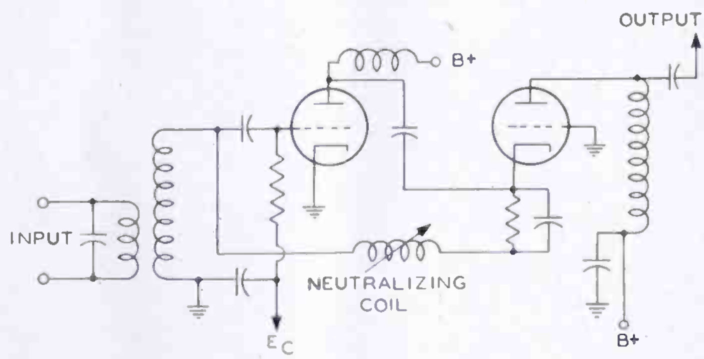

Fig. 2—Cascode circuit.

nels where the value of this capacitive reactance approaches the input impedance of the grounded-grid-section. This input impedance is approximately the reciprocal of the transconductance and is in the order of 200 ohms in a tube having a transconductance of 5000 micromhos. A distributed capacitance of only 2 micromicrofarads, because it has a reactance of only 400 ohms at 200 megacycles, appreciably reduces the input impedance of the grounded-grid unit. This effect reduces the gain, causes degeneration due to the capacitive phase angle, and allows the noise of the output unit to contribute to that produced by the input unit, impairing the noise figure.

Driven-Grounded-Grid Circuit

Figure 3 is one of the new circuits developed for the 6BQ7 which, for identification purposes, is called the driven-grounded-grid circuit,[5] although the term is also descriptive of the inverted amplifier and cascode circuits. Note that neutralization is accomplished by means of

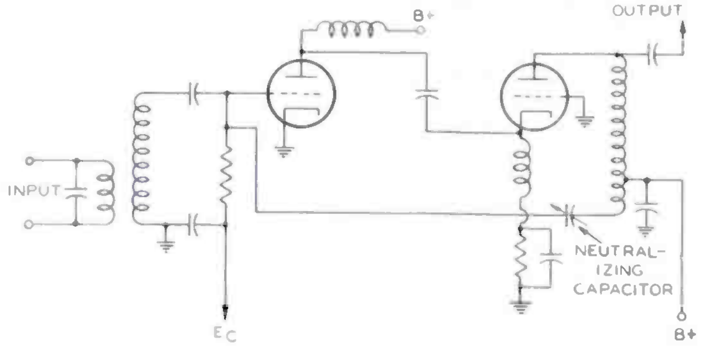

Fig. 3 — Driven-grounded-grid circuit.

a bridge circuit commonly employed with single triode amplifiers. This method of neutralization has the distinct advantage of being relatively independent of frequency, provided the connecting leads in series with the neutralization capacitor are short. This circuit requires less involved switching than in the cascode circuit, but requires one more switch contact than a pentode circuit.

Direct-Coupled Driven-Grounded-Grid Circuit

Figure 4 is another version of the driven-grounded-grid circuit in which the plate of the input triode is directly coupled to the cathode of the output triode. Neutralization is accomplished in the same manner as previously described. This circuit has the advantage that several components are eliminated from the coupling network between the two units; consequently, the distributed capacitance to ground is reduced and the gain at the higher channels is increased. Another important advantage is that application of bias to the input triode causes the voltage between plate and cathode to increase, extending the cutoff of the tube. This extension reduces cross-modulation, without the use of a remote-cutoff tube. Such a tube would adversely affect the signal-to-noise ratio, either by increasing input loading or by reducing transconductance. Because of lower distributed capacitance between the plate circuit of the input triode and ground, this amplifier

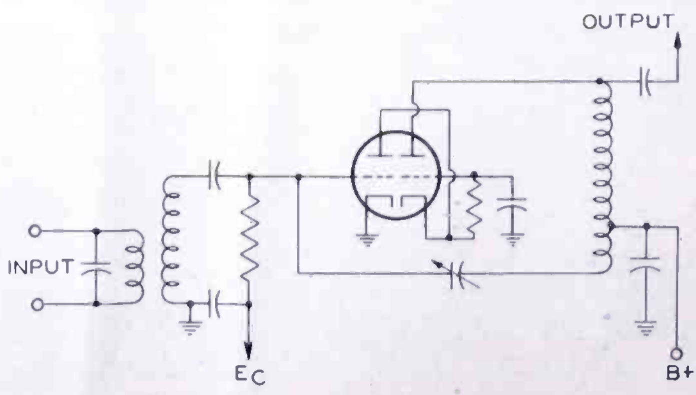

Fig. 4 — Direct-coupled driven-grounded-grid circuit.

can give fairly satisfactory results on the low channels without neutralization. It is interesting to note that the number of components in this circuit equals the number required for a conventional pentode amplifier, the grid resistor and bypass capacitor in the grounded-grid triode circuit being equivalent to the screen resistor and capacitor of the pentode circuit.

The foregoing driven-grounded-grid circuits have an input impedance and an admittance from output to input terminals which are dependent to a large extent on certain characteristics of the tubes employed. The 6BQ7 is primarily designed to provide the characteristics needed for good performance in the various driven-grounded-grid circuits. The following detailed description of the 6BQ7 correlates its design features and electrical characteristics with the specific requirements of the driven-grounded-grid circuits.

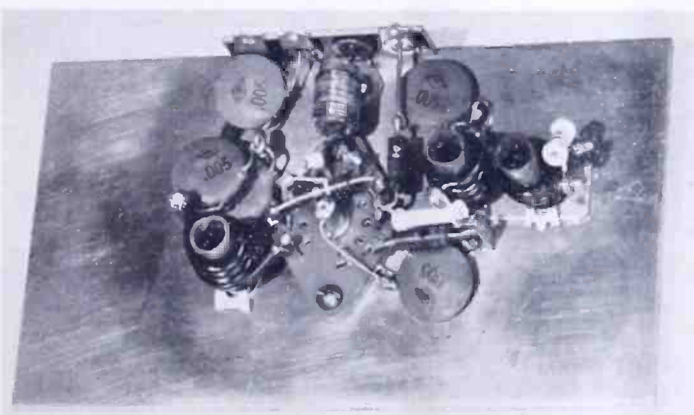

Fig. 5—Layout for direct-coupled driven-grounded-grid circuit using 6BQ7.

Characteristics of the 6BQ7

Unit one of the 6BQ7 has the plate, grid, and cathode connected to pins 6, 7, and 8 respectively; unit two, which is electrically identical with unit one, has these elements connected to pins 1, 2, and 3 respectively. It is recommended that unit one be used for the input section and unit two for the output grounded-grid section. This connection permits the use of shorter leads and consequently results in less capacitance between the plate and cathode leads of the output section. The shield between the sections aids in preventing excessive coupling between the units. Figure 5 indicates how the selected basing arrangement simplifies wiring and layout.

Table I indicates important 6BQ7 electrical characteristics and ratings. The two units are identical so that the tube can be used not only in the cascode and driven-grounded-grid circuits, but also in push-pull grounded-grid amplifiers, high-frequency counter circuits, and other applications. Fortunately, it is possible to make the tube units identical without any compromise in grounded-grid circuit performance. The more versatile arrangement should result in higher-volume production and reduced cost. The transconductance value of 6000 micromhos obtained at a plate current of only 9 milliamperes results in high gain and a reduction of equivalent noise resistance. The use of fine grid laterals and close spacing between grid and cathode accounts for this unusually high ratio of transconductance to plate current. The shield used in the 6BQ7 is provided by a shaped grid connector which effectively reduces the plate-cathode capacitance to an average value of 0.135 micromicrofarads without increasing the other critical capacitances. This method of shielding permits either triode to be used for grounded-grid or grounded-cathode operation.

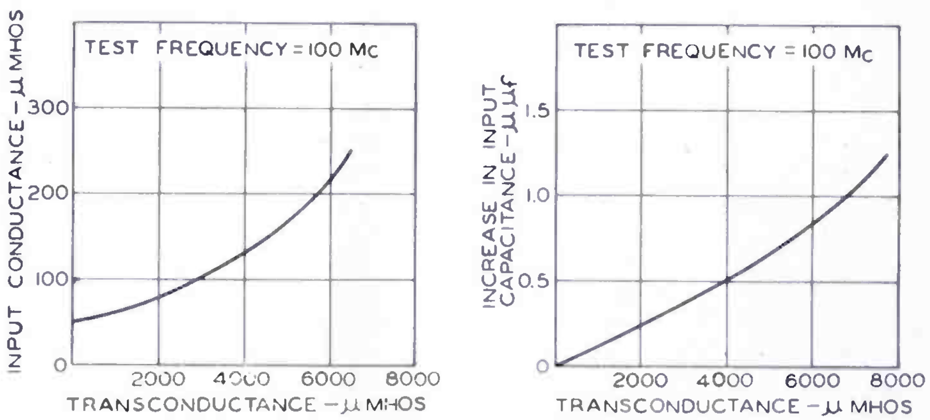

Fig. 6—Variation of input conductance and input capacitance with transconductance in 6BQ7.

Figure 6 gives the input admittance characteristics of the 6BQ7.[10] Because induced grid noise increases with input conductance when high-impedance input circuits are used, low values of input conductance are desirable. The theoretical noise figure for the 6BQ7 obtained by the methods given by Herold[11] and Harris[12] is 3.1 decibels at 70 megacycles and 8.6 decibels at 200 megacycles, if it is assumed that the input circuit is impedance-matched, has a 6-megacycle bandwidth and does not benefit from coherence between plate and grid noise. The minimum theoretical noise factors of the 6BQ7 at these frequencies are 3 decibels and 6.4 decibels, respectively.

[10] Taken on admittance meter described in *RCA Application Note AN-118*, April 15, 1947.

[11] E. W. Herold, "An Analysis of the Signal-to-Noise Ratio of Ultra-High-Frequency Receivers", *RCA Review*, Vol. VI, No. 3, pp. 302-331, January, 1942.

[12] W. A. Harris, "Some Notes on Noise Theory and its Application to Input Circuit Design", *RCA Review*, Vol. IX, No. 3, pp. 406-418, September, 1948.

In addition to influencing the generation of noise, too high an input conductance may limit the voltage gain from the antenna to the input grid. The input conductance of the 6BQ7 is only 200 micromhos at 100 megacycles and 800 micromhos at 200 megacycles. This latter value, equivalent to an input resistance of 1250 ohms, permits an antenna voltage gain of greater than two in the high-frequency channels, if a matched-impedance input circuit is used. As shown in Figure 6, the input conductance of the 6BQ7 decreases as bias voltage to the control grid is increased. A damping resistor of 10,000 ohms in shunt with the grid circuit is recommended to prevent excessive changes in bandwidth and input impedance as a result of variations in automatic-gain-control bias. The indicated change of input capacitance with bias is sufficient to cause noticeable detuning of the input circuit. When the tube is operated with an unbypassed cathode resistance of 68 ohms, the change of input capacitance with bias is reduced to a negligible value and the variation in conductance is also reduced. However, the resultant degeneration reduces the effective transconductance to 5150 micromhos or by approximately 14 per cent. This degeneration causes a proportionate reduction in gain but does not affect the noise factor. When the tube is used in the series-connected circuit with no unbypassed cathode resistor, the minimum allowable bias is 1.25 volts. When the bias is varied from 1.25 volts to cutoff, the change in input capacitance is 0.3 micromicrofarads. The resultant detuning will be approximately 1.7 megacycles in a high-impedance input circuit having a total capacitance of 20 micromicrofarads shunting the input coil and tuned to channel thirteen. This value of detuning is lower than that which occurs with other tubes.

Figure 7a gives the plate family of characteristic curves for the 6BQ7. The tube has a sharp-cutoff characteristic which results in low input loading, although the cross-modulation is thereby increased to a degree comparable to that obtained with the pentodes. When the series-connected direct-coupled circuit is used, the overall plate characteristic curve for the two tubes is that shown in Figure 7b. The cutoff is extended by a factor of two, without adverse effect on the input loading.

Because the curve more nearly approaches a square-law characteristic, the theoretical requisite for absence of cross-modulation, this type of interference is greatly reduced. Interference measurements indicate that cross-modulation with the direct-coupled circuit is one-eighth that with the capacitively coupled circuit, an improvement which agrees well with theoretical calculation. Figure 7c shows the transconductance variation with bias voltage for the single triode unit and for the series-connected arrangement.

Finally, it is necessary to consider the effect of the radio-frequency amplifier on oscillator radiation. Oscillator radiation is a function of the capacitance between the output plate and input grid and of the terminal impedances, as shown in Figure 8. Most of the attenuation occurs between the plate and cathode of the grounded-grid unit because of the low value of the parallel combination of the plate-to-cathode and plate-to-plate capacitances, and also because of the low impedance between the cathode and ground. The plate-to-plate capacitance, C_2, is reduced to a very low value by the shield between the units. The total voltage attenuation is calculated to be 35.6 decibels at 200 megacycles, assuming the plate-to-grid capacitance of the input unit is not neutralized.

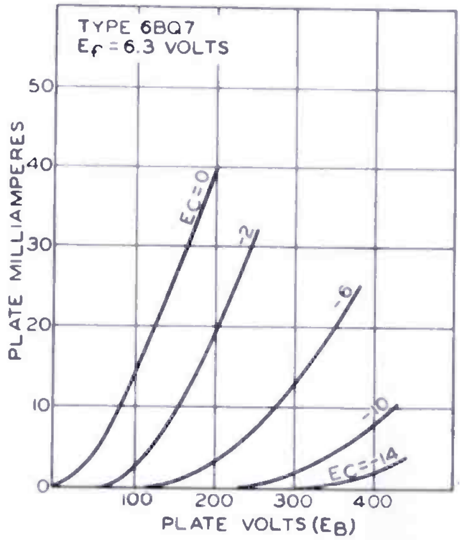

Fig. 7a—Average plate characteristics of each unit.

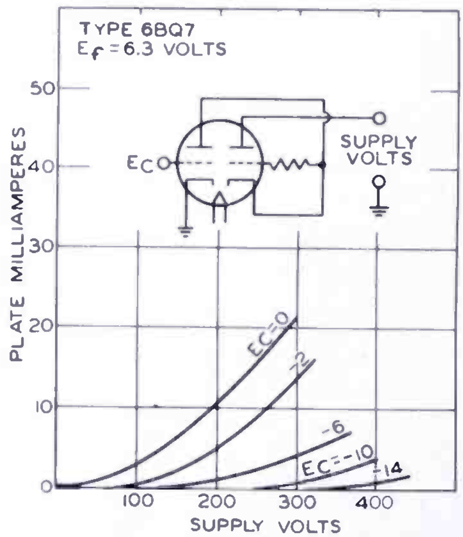

Fig. 7b—Average plate characteristics, series-connected.

This attenuation is slightly larger than that for pentode circuits, and is several times larger than that of the grounded-grid arrangement. Measurements of the actual oscillator radiation from the neutralized circuit indicate that the oscillator radiation is not appreciably affected by the neutralization. Operation without neutralization is considered later when the practical problems involved in applying the circuits to a television receiver are discussed.

Performance Measurements

There are no standard methods of measuring the performance of a tube in tuner circuits. It is possible to obtain a tuner having a pentode radio-frequency stage, measure the performance of the tuner, replace the radio-frequency stage with the 6BQ7 in suitable circuits, and repeat the measurements. However, to compare one circuit arrangement with another would necessitate rebuilding the radio-frequency stage of the tuner for each circuit tried, a laborious and time-consuming process. The results obtained would be dependent to some extent on

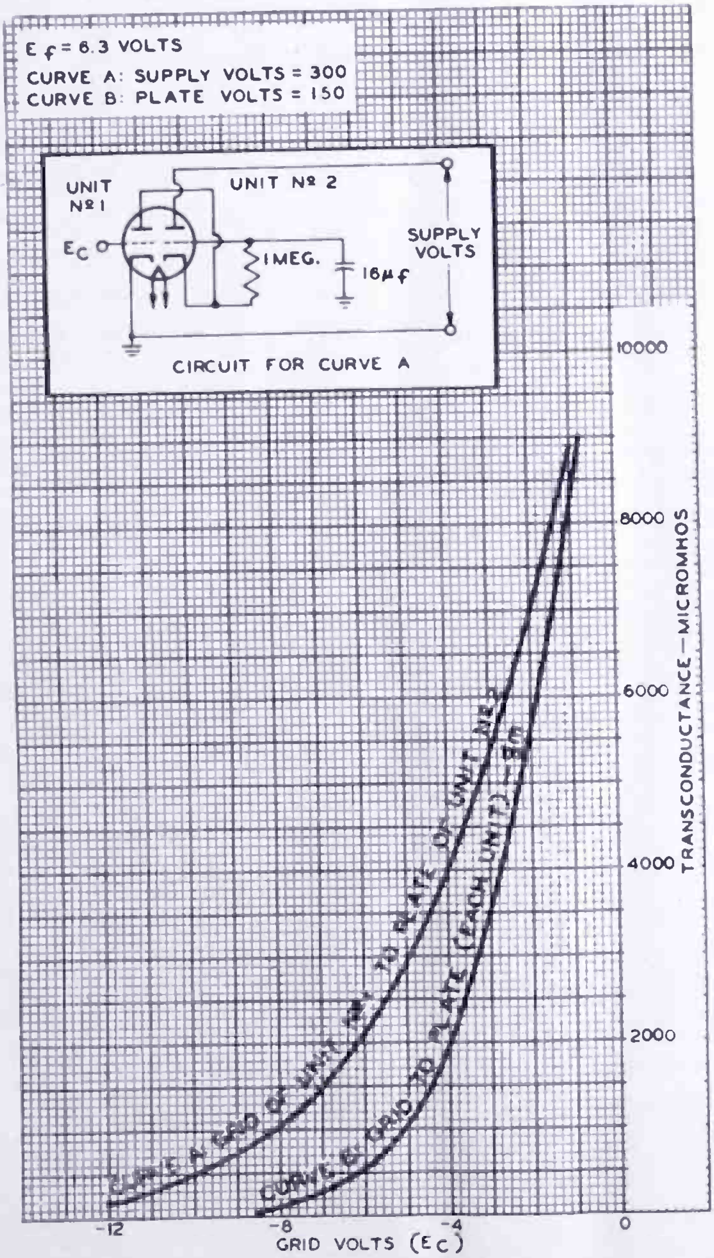

Fig. 7c — Variation of transconductance with bias voltage for single 6BQ7 unit and for series-connected arrangement.

the particular mechanical arrangement of the tuner selected and would not necessarily be indicative of what could be expected from other types of tuners. The switching mechanism introduces additional variables in the form of inductance and shunt capacitances which limit the performance of the system. To avoid speculative valuations of

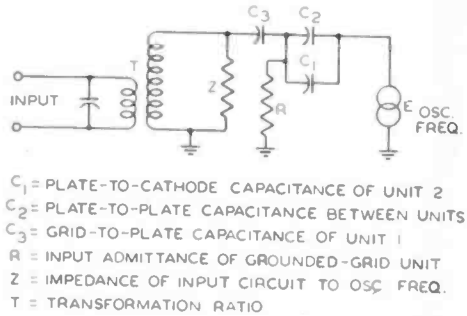

Fig. 8—Equivalent circuit for oscillator radiation of driven-grounded-grid circuit.

these limitations, tuner measurements were made on a breadboard tuner having no switches or turrets. The results, while admittedly optimistic, are at least indicative of tube capabilities and may be conveniently compared with similarly obtained data[1] on other tubes which are currently used in tuners. The circuit which performed best in the breadboard arrangement was installed in a turret-type tuner and the two sets of data were compared in order to evaluate the relative efficiency of the tuning unit alone. The performance is believed to be typical of what can be expected from turret-type tuners.

Figure 9 shows the block diagram of the test setup employed in measuring performance. Data on each unit are obtained on channels four, eleven, and thirteen. Noise measurements are made using a square-law vacuum-tube voltmeter and a high-gain intermediate-frequency amplifier having a bandwidth of 4.5 megacycles.

The signal generator has a balanced output with a matching resistor in each conductor to provide a total balanced impedance of 300 ohms. A noise generator incorporating an emission-limited diode is used to check the signal-generator calibration. The noise information is presented as a noise figure which indicates the ratio of the noise produced by the receiver to that of an ideal system having as a source of noise only the 300-ohm antenna resistance. The noise produced by the receiver is measured by reducing the output of the signal generator to zero and noting the output on the meter. This output is proportional to the square of the amplified noise voltage and thus is a function of noise power. An unmodulated carrier is then applied to the input terminals by the generator and the value of the input signal is adjusted to double the noise reading at the output. The receiver noise, referred to the input terminals, is equal to the signal-generator input voltage; the ratio of this voltage to that calculated to be produced by

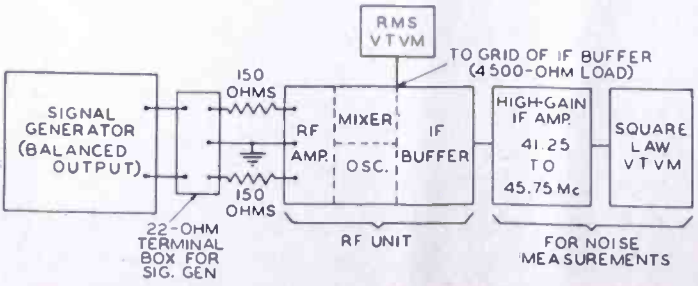

Fig. 9—Block diagram of test set-up.

the 300-ohm antenna resistance is expressed in decibel relationship as the noise figure.

Gain figures are obtained in the following manner. The 6X8, a new triode-pentode, is used as an oscillator-mixer and its output is measured across a 4500-ohm load at the grid of the first intermediate-frequency amplifier stage. The voltage output divided by the value of signal input voltage as indicated on the signal generator is the overall gain from the antenna to the first amplifier, including the mixer gain which is maintained constant for the various circuits tested. This method avoids the uncertainty involved in direct measurement of a 200-megacycle signal. The frequency of the mixer output voltage is 45 megacycles, at which frequency it is possible to measure voltages with good accuracy. Values for the gain of the radio-frequency stage alone are closely equal to the overall gain divided by the mixer gain, which is approximately five.

Antenna termination is measured by determining the standing-wave ratio of the transmission line by means of a small single-turn loop tuned to the signal frequency and loosely coupled to the line. The energy picked up by the loop is applied to the input of a sensitive receiver having a calibrated diode detector at the intermediate-frequency output. The values shown are voltage standing-wave ratios measured along a 75-foot 300-ohm twin-lead line having the antenna end connected to the signal generator. Placing a 5000-ohm resistor in series with each lead effectively makes the antenna end of the line an open circuit. When the receiver properly terminates the transmission line, there are no standing waves on the line.

Measurements of the performance of the 6BQ7 in the push-pull grounded-grid circuit, the driven grounded-grid circuit, and the directly coupled driven grounded-grid circuit are presented in Table II and are compared with measurements on other tube types in Table III. No data were taken on the neutralized triode and cascode amplifiers because neutralization difficulties make these circuits unsuitable for this application.

Figure 10 shows the push-pull grounded-grid circuit. Bifilar chokes are used in the heater circuit to prevent variations in heater-cathode capacitance from affecting the tuning of the input circuit. For optimum performance, the input circuit should match the antenna to the low input impedance of the tube, and the bandwidth of the circuit should not exceed six megacycles. Since the input circuit is heavily loaded by the low input resistance of the tube, the circuit must have an extremely low L-to-C ratio in order to meet this bandwidth requirement. The requirement cannot quite be met on the higher channels.

For example, at 213 megacycles (the center of the highest very-high-frequency channel), the capacitance needed across a parallel tuned circuit to obtain the required selectivity is 87 micromicrofarads; the inductance required has the unattainably low value of 0.0064 microhenry.

Extreme care must be taken to ground the grid with the shortest practical lead, and wafer sockets of a special design having the contact lug emerging from the edge of the socket between the wafers are recommended. These sockets should also be used in the other circuits which require the grounded-grid connection. No automatic-gain-control bias is applied to the amplifier since the resultant change in input impedance of the tube would be intolerable from an antenna termination standpoint. The performance data for this circuit is shown in Table II. It is probable that the figures shown here for image rejection are somewhat optimistic since it is easier to develop

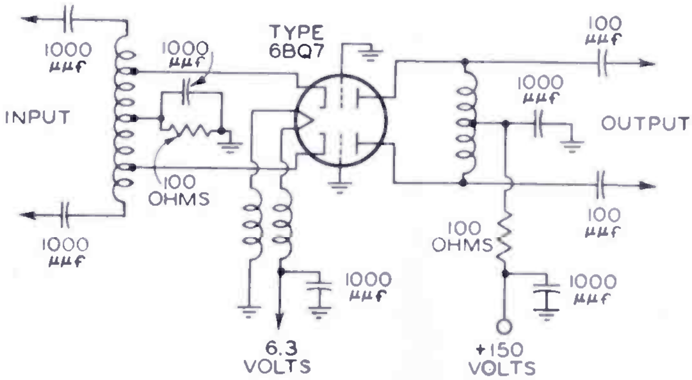

Fig. 10 — Push-pull grounded-grid circuit.

the proper input circuit on the experimental breadboard than it would be in a commercial tuner or preamplifier.

Figure 11 shows the neutralized driven-grounded-grid circuit having conventional capacitance coupling between tube units. On low-frequency channels, optimum performance occurs when the input circuit is double-tuned, the primary is matched to the 300-ohm antenna, and the secondary is operating at the highest impedance attainable without reducing the bandwidth to less than six megacycles. On the high channels, very close coupling is used in the input transformer to reflect a low value of impedance to the secondary winding. The input transformer should be slightly overcoupled, and it is desirable to reduce the capacitive coupling by winding the secondary in a figure-eight configuration or by using electrostatic shielding. One convenient way of obtaining such shielding is to place some high-dielectric-constant ceramic material between the coils with the edge of the ceramic shield grounded, so as to effectively short circuit to ground the capacitance between the shield and each coil. When the grounded-grid unit is wired, care must be taken to place the leads so that the plate-cathode capacitance is not unduly increased, or the grounded-grid stage will oscillate. Because the heater leads and the plate of the input stage are at the radio-frequency potential of the cathode of the output unit, they must be placed so that coupling to the output plate is avoided. The small capacitor shunting the plate coil couples one side of the coil to the other in order to avoid a parasitic effect. Without this capacitor, the portion of the coil between the tap and the lower end of the coil forms a series circuit with the neutralizing capacitor which absorbs signal energy from the input circuit; furthermore, the lower

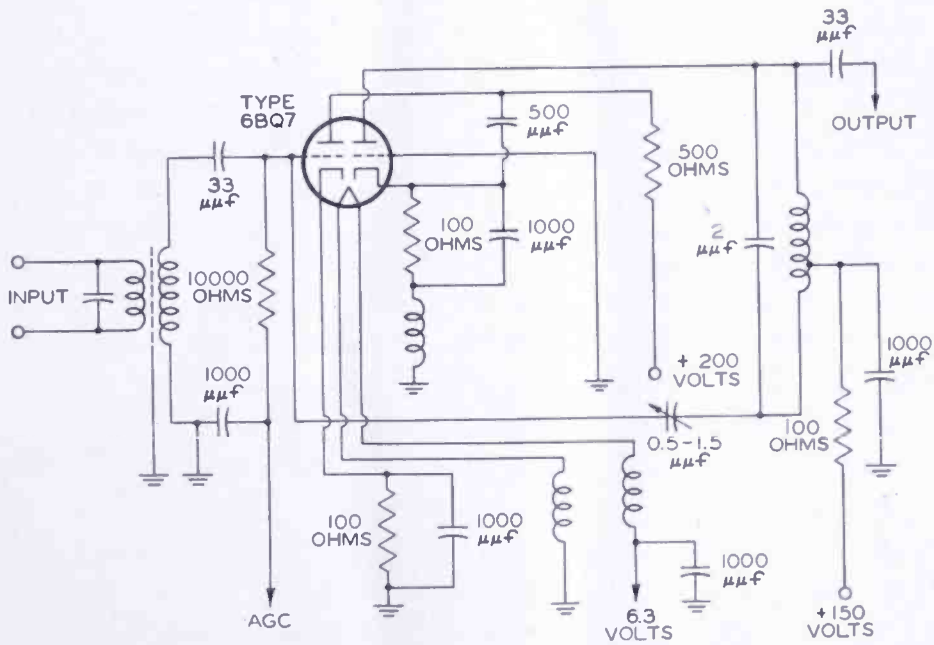

Fig. 11—Driven-grounded-grid circuit.

section of the coil has no effect on the resonant frequency of the output circuit. These effects are due to insufficient coupling between the two sections of the coil. The addition of the tank capacitor proves to be the most practical method of obtaining the necessary coupling.

The neutralization method shown is quite practical when applied to a turret-type tuner, requiring only one additional switch contact and a non-critical adjustment; it is not practical for the conventional switch-type tuner unless an extra switch wafer is provided. When this circuit is not neutralized, serious degeneration results at the higher frequency bands, but operation at the lower frequencies is only slightly impaired.

As a result, the following arrangement appears to be a practical way of operating driven-grounded-grid circuits in the very-high-frequency tuner application when it is desired to reduce the neutralization cost. The heater chokes are adjusted to be approximately in resonance with the plate-to-ground capacitance of the first unit at a frequency of 200 megacycles. Since the amount of degeneration is a function of the magnitude of the capacitance from plate to ground, tuning out this capacitance eliminates degeneration. The resultant resonance is effective throughout the high-frequency band because the coil is heavily damped by the low and unvarying input impedance of the grounded-grid unit. No neutralization is provided for the low band. Such neutralization improves the noise figure only by one decibel and requires additional circuit complexity which is economically unjustified.

It should be noted that the tuning out of the plate-to-ground capacitance of the input unit mentioned above is not completely realized. The capacitance actually shunting the heater chokes consists of two parallel components: namely, the heater-to-cathode capacitance of the input unit, and the series combination of the heater-to-cathode capacitance of the second unit and the capacitance between the cathode of the second unit and ground. Because the heater-to-cathode capacitance is of the same order of magnitude as the cathode-to-ground capacitance of the output unit, the cathode is effectively tapped down on the resonant circuit. Any attempt to remedy this situation by increasing the heater-to-cathode capacitance is impractical, because such a step causes deterioration of performance on the low channels.

Table II compares the performance of the driven-grounded-grid circuit for the two methods of operation. With the heater-choke arrangement, the noise factor is greatest but only by two decibels in the worst case, on channel six. The standing-wave ratio is satisfactory and is comparable to that of the better pentode circuits. Cross-modulation in the radio-frequency amplifier is comparable to that experienced with the sharp-cutoff type of pentode tubes now used.

Figure 12a shows the direct-coupled driven-grounded-grid circuit which gives the most satisfactory performance of the various circuit arrangements tested. The reduction of distributed wiring capacitance in the coupling circuit results in higher gain and lower noise as shown in Table II. Antenna termination and cross-modulation are improved considerably over the results obtained with the capacitor-coupled circuit. Operation of the series circuit with a minimum bias of —2 volts, which is recommended to minimize variation in input admittance, results in improved termination without impairment of noise figure. The extension of cutoff as the result of the series connection reduces the cross-modulation.

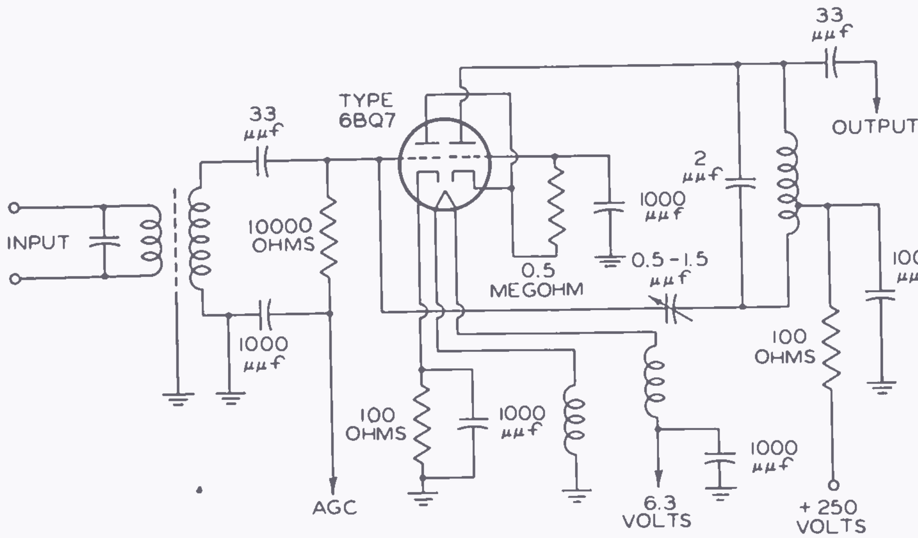

Fig. 12a—Direct-coupled driven-grounded-grid circuit.

Another modification of the driven-grounded-grid circuit[13] is that shown in Figure 12b. Here, the heater chokes, which are nonresonant at the television frequencies, are used only to reduce undesirable microphonic effects caused by heater-cathode capacitance variations. Coil L_1 and the distributed circuit capacitance C_d between the cathode of the output unit and ground, are series resonant at a frequency of 200 megacycles. This series circuit presents a very low impedance between

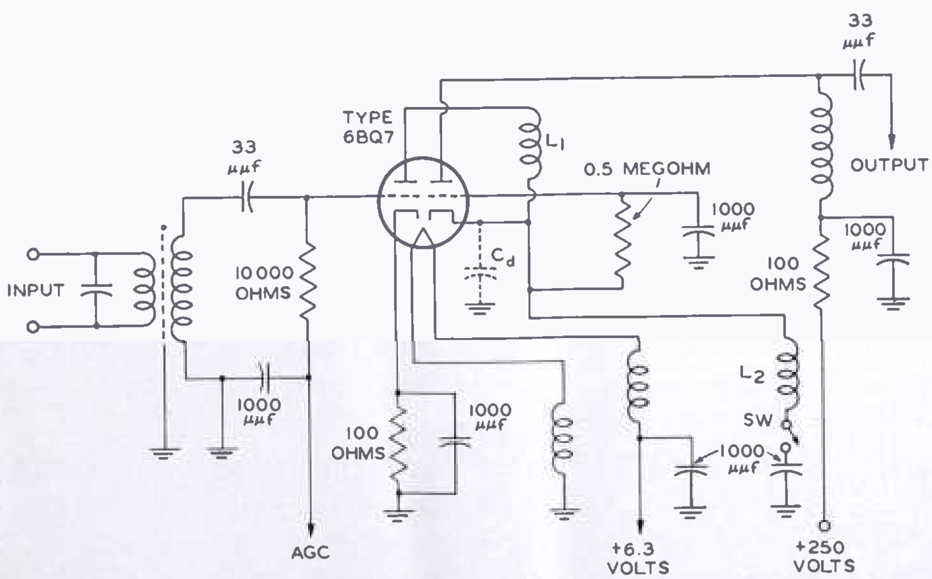

Fig. 12b—Alternate method of utilizing direct-coupled driven-grounded-grid circuit. (L_1 = 8 turns 24E, 3/16-inch diameter air core, spaced one diameter. L_2 = 16 turns 24E, 3/16-inch diameter air core, spaced one diameter.)

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[13] This circuit was developed by J. C. Achenbach and P. C. Swierczak of Home Instrument Department, RCA Victor Division, Camden, N. J.

the input plate and ground, thereby reducing the radio-frequency voltage on the input plate sufficiently to make conventional neutralization unnecessary. The fact that C_d is shunted by the input impedance of the output unit limits the resonant voltage across C_d to a value which is nearly equal to that applied to the input grid.

The high-channel performance of this circuit equals that obtained with feedback neutralization. Like the previously discussed heater-choke arrangement, this circuit has a sufficiently wide frequency response to permit the use of fixed components. This circuit has the additional advantages of lower cost and greater ease of adjustment.

If better low-channel noise performance is desired in this circuit, an improvement of approximately one-decibel results when a modification of the series-tuned circuit is used. Coil L_2 and blocking capacitor C in series form a parallel circuit with C_d that is resonant at the center of the low-frequency band. For high-band operation, L_2 is switched out at the low-impedance terminal. This coil must be so positioned that the capacitance between it and ground is minimized.

Table III shows noise data taken on other tubes in typical circuits and is included for comparison purposes. The data are obtained with similar testing methods and show the relative merits of the 6BQ7. Only the push-pull neutralized 6J6 affords comparable results, but this circuit is impractical because of difficulties in neutralization.

Practical Results in a Turret Tuner

It was thought desirable to substantiate the encouraging results obtained in the laboratory test set-ups with field tests in commercial-type receivers. The direct-coupled driven-grounded-grid circuit, because it provides the most satisfactory operation, was built into a turret-type television tuner which originally had a 6CB6 pentode stage. The following is a discussion of the problems experienced in installing this improved tuner in a television receiver.

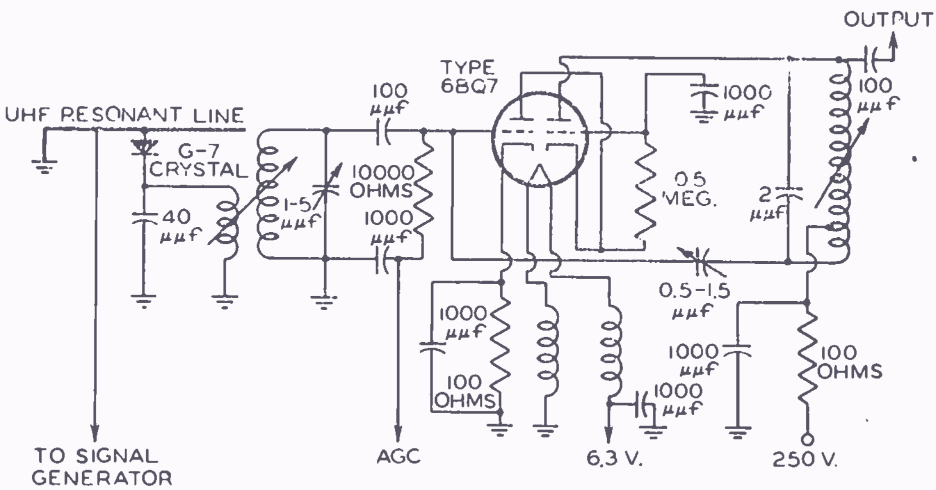
As the sensitivity of the receiver is increased, its susceptibility to interference from other sections of the receiver which radiate energy also increases. The magnetic-deflection system usually generates a multitude of extraneous frequencies which are capable of causing interference at the signal frequency, either by heterodyning with harmonics of the local oscillator, or by cross-modulation. The elimination of radio-frequency interference produced by the deflection system is a subject in itself and will not be discussed other than to mention that additional shielding and supply-voltage filtering were found to be necessary. Another source of interference which will require additional shielding in most receivers is radiation from the second detector, the harmonics of which couple back to the radio-frequency circuits. In the receiver used, this type of interference was particularly troublesome; it was finally limited to reasonable proportions by extensive shielding of the detector circuit.

The action of the automatic-gain-control system is another important factor in affecting the results obtained in receivers. Too early an application of control voltage to the radio-frequency amplifier reduces its gain, permitting the converter noise to add to the overall noise figure. On the other hand, too great a delay in the application of automatic-gain-control voltage to the radio-frequency stage causes cross-modulation. The correct adjustment is therefore a compromise. The relationship between radio-frequency and intermediate-frequency bias-control voltages must be carefully selected because it has an appreciable effect on overall performance. To take full advantage of the remote cutoff characteristics of the radio-frequency amplifier in the direct-coupled circuit, an amplified automatic-gain-control system appears desirable.

As expected, some degradation in performance is experienced when the tube and circuit are installed in the turret-type tuner. Measurements were made first with the circuit having feedback neutralization, and then with the circuit having resonant heater chokes. Table IV includes data taken before the tuner was converted, thus indicating the degree of improvement directly attributable to the use of the 6BQ7. Field tests were made with the receiver and the results confirmed the laboratory data to a satisfactory degree.

Use of the 6BQ7 in Ultra-High-Frequency Receivers

Analysis of the ultra-high-frequency tuner problem for a proposed carrier frequency range of approximately 470 to 890 megacycles in a receiver having an intermediate frequency of 43 megacycles and a crystal mixer, reveals that the characteristics of the first intermediate-frequency amplifier tube are important in determining the noise figure. To a close approximation, the noise figure of the intermediate-frequency system, determined by the first intermediate-frequency amplifier stage, added to the noise figure of the radio-frequency system equals the overall noise figure. Unless radio-frequency amplifier tubes are used ahead of the crystal mixer, the signal level at the grid of the first intermediate-frequency tube will be low due to the approximately 9 decibels of attenuation in the crystal mixer stage. An intermediate-frequency preamplifier stage is needed to provide an overall receiver gain on the ultra-high-frequency band equal to that obtained on the very-high-frequency bands. It is a relatively simple problem to use the 6BQ7 as an intermediate-frequency preamplifier stage at 43 megacycles, either in the cascode circuit or in the driven-grounded-grid circuit. Because of the very low plate-cathode capacitance of the 6BQ7, it is not necessary to neutralize the output section when the circuit wiring is carefully oriented to avoid excessive coupling from plate to cathode. The input section should be neutralized to avoid degeneration and a resultant 2-decibel noise increase. The noise figure is approximately 4.5 decibels in a neutralized circuit operating at a center frequency of 43 megacycles and with a five-megacycle bandwidth. The use of the direct-coupled circuit has advantages with respect to cross-modulation and is recommended when gain-control voltage is to be applied to the intermediate-frequency preamplifier. Figure 13 shows



BANDWIDTH	LOAD	GAIN*	NOISE FACTOR	SELECTIVITY
4.5 Mc	4000 OHMS	24	4.5 DB	10.2 DB AT 10 Mc

* FROM GRID OF 6BQ7 TO GRID OF FIRST IF AMP.

Fig. 13—Intermediate-preamplifier for ultra-high-frequency applications.

the crystal mixer and the 6BQ7 as the intermediate-frequency preamplifier. The low output impedance of the crystal mixer, in the order of 300 ohms, should be matched to the input circuit of the intermediate-frequency tube which should have as high an impedance as it is practical to develop without reducing the required bandwidth.

If the total input capacitance is 15 micromicrofarads, it should be possible to develop an input impedance of 2100 ohms. The input resistance of the 6BQ7, which at this frequency is approximately 20,000 ohms, is, therefore, not a limiting factor in obtaining the required circuit impedance, but must be considered when the required value of damping resistance is calculated. It should also be recognized that the input resistance increases rapidly with the application of bias voltage, and it may be necessary to use a lower value of damping resistance to effect a satisfactory compromise. An unbypassed cathode resistance of 68 ohms may also be used to minimize the variation in input admittance.

Further References

"Grounded-Grid Amplifiers," by E. E. Spitzer in *RCA Broadcast News,* October 1946, pages 66-69, and in *Electronics,* April 1946, pages 136-142.

"Low Noise Preamplifier," by F. B. Llewellyn in *Electronics,* April 1946, page 97.

"Cathode-Coupled Wide-Band Amplifiers," by G. C. Sziklai and A. C. Schroeder in *Proc. IRE,* October 1945, pages 701-708.

Patents — USP 2460907, Schroeder; 2093078, R. A. Husing (Filed 1934); 1986597, A. Nyman (Filed 1931); 1886386, D. T. Francis (Filed 1928).

Acknowledgment

The author wishes to acknowledge the valuable work contributed by H. J. Prager and J. Johnston in the design of the 6BQ7 and by E. M. Troy in establishing electrical ratings and characteristics.

Table I — RCA 6BQ7 Electrical Characteristics and Tentative Ratings

Direct Interelectrode Capacitances (micromicrofarads)

	Unit 1	Unit 2
Gride to Plate	1.15	1.15
Plate to Cathode	0.15 max.	0.15 max.
Heater to Cathode	2.20	2.30
Input	2.85	4.95*
Output	1.35	2.27*
Plate of Unit 1 to Plate of Unit 2	0.010 max.	
Plate of Unit 2 to Plate & Grid of Unit 1	0.024 max.	

Class A_1 Amplifier

Max. Ratings, Design-Center Values: (each unit)

DC Plate Volts	250 max. volts†
DC Cathode Current	20 max. milliamperes
Plate Dissipation	2.0 max. watts
Peak Heater-Cathode Volts	
Positive	200 max. volts
Negative	200 max. volts†

* Read as grounded-grid amplifier.
† This rating may be as high as 300 volts under cutoff conditions.

Table I — (Continued)

Characteristics (each unit)	
Plate Volts	150 volts
Cathode-Bias Resistor	220 ohms
Amplification Factor	35
Transconductance	6000 micromhos
Plate Resistance	5800 ohms
Plate Current	9 milliamperes
Maximum Circuit Values (each unit)	
Grid-Circuit Resistance	0.5 max. megohm
Typical Operation in Push-Pull Grounded-Grid Circuit (Values are for each unit)	
Plate Volts	150 volts
Grid Volts‡	—2 volts
Plate Current	10 milliamperes
Cathode Resistor (common to both units)	100 ohms

Typical Operation in Driven-Grounded-Grid Circuit with Direct-Coupled Drive. Unit 1 (driver unit) is directly coupled to Unit 2 (driven-grounded-grid amplifier unit).

	Unit 1	Unit 2
Plate-Supply Volts	250	250 volts
Plate Volts	135	115 volts
Grid Volts	— 1	— volt
Grid-Resistor	—	0.5 megohm
Plate Current	10	10 milliamperes
Grid Current	0	0 milliamperes
Grid Volts (Approximate) for plate current of 10 microamperes	—14	— volts
Heater-Cathode Volts Heater negative with respect to cathode	—	225 volts

‡ Obtained from cathode resistor.

Table II — Tabulation of Circuit Performances with 6BQ7

Circuit	Channel	RF Gain	Overall Gain	Noise Figure (db)	Image Rejection (db)	Standing Wave Ratio
Push-pull grounded-grid	4	9	45	7	42	less than 1.1 when no AGC is used
	11	9	45	7	38	
	13	8.5	42	7	35	
Driven-grounded grid (a) feed-back capacitor	4	15	75	6.8	45	1.15
	11	14	70	7.0	42	1.2
	13	14	70	7.2	42	1.2
(b) resonant heater choke	4	14	70	7.1	42	1.15
	11	12	60	8.1	42	1.2
	13	12.5	62.5	7.9	41	1.2
Direct-coupled, driven-grounded-grid	4	17	85	6.0	45	1.15
	11	16	80	6.0	42	1.2
	13	16	80	6.0	42	1.2

Table III — Comparison of 6BQ7 with Other Tubes in Typical Circuits

Tube Type	Circuit	Measured Gain Channel No. 4	Measured Gain Channel No. 11	Image Rejection (db) Channel No. 4	Image Rejection (db) Channel No. 11	Noise Figure (db) Channel No. 4	Noise Figure (db) Channel No. 11
6J6	Push-pull neutralized grounded-cathode with untuned input	60	60	35	35	13	13
6J6	As above with tuned input and neutralized	120	120	45	45	6	6
6AU6	Grid-cathode input input circuit untuned	30	25	35	30	20	20
6J6	Grounded-grid RF amplifier and 6J6 cathode-coupled mixer	15	15	40	40	8	9
6J4	Grounded-grid RF amplifier and 6J6 cathode-coupled mixer	30	30	40	40	6	6
6BQ7	Direct-coupled driven grounded-grid circuit	85	80	45	42	6	6

Table IV — Performance Data of Turret-Type Tuner

RF Amplifier	Channel	RF Gain	Noise Figure (db)	Image Rejection (db)	Standing Wave Ratio
Type 6CB6	2	14	7.2	80	1.6
	6	11	11.0	70	1.2
	7	9	13.2	70	1.3
	13	8	14.0	70	1.35
Type 6BQ7 in direct-coupled circuit having resonant heater chokes	2	14	7.1	80	1.25
	6	12	7.8	70	1.2
	7	12	8.1	70	1.25
	13	12.5	8.9	70	1.25
Type 6BQ7 in direct-coupled circuit using feedback neutralization	2	15.5	6.8	80	1.25
	6	15.0	7.2	70	1.25
	7	14.0	8.1	70	1.25
	13	14.0	8.5	70	1.25

SHORTWAVE RADIO PROPAGATION CORRELATION WITH PLANETARY POSITIONS*

By

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Summary—An examination of shortwave radio propagation conditions over the North Atlantic for a five-year period, and the relative position of the planets in the solar system, discloses some very interesting correlations. As a result of such correlations, certain planetary relationships are deduced to have specific effect on radio propagation through their influence upon the sun. Further investigation is required to fully explore the effect of planet positions on radio propagation in order that the highly important field of radio weather forecasting may be properly developed.

Introduction

MANY investigators of solar activity in the past have conducted extensive studies of planetary phenomena in an effort to account for the maximum and minimum of the eleven-year sunspot cycle and also the shorter period variations in sunspot numbers which take place from month to month. The results of several of these investigators appear to indicate a connection between the interrelationship of the planets and the degree of spottedness of the solar surface. The works of Huntington,[1] Clayton,[2] and Sanford,[3] were found to be particularly applicable to the subject matter of this paper.

The results of their investigations suggested that a similar study relating planetary phenomena to radio disturbances over the North Atlantic might reveal information of value. Since June, 1948, the author has conducted research on this subject, and this paper presents the correlation that has been found between shortwave radio disturbances and certain planetary phenomena as described below.

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* Decimal Classification: R 113.216.

[1] E. Huntington, EARTH AND SUN, Yale University Press, New Haven, Connecticut, 1923.

[2] H. H. Clayton, SOLAR RELATIONS, Clayton Weather Service, Canton, Massachusetts, 1943.

[3] F. Sanford, INFLUENCE OF PLANETARY CONFIGURATIONS UPON THE FREQUENCY OF VISIBLE SUNSPOTS, Smithsonian Institution, Washington, D. C., 1936.

Motion of the Planets

The heliocentric interrelationship between Mercury, Venus, Earth, Mars, Jupiter and Saturn was extracted from the AMERICAN EPHEMERIS AND NAUTICAL ALMANAC published by the U. S. Naval Observatory in Washington, D. C. for the years 1942, 1944, 1947, 1948, and 1949. Dates when the heliocentric relationship of any two planets was 0°, 90°, 180° or 270° were recorded. At 0° an inner planet is in line on the same side of the sun with an outer planet; at 90° an inner planet is 90° *ahead* of an outer planet; at 180° an inner planet and an outer planet are in line on opposite sides of the sun; at 270° an inner planet is 90° *behind* an outer planet. These relations are hereinafter referred to as configurations.

In addition to plotting the positions of the various planets in this way, a record was made of the solar quadrants over which each configuration took place. The solar quadrants in this study are determined by the Earth-Sun relationship under which the Sun is divided into its four quadrants as follows: The first quadrant is the visible sector of the eastern hemisphere of the Sun as viewed from the Earth; the second quandrant is the invisible sector of the eastern hemisphere; the third quadrant is the invisible sector of the western hemisphere; and the fourth quadrant the visible sector of the western hemisphere.

Every configuration of the type previously mentioned was calculated between Mercury and Venus, Mercury and Earth, Mercury and Mars, Mercury and Jupiter, and Mercury and Saturn. Following this the same method was used with Venus and its outer planets, Earth and its outer planets, Mars and its outer planets, and Jupiter and Saturn. The analysis shows that these configurations are quite random in time and vary from cases where only one configuration between two planets takes place in a 14-day period to cases in which five of these six planets are involved in a configuration with some other planet within a forty-eight hour period. Cases where an inner planet makes a configuration with two outer planets within a twenty-four hour period are quite numerous.

At perihelion, Mercury moves 6° 19′ per day and therefore will make more configurations per unit time at this point in its orbit than at aphelion, where its speed is reduced to 2° 44′ per day. The variation in orbit speed of the other planets is negligible for this particular study.

Method of Observation

Each technician at the RCA Communications, Inc. Receiving Station at Riverhead, Long Island, New York, maintains a log record during his tour of duty which provides a record of conditions on shortwave signals over the North Atlantic during the midnight to 8:00 A.M. period, the 8:00 A.M. to 4:00 P.M. period and the 4:00 P.M. to midnight period. All of these eight-hour periods for the years 1942, 1944, 1947, 1948 and 1949 were correlated with existing planetary configurations and the results systematically recorded. The correlation between shortwave disturbances and planetary configurations is shown in Figure 1 for each of the years studied.

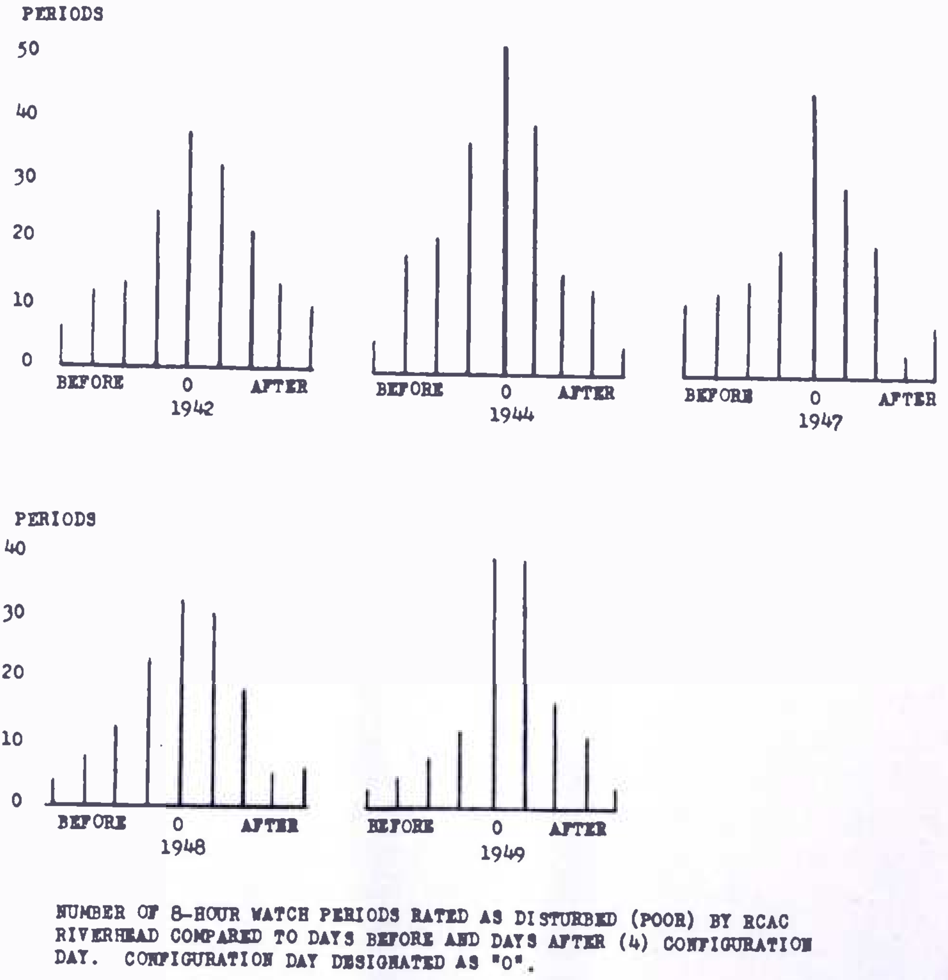


Fig. 1—Correlation between disturbed periods and day of configuration.

It can be readily seen from these graphs that disturbed conditions show good correlation with planetary configurations. This strongly indicates a relationship between North Atlantic shortwave signal behavior and planetary configurations. All configurations, however, are not accompanied by disturbed conditions. Analysis shows that some 0°, 90°, 180°, or 270° configurations between two planets can be nullified if one of them is close to 120° from another planet on the same day.

It is, however, definitely shown that each of the six planets studied is effective in some configurations.

There seems to be a "quadrant effect" in that planets over some solar quadrants show better correlation with disturbed conditions than when over other quadrants. The two most effective quadrants are shown in the study to be the first and third quadrants and the least effective the fourth quadrant. Good correlation exists with configurations of the 180° type wherein the inner planet is over the first quadrant and the outer planet over the third quadrant.

The results of a special study of Mercury when this planet was over the first quadrant and the outer planet was over the third quadrant is shown below. (The period of plus and minus one day from configuration day or a three-day period was considered in determining whether correlation was made with radio conditions.)

Planets	Number of Configurations of Above Type	Number of Configurations Disturbed
Mercury — Venus	4	4
Mercury — Mars	6	5
Mercury — Jupiter	5	3
Mercury — Saturn	4	3

The best correlation is found between shortwave disturbances and configurations of the multiple type. Analysis shows that the closer a configuration comes to being a multiple, the greater the likelihood of a disturbance. In a multiple configuration one fast inner planet will make a configuration with two slower outer planets while the two outer planets are themselves actually in configuration or close to a configuration. An example would be Mercury reaching a position where it was 90° behind Venus and 180° from Jupiter which places Venus 90° behind Jupiter. Configurations of this type, surprisingly, are quite common. Six examples are shown below.

(1)	February 7, 1944	Mercury — Venus	0°
	February 8, 1944	Mercury — Jupiter	90°
	February 8, 1944	Venus — Jupiter	90°

(February 7th to 10th were severely disturbed)

(2)	April 12, 1949	Mercury — Venus	0°
	April 12, 1949	Venus — Jupiter	90°
	April 12, 1949	Mercury — Jupiter	90°

(April 11th to 13th were severely disturbed)

(3)	January 23, 1947	Mercury — Earth	180°
	January 24, 1947	Mercury — Saturn	180°
	January 26, 1947	Earth — Saturn	0°
	(January 25th was severely disturbed)		
(4)	May 14, 1947	Earth — Jupiter	0°
	May 15, 1947	Mercury — Jupiter	180°
	May 15, 1947	Mercury — Earth	180°
	(May 13th to 17th were severely disturbed)		
(5)	February 22, 1948	Venus — Jupiter	180°
	February 23, 1948	Mercury — Venus	90°
	February 23, 1948	Mercury — Jupiter	270°
	(February 23rd to 25th were severely disturbed)		
(6)	April 18, 1948	Mercury — Venus	180°
	April 19, 1948	Mercury — Jupiter	90°
	April 21, 1948	Venus — Jupiter	270°
	(April 19th to 23rd were severely disturbed)		

The relationship between the positions of Jupiter and Saturn is very important in respect to multiple configurations and during those years when these two planets are separated by 0°, 90°, 180°, and 270°, there will be a greater number of multiple configurations since these planets move very slowly. Hence, the faster inner planets will make a double configuration in rapid time sequence every time one of them makes a contact with either Jupiter or Saturn. This is particularly so in the case of Mercury or Venus.

Correlation of Observations

The encouraging correlation found between ionospheric disturbances over the North Atlantic and configurations (particularly of the multiple type) for 1942, 1944, 1947, 1948, 1949, suggest the following deductions:

(1) That the most disturbed twelve-month periods will be those preceding and following configurations of the 0°, 90°, 180°, and 270° type between Saturn and Jupiter.

(2) That the most disturbed parts of the periods in (1) will be those in which Mars is close to a configuration of the 0°, 90°, 180°, and 270° type with either Saturn or Jupiter.

(3) That the most disturbed part of the periods in (2) will be weeks when Earth, Venus, or Mercury has a configuration of

the 0°, 90°, 180°, or 270° type with either Saturn, Jupiter, or Mars.

(4) That the most severe disturbances of all will come when the combined influence of Mars, Earth, Venus, and Mercury are such that all four will be arranged in positions where there will be a great concentration of planetary influence near the 0°, 90°, 180°, or 270° points of the Saturn-Jupiter team during the configurations mentioned in (1).

(5) That the least disturbed periods will be those preceding and following periods when Saturn and Jupiter are separated by 120°, the principal disturbances during these periods coming from configurations of the 0°, 90°, 180°, or 270° type that the inner planets Mars, Earth, Venus, and Mercury make among themselves, or as a multiple with either Saturn or Jupiter.

(6) That the least disturbed periods of all will be those when Saturn, Jupiter, and Mars are equally spaced by 120°, the principal disturbances during these periods coming from configurations that Earth, Venus, and Mercury make among themselves, or as multiples with Saturn, Jupiter, or Mars. Configurations of the multiple type are less frequent during an arrangement of 120° among these three slow outer planets.

(7) That 60° relationships between planets will also tend to produce "least disturbed periods" since 60° is one half of 120°.

An exact arrangement of 120° as mentioned in (6) is rare but a very close approach to it occurred in 1934 when Jupiter was 120° behind Saturn on June 1st. During August, Mars came to the 120° position with both Jupiter and Saturn within a few days, while Jupiter and Saturn were 117° apart. Magnetic activity records show that the 1934 yearly average was the lowest recorded between 1930 and 1949.

Considerable scientific study has been devoted to the great magnetic storm and aurora of July 26th to 30th, 1946. Figure 5 shows the positions of the six planets as they were at zero Greenwich mean time July 27th, 1946. The space separation between the planets is given below.

Jupiter—Saturn	91° 45′
Jupiter—Earth	92° 55′
Saturn —Earth	175° 20′
Mars —Mercury	89° 12′

The speed of Mercury carried it past the 90° point in relation to Mars a few hours after zero and into a multiple configuration with Saturn and Jupiter on July 30th.

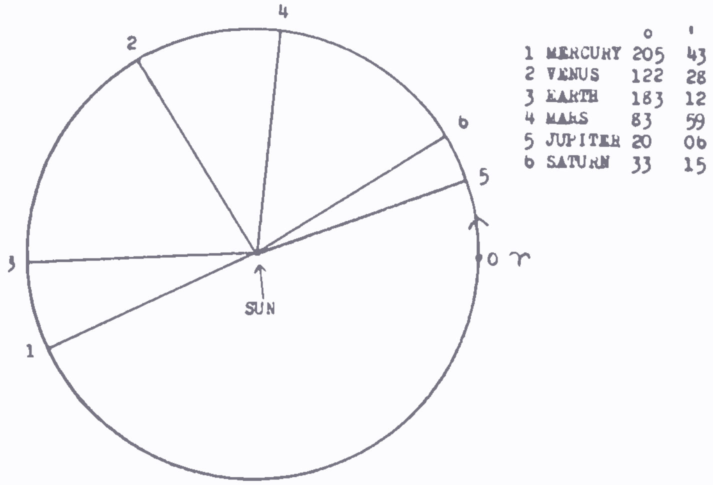


Fig. 2—Heliocentric longitude of planets during disturbances on March 24, 1940.

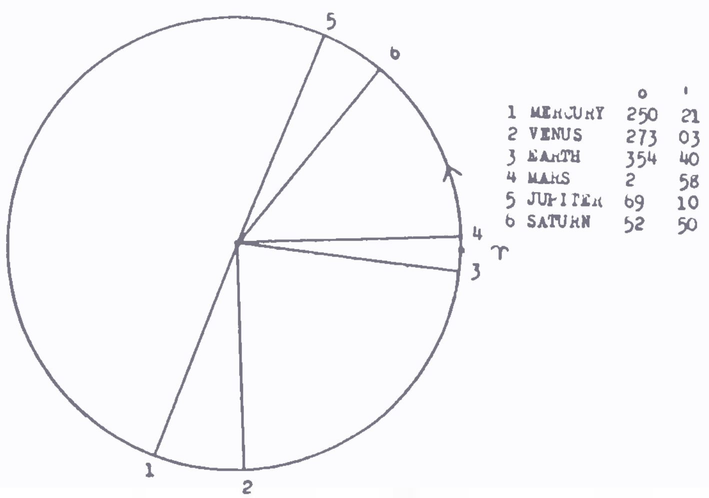


Fig. 3—Heliocentric longitude of planets during disturbances on September 18, 1941.

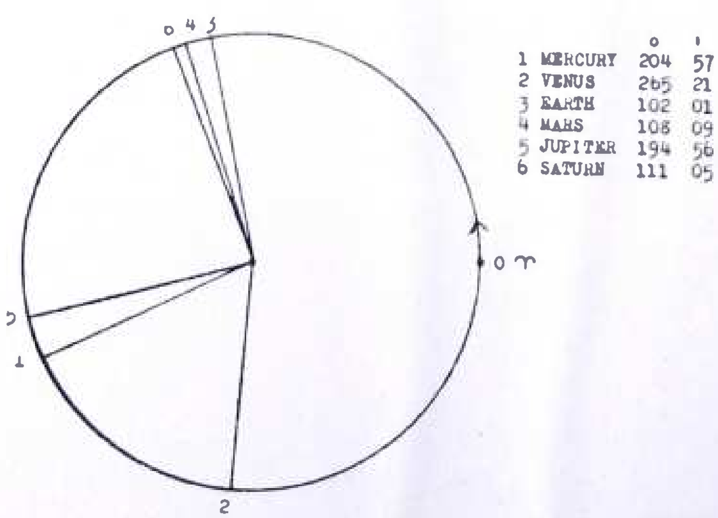
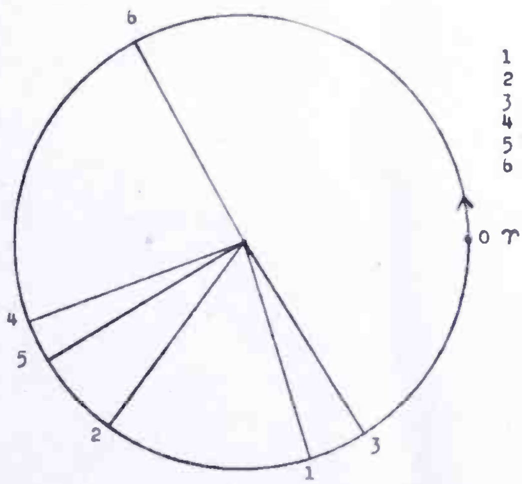


Fig. 4—Heliocentric longitude of planets during disturbances on January 3, 1946.

The heliocentric positions of the planets are shown in Figures 2, 3 and 4 for three other severe storms, March 23rd to 26th, 1940, September 18th to 20th, 1941 with aurora and January 3rd and 4th, 1946 with aurora.

Since 1946 a short term (24 hours) forecasting system has been under development at the Central Radio Office of RCA Communications, Inc. in New York City. An observatory, housing a six-inch refracting telescope, is maintained and daily solar observations, weather permitting, are made and correlated with existing radio conditions. From these solar observations, consisting of a study of the solar surface by eyepiece, and the mapping and classifying of all sunspots, a forecast for the following twenty-four hours is made. These forecasts when compared to actual radio conditions have been attaining an accuracy of around 80 per cent, as reported by RCA Communications, Inc., Riverhead, New York and Radio Suisse, Berne, Switzerland.



		°	'
1	MERCURY	288	45
2	VENUS	234	03
3	EARTH	303	21
4	MARS	199	33
5	JUPITER	210	26
6	SATURN	118	41

Fig. 5—Heliocentric longitude of planets during disturbances on July 27, 1946.

Conclusion

The research conducted at this Observatory since 1946 has quite definitely indicated that sunspots themselves are not the full answer to the problems that are manifest. There is very strong evidence that some other forces are at work in addition to the sunspots. The need of a new approach is indicated. The study of the planets as a new approach to propagation analysis has netted the encouraging results that are given in this paper and shows sufficient promise to warrant further and deeper study. A highly developed forecasting technique of this type would enable forecasting to be done several years ahead since advance planetary phenomena can be calculated with very great accuracy.

ACKNOWLEDGMENTS

Grateful acknowledgment is made to H. C. Ingles, President of RCA Communications, Inc., for his approval, interest, and encouragement given the author throughout the project. Appreciation is expressed for the valuable daily radio propagation reports received from: the Riverhead staff, in particular, A. T. Ellwood; Compagnie Radio-France, Paris, France; and Direction Generale des Telegraphs de Suede, Stockholm, Sweden. Valuable assistance is acknowledged to Carl W. Nelson, Gloucester, Massachusetts, in the determination of the "least disturbing angles".

AN AUTOMATIC NONLINEAR DISTORTION ANALYZER*

BY

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Summary—*This paper describes a system for automatically recording the nonlinear distortion frequency characteristic of a loud speaker. The system consists of the standard automatic recorder for obtaining the response frequency characteristic, coupled with a system of electronically switched high-pass filters which suppress the fundamental, and pass the harmonics for obtaining the distortion frequency characteristic. Two characteristics are obtained on the same graph sheet, namely, the response frequency characteristic and the distortion frequency characteristic. The vertical displacement between the two characteristics thus obtained gives the amount of distortion at any frequency.*

INTRODUCTION

THE nonlinear distortion characteristic of a loud speaker is a plot of the total distortion in per cent of, or in decibels below, the fundamental, versus the frequency at a specified power input. Although there are several good instruments commercially available for a "point by point" distortion analysis, the length of time required for such an analysis coupled with the possibility of overlooking narrow frequency bands of high distortion makes this type of equipment impractical for distortion analysis of loud speakers. If the nonlinear distortion characteristic of loud speakers is to be examined on a basis comparable to that with which the frequency response of a loud speaker is obtained today, some sort of automatic equipment is required for obtaining the nonlinear distortion characteristic.

In general, loud speakers exhibit unsymmetrical even harmonic and symmetrical odd harmonic distortions. However, components above the second and third are of such small magnitude as to be of secondary importance. Although it may be useful to know whether the principal contribution to the overall distortion is due to second or third harmonic distortion, it appears that for most loud-speaker research and development work, an automatic recording system which depicts the total distortion produced by a loud speaker would be very useful. It is to

* Decimal Classification: R265.2.

this end that a reliable system has been developed, designed and built, by means of which the total root-mean-square distortion is automatically traced on the same graph sheet with the response, but vertically displaced on the logarithmic ordinate scale with respect to a conventional response frequency characteristic. It is the purpose of this paper to describe this system for obtaining the nonlinear distortion frequency characteristic of a loud speaker or any other transducer.

THEORY

The automatic nonlinear distortion analyzer consists of the conventional system for obtaining a response frequency characteristic of a loud speaker coupled with an automatic means for suppressing the fundamental frequency (Figure 1). The loud speaker is supplied by a pure tone from a low distortion oscillator and power amplifier combination. The sound output of the loud speaker is picked up by a calibrated microphone. Both the loud speaker and the microphone are located in a free-field room. The output of the microphone is amplified

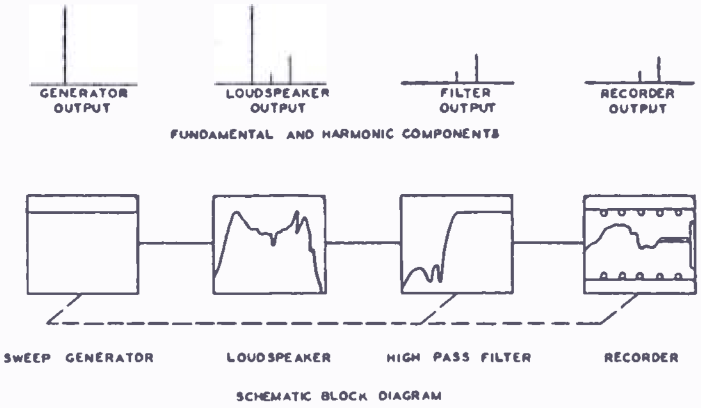

Fig. 1 — Functional diagram of a nonlinear distortion analyzer.

and fed to a recorder and a response frequency characteristic of the loud speaker is obtained with this system in the conventional manner. To obtain the distortion frequency characteristic which depicts the distortion generated by the loud speaker as a function of the frequency, the system for automatically suppressing the fundamental is connected between the microphone, amplifier and the recorder. Under these conditions, the voltage applied to the recorder is the root-mean-square total of the harmonic frequencies generated by the loud speaker.

In the system described above, the major problem becomes one of attenuating the fundamental frequency in a dependable and fairly rapid manner. Of the several methods available for eliminating the fundamental, a reliable and straightforward one is shown in Figure 2. This consists of a series of high-pass filters sequentially interposed between the microphone pickup and the recording equipment to attenuate the 40 to 15,000 cycle sweep fundamental. The primary advantage of using this method for harmonic distortion measurements is its dependability. The filters themselves may be made very rugged. Furthermore, should the filter switching circuit fail to function properly, the distortion readings will immediately go to 100 per cent, thus reading fundamental rather than harmonics, and thereby providing a positive check against a possible switching error.

DESIGN

The design of the system described in the preceding section is not difficult. However, several practical problems are involved. The useful frequency range of each filter is determined by two frequencies, namely, f_c and f_∞. The frequency at which the response is down one decibel is f_c. This frequency sets the lower limit of the useful pass band of each

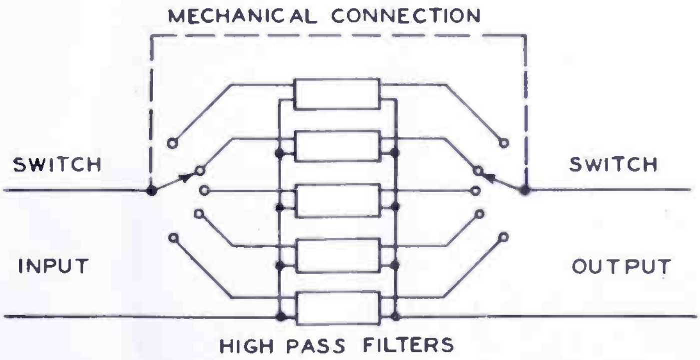

Fig. 2—A nonlinear distortion analyzer consisting of a series of high pass filters which can be switched to suppress the fundamental and pass the harmonics.

high-pass filter when recording distortion to an accuracy of 10 per cent. The frequency at which the response is down sixty decibels is f_∞. This frequency sets the upper limit of the useful rejection band of each filter when recording distortion to an accuracy of 10 per cent for a 0.3 per cent 2nd harmonic distortion value. The f_c and f_∞ overlap characteristic of adjacent filters are very close at the lower frequencies, and it therefore becomes very important that the frequency at which a filter is switched be held to a close frequency tolerance if the full possible accuracy of the distortion analysis is to be realized. For this reason, an electronic rather than a mechanical system for the detection of the switch frequency is used. A typical bridged-T network employed for switch frequency detection is shown in Figure 3. The response frequency characteristic of the network is also shown in Figure 3. The inductors, capacitors and resistors of the network are sufficiently stable so that frequency adjustments are infrequent. In order to make up for the difference in amplitude between the fundamental and the harmonic levels normally encountered, a microphone preamplifier is necessary. It is important that this amplifier generate a very low level of harmonics, and be essentially noise-free due to its position in the system. The network of the preamplifier, the high-pass filters, and the automatic switch frequency detection and the switching together make up a system which, when used with any conventional automatic oscillator-recorder system, provides a recording instrument for total root-mean-square distortion analysis.

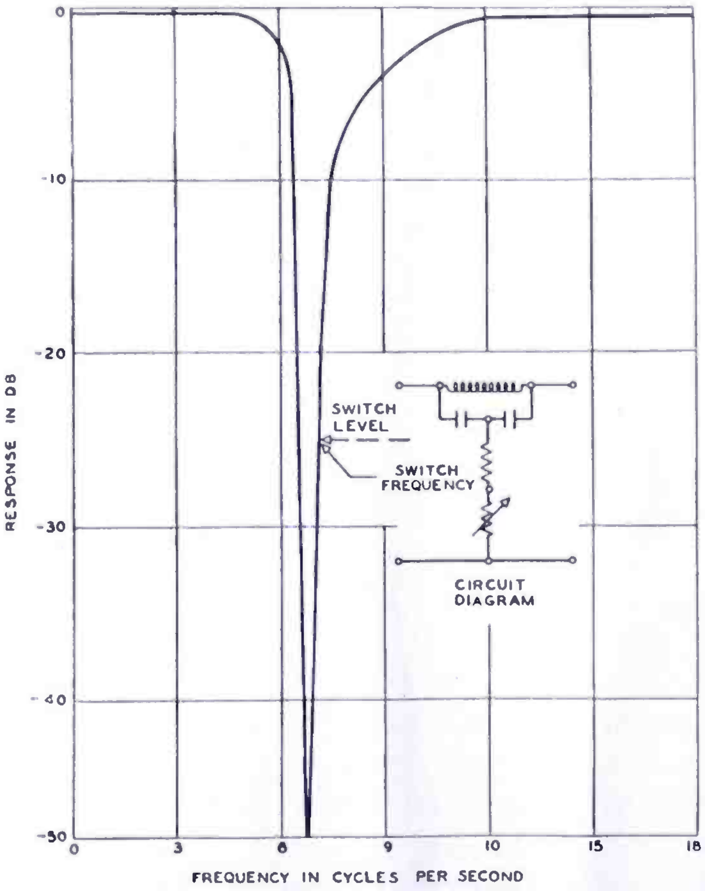

Fig. 3 — Bridged-T network used for initiating the switching of the filters. The graph depicts the response characteristics of the network.

A functional block diagram of the complete analyzer recorder system is shown in Figure 4. The oscillator, power amplifier, microphone amplifier and recorder are located in permanent racks. A photograph of the recording equipment is shown in Figure 5. The distortion analyzer is located in a portable cabinet. A photograph of the distortion analyzer is shown in Figure 6.

The sweep oscillator, an RCA type 68-B beat-frequency oscillator of exceptionally low harmonic content, is rotated, through a mechanical linkage, by a Leeds and Northrup Speedomax recorder at a sweep speed of 2½ minutes for 30 to 17,000 cycles. The output of the oscillator is fed to both the amplifier which drives the loud speaker and the preamplifier of the distortion analyzer. The output level of the amplifier

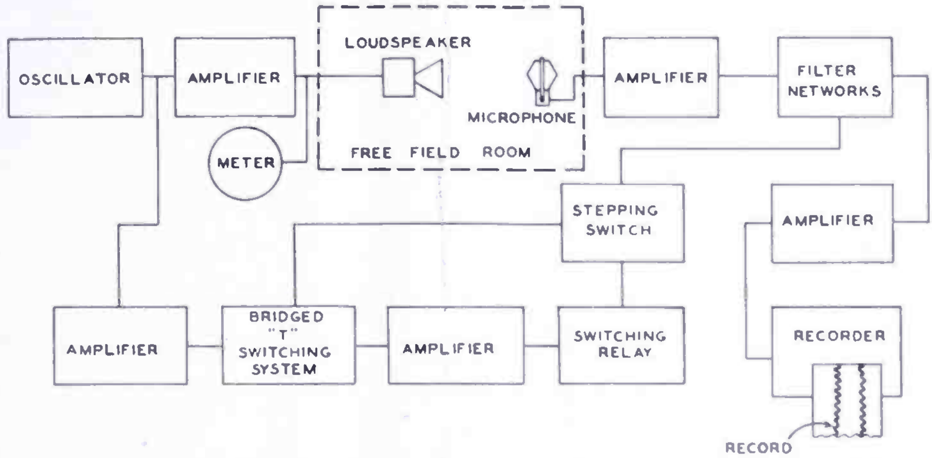

Fig. 4—A complete block diagram of the automatic nonlinear distortion recording system.

which drives the loud speaker is shown on the output meter. The output of the preamplifier of the distortion analyzer is fed to the bridged-T switching network. The sound output produced by the loud speaker is picked up by an RCA 44B velocity microphone and amplified and fed to the filter system of the distortion analyzer. The inputs and

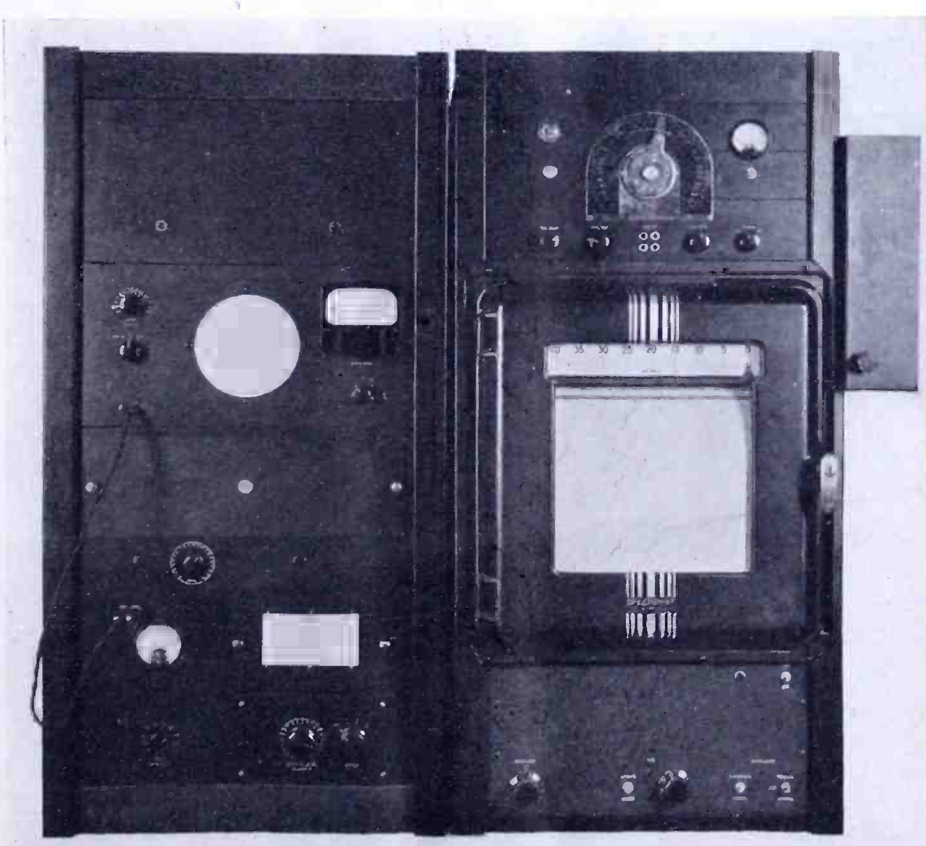

Fig. 5—An automatic response frequency recorder.

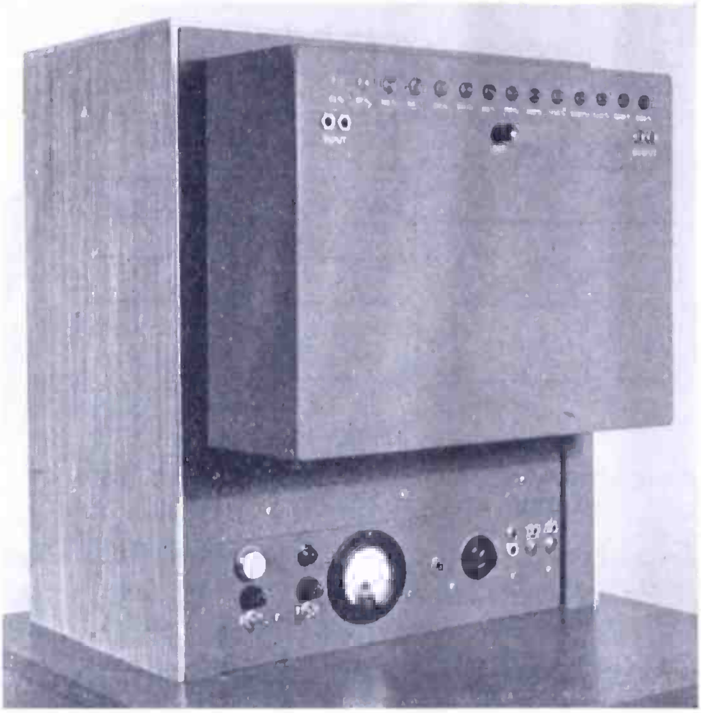

Fig. 6 — The nonlinear distortion analyzer.

outputs of the fourteen low-pass filters are connected to two banks of a relay-operated step switch. The step relay is operated from the output of the bridged-T switching network. Considerable care was taken so the filters met the f_o and f_∞ requirements. The f_o/f_∞ ratio was eventually fixed at 4:3 which gives a useful frequency range of ½ octave per filter or 14 filters for the 40 to 15,000 cycle audio range. Figure 7 shows the response frequency characteristics of 14 high pass

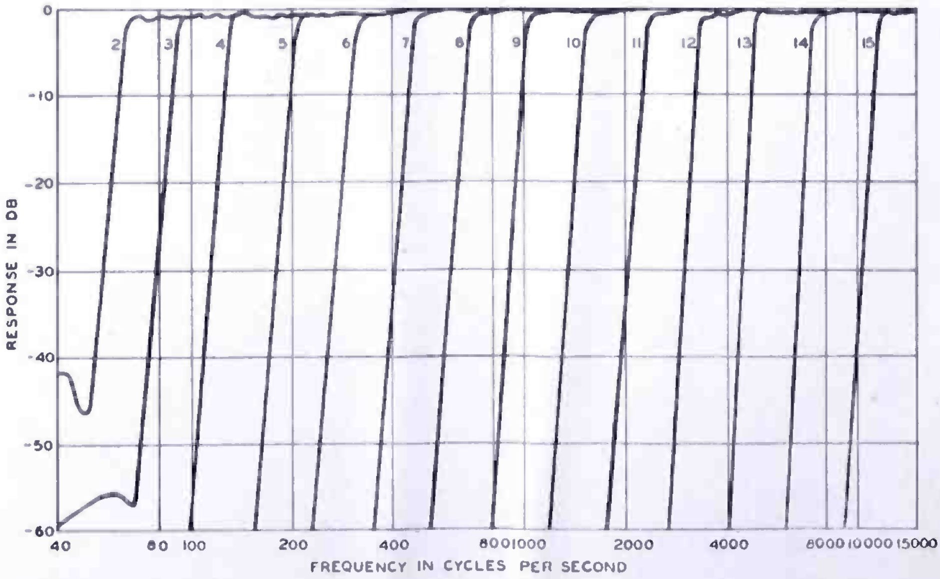

Fig. 7—The response frequency characteristics of the fourteen high-pass filters used in the nonlinear distortion.

filters used in the system. Filters 2, 3, 4 and 5 have respectively 40, 45, 50 and 55 decibels attenuation for f_∞. The remainder have over 60 decibels attenuation for f_∞. The filters are 600-ohm, single-ended, M-derived, torroidal-core networks. The insertion loss, which varies from filter to filter, was equalized by suitable resistance pads across the output terminals. Filters 2 and 3 were heavily shielded to reduce hum pickup. A series of panel lights connected to the third bank of the step switch indicates the particular filter in operation at any time. Electronic frequency detection of the proper switch frequency and the switching operation occurs as follows. A constant amplitude signal supplied by the oscillator is fed to the preamplifier on the distortion analyzer panel. The amplified signal is then fed to the frequency selective network as shown in Figure 4. Fourteen of these selective networks, one for each high-pass filter, are connected in sequence to the fourth bank of the step switch. As the oscillator sweeps through the null frequency of the particular bridged-T network in the circuit, the switching relay closes, and thereby energizes the solenoid of the step relay. This solenoid action sets the step switch. As the oscillator frequency increases, the relay is de-energized, and the step switch advances thereby placing in the circuit both the next bridged-T switch network and the next high-pass filter. This process is repeated for each network until the 14 bridged-T switch networks and the 14 high-pass fundamental removing filters have been used. The output of each filter contains the harmonics and the highly attenuated fundamental. The output of each filter in turn is fed to an amplifier and the recorder. The recorder has a range of 0 to 40 decibels above a zero input level of 0.005 volt, with a zero range adjustable ±5 decibels. The recorder is designed to operate between 40 and 15,000 cycles to an accuracy of one-half per cent of full scale, and has a response speed such that the pen will traverse any part of the nominal ten-inch chart and arrive at complete balance without overshooting in two seconds.

For a loud-speaker distortion analysis, a conventional amplitude versus frequency response curve is first run with the power amplifier adjusted to furnish the proper power level to the loud speaker under test, and with the distortion analyzer step switch solenoid power turned off. The recorder preamplifier is adjusted to a level such that the recorder will not go off scale for the response curve. The response frequency characteristic is run with this gain setting. This procedure is repeated with the distortion analyzer step switch solenoid power turned on and with the gain control of the preamplifier turned some 20 to 40 decibels higher. The resultant curve, with due consideration

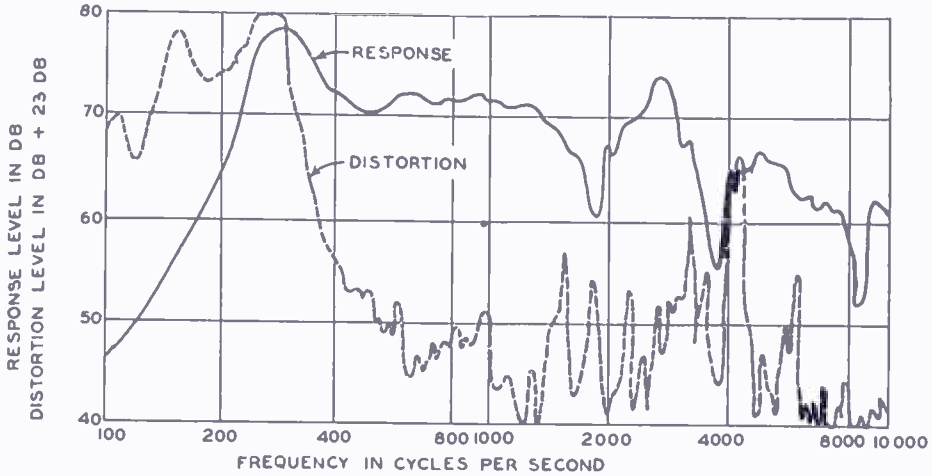

Fig. 8—The response frequency characteristic and the distortion frequency characteristic of a three-inch direct radiator loud speaker for an input of .1 watt. Note that the level of the distortion curve has been raised 23 decibels.

for the difference in preamplifier settings, gives the relative distortion frequency characteristic of the loud speaker under test.

A response frequency characteristic and a distortion frequency

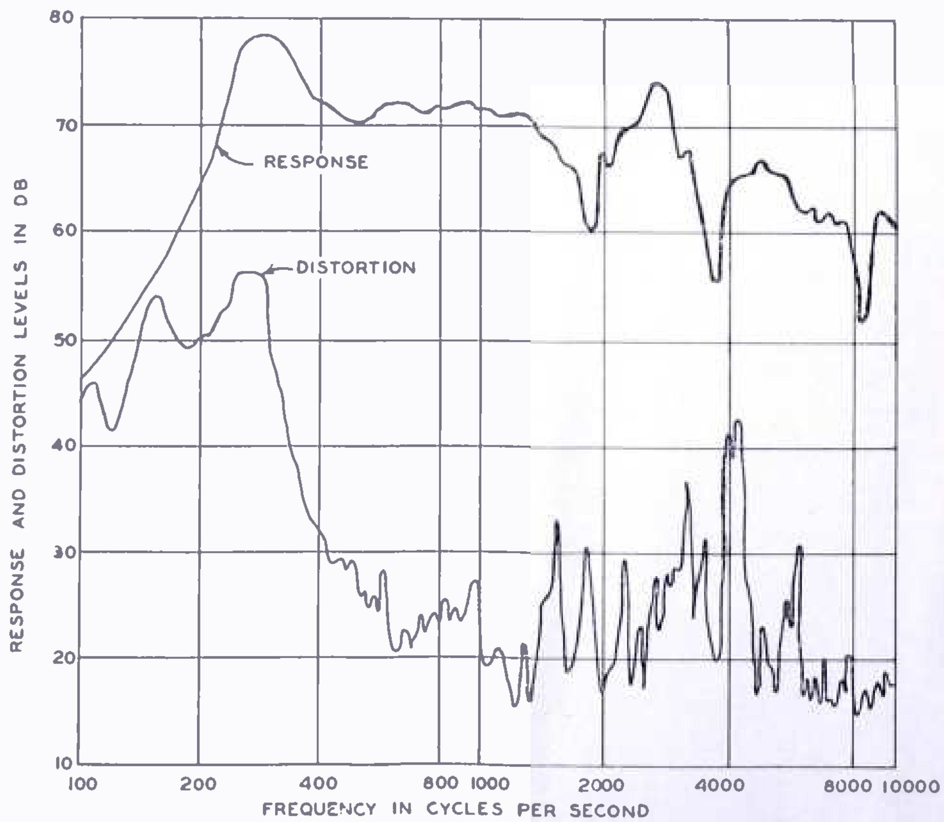

Fig. 9—The response frequency characteristic and the distortion frequency characteristic of a three-inch direct radiator loud speaker for an input of .1 watt. Data obtained from Figure 8.

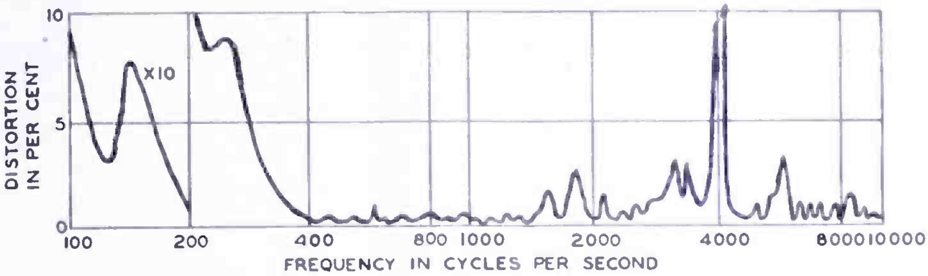

Fig. 10—Distortion frequency characteristic of a three-inch direct radiator loud speaker in per cent of the fundamental. Data obtained from Figure 8.

characteristic taken on a three-inch dynamic loud speaker for an input of 0.1 watt is shown in Figure 8. The gain of the preamplifier was increased 23 decibels for the distortion frequency characteristic. The response frequency characteristic and the distortion frequency characteristic drawn in the true relation on the same graph are shown in Figure 9. The distortion in per cent of the fundamental as a function of the frequency obtained from the data of Figures 8 and 9 is shown in Figure 10. It will be seen that the distortion is relatively large. This is due to the small light-weight diaphragm used in this loud speaker to obtain high sensitivity.

A response frequency characteristic and distortion frequency characteristic taken on a twelve-inch loud speaker for 1 watt input is shown in Figure 11. The distortion in per cent of the fundamental is shown in Figure 12. It will be seen that the distortion produced by this loud speaker is very low. This loud speaker will deliver a sound level of 80 decibels in the average living room for an input of 0.1 watt. Thus it will be seen that for reproduction in the home, a loud speaker of this type exhibits less distortion than some of the other elements in the reproducing system.

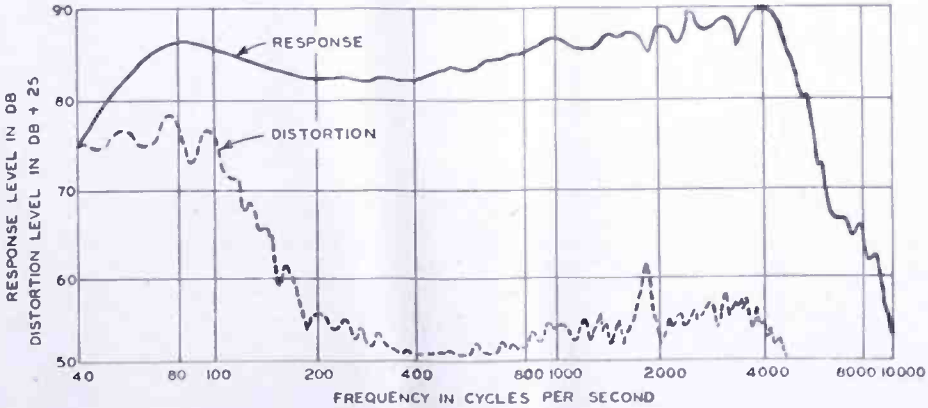

Fig. 11—The response frequency characteristic and the distortion frequency characteristic of a twelve-inch dynamic loud speaker for an input of one watt. Note that the level of the distortion curve has been raised 25 decibels.

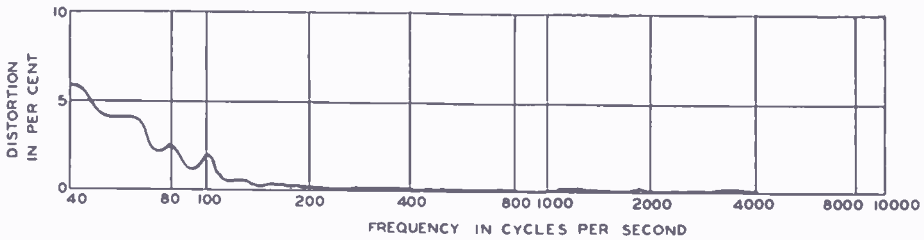

Fig. 12—Distortion frequency characteristic of a twelve-inch direct radiator loud speaker in per cent of the fundamental. Data obtained from Figure 11.

From the characteristics depicted in this paper, it is seen that the distortion frequency characteristic of a loud speaker exhibits larger variations than the response frequency characteristic. In order to obtain an accurate characteristic under these conditions and in a reasonable length of time requires some sort of automatic or semi-automatic recording means. For example, in the case of the distortion frequency characteristic of Figure 8, it would be practically impossible to obtain this curve by a point by point method. Typical distortion characteristics shown in this paper illustrate the usefulness of an automatic curve-tracing distortion analyzer for obtaining the distortion characteristics of loud speakers.

OPEN-FIELD TEST FACILITIES FOR MEASUREMENT OF INCIDENTAL RECEIVER RADIATION*

By

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Summary—In order to determine accurately the amount of incidental radiation from a television receiver, it is necessary to make actual measurements in the field. A test setup permitting such measurements is described.

INTRODUCTION

THE problem of evaluation and reduction of interference radiation propensities of television and frequency-modulation receivers is one of major import to manufacturers, and has been receiving increasing attention. Progress in this direction was initially greatly hampered by the lack of standards or facilities to permit measurement in terms of actual inductive or radiated field intensity. Some time ago the Industry Service Laboratory was asked to propose means of measurement of incidental radiation, including specification of the physical characteristics of the required open-field facilities. Upon receipt of favorable comment on the proposal the construction of such a setup was started on the grounds of RCA Laboratories Division at Princeton, N. J. Construction was completed and the facilities were placed in operation early last Summer, since which time they have been in nearly constant use. Results have been up to expectations, and it is becoming increasingly apparent that facilities of this type are essential in the evolution of receiver designs which will conform to whatever standards of good engineering practice are eventually established on incidental radiation.

CHOICE OF SITE

Due care must be exercised in choosing a site for making radiation measurements, and in constructing the test apparatus, if accurate and reproducible measurements are required. The site should be one where the ground is relatively level and of uniform electrical characteristics. It should be clear of all overhead wires or other metallic objects. The minimum distance from either the receiver under test or the field-strength meter to objects such as eaves troughs, house plumbing, etc.,

* Decimal Classification: R261.5 × R583.5.

should preferably not be less than 3 times the distance between the receiver under test and the field-strength meter. Furthermore, the site should be relatively free from unwanted signals such as ignition noise or strong radio-frequency carriers, as from locally generated oscillations or near-by frequency-modulation and television stations, in order that the wanted signal may be easily and correctly identified and accurately measured. It is also desirable, for practical reasons, that the chosen location be relatively accessible.

The grounds of RCA Laboratories Division at Princeton essentially fulfill these requirements. A site was chosen on the clear, flat lawn in the rear of the laboratory, about 1200 feet from the main laboratory building, 600 feet from the nearest building of any type, and about 200 feet from the nearest tree. A general view of the chosen site with installed facilities is shown in Figure 1.

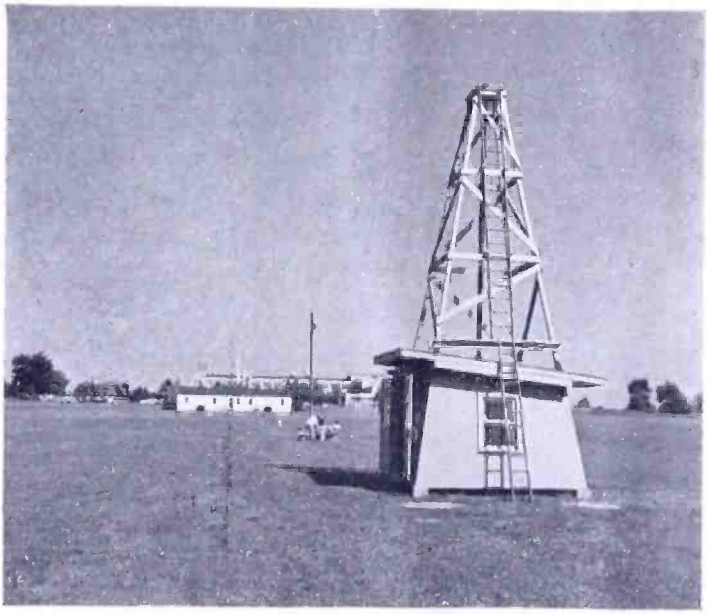

Fig. 1—General view of facilities and site.

Test Layout

The test layout consists of a receiver station (with tower, Figure 1), into which the test receiver is placed, and a set of five listening posts, spaced at distances of 100, 200, 300, 400 and 500 feet from the center of the receiver station. At any one of these listening posts the field-intensity meter and its associated equipment may be operated. The field-intensity meter is shown at the 100-foot location in Figure 1. 60-cycle power is available at all of the listening posts and at the receiver station. All construction is of either wood or plastic except for such items as nails, small braces, etc. In no case, however, did the size of the metallic materials used exceed 6 inches.

Power

60-cycle alternating-current power is supplied to the receiver station and the various listening posts by means of a two-wire trench-lay cable 600 feet long, buried throughout to a depth of twelve inches or more. The outlets at the listening posts are of the all-weather type, and are mounted 18 inches above the ground on substantial posts (see Figure 3). The power cable is totally contained in pipe conduit between the receiver-station outlet and the outlet at the first listening post. Short lengths of pipe conduit are also used as risers at the other listening-post outlets. The voltage drop in the cable (No. 4 B. and S. gauge) when supplying an average television receiver and field intensity meter is approximately 1½ volts. An automatic voltage regulator of the diode-bridge-control type is used to maintain the line voltage at 117 volts. It is located in a small auxiliary office-laboratory (left of center, Figure 1), at the input to the underground line.

Receiver Station

Functionally, the receiver station consists of a turntable upon which the test receiver is placed, an antenna mast which is rigidly attached to the turntable, and a tower which supports the mast. The lower portion of the tower is sided with exterior grade plywood and roofed over to form a small shed (11 by 11 foot floor, 7½ feet high) for protection of the test receiver and for overnight storage of measuring, servicing, and incidental gear. This tower and house structure is anchored on concrete piers at the corners. Adequate access, light, and ventilation are provided by double doors on one side, and windows on the other three.

The receiver turntable (Figure 2) is a circular wooden platform about 4½ feet in diameter. It is constructed of two sheets of one-inch plywood, with a framing structure and hub sandwiched between. The turntable rests upon twelve plastic croquet balls arranged annularly in two concentric rings. The balls are held in proper spacing by a Masonite retainer pivoted on the hub. Masonite facings are also applied to the floor and to the bottom surface of the turntable to provide smooth rolling surfaces for the balls. The receiver power-line conduit is brought up through the arbor upon which the turntable hub rotates. The alternating-current outlet box is located in the center of the turntable, but does not rotate therewith. The circumference of the turntable framing members and the plywood top and bottom sheets form a flanged pulley for a rope drive. Rotation is accomplished by means of a hand-powered rope "belt" extending to the first listening post. This permits simultaneous control of rotation and observation of its effect on field intensity.

The top of the turntable is 18 inches above ground. Console receivers, and receivers which include a stand, are placed directly on the turntable for test. For table model receivers, the 30-inch-high table shown in Figure 2 is placed on the turntable in order to raise the receiver chassis to a height of four feet above ground.

The antenna mast is carried on the cap plate of the four-legged "spider" structure based on the turntable (Figure 2). Thus, the mast and turntable rotate as a unit. Bearings are provided for the mast at roof level and at the top of the tower. These bearings are accurately plumbed with the turntable bottom pivot to avoid binding.

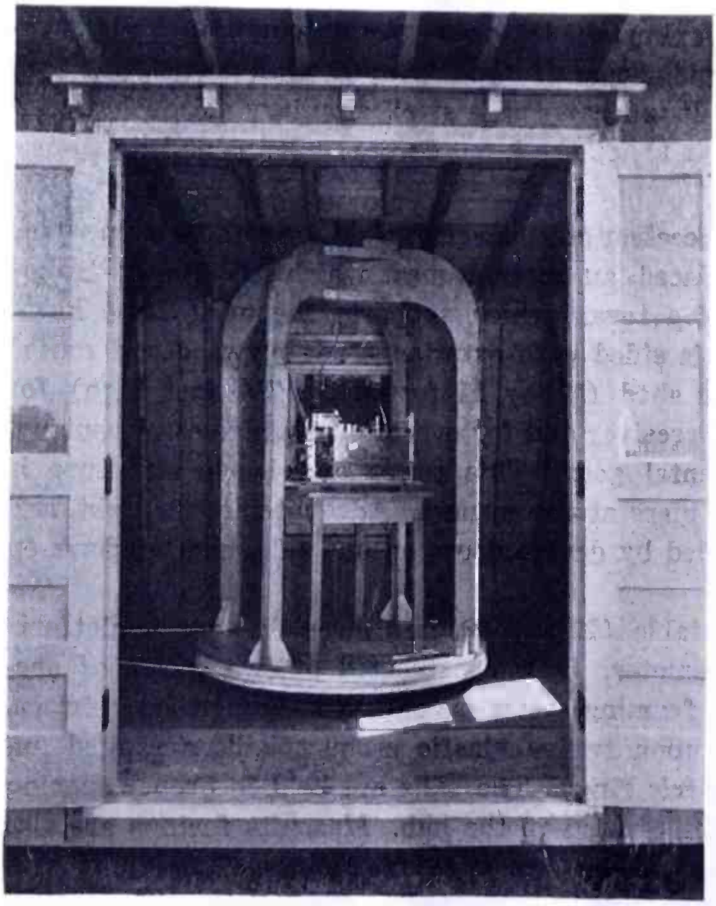

Fig. 2—Receiver station.

The antenna lead-in is 28 feet long and of whatever type is specified by the receiver manufacturer, or standard 300-ohm ribbon line if no specification is made. The lead-in is suspended down through the center of the antenna mast. The antenna elements are clamped to a bakelite cap which weather-proofs the top opening. The antenna is at an elevation of 30 feet above ground. Present practice is to use a single dipole, 88 inches long for television receivers, and 58 inches for frequency modulation receivers. The dipoles are made of ½-inch diameter aluminum tubing. A ladder attached to the side of the tower provides access for changes or removal of antenna and lead-in.

Listening Posts

The field intensity of receiver radiation or induction fields can be measured at any of the five listening posts, using any suitable field-intensity meter or the equivalent. For example, a Stoddart NMA-5 meter is being used for very-high-frequency (VHF) measurements, and a Ferris 32-A in the range of 150 kilocycles to 20 megacycles.

As may be seen in Figure 3, the VHF field intensity meter antenna and its control means are mounted on a plywood mast, which is ap-

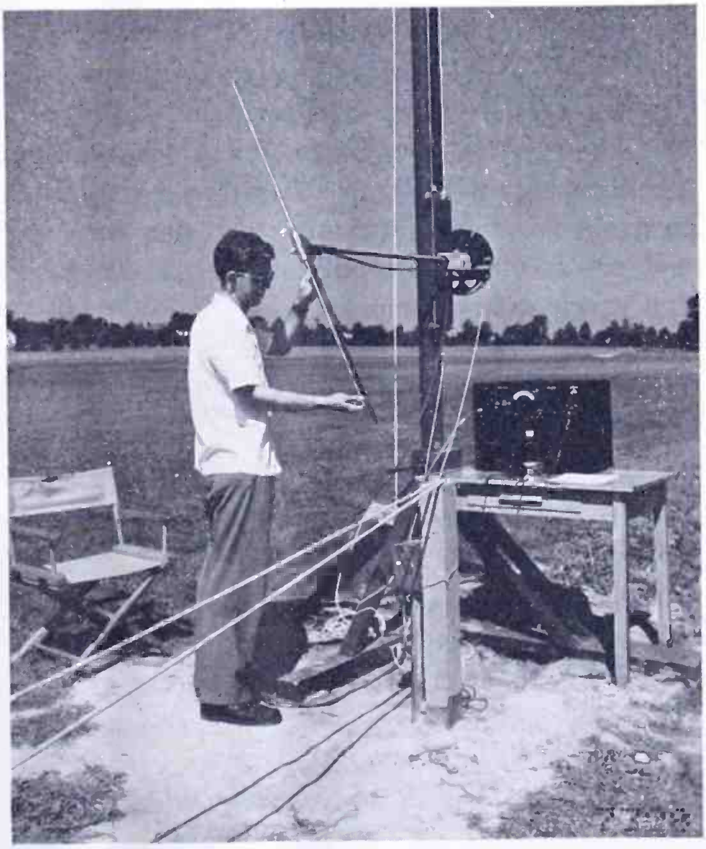

Fig. 3—Listening post.

proximately 22 feet tall. In order to provide a standardized clearance between the dipole and its transmission line lead-in when the dipole is oriented for vertically polarized signals, and to minimize mechanical stress on the line at the antenna plug, the dipole itself is horizontally displaced an arbitrarily selected distance of two feet from the mast on the end of a bakelite pole. The transmission line is suspended horizontally along this pole before being dropped vertically down to the field intensity meter. The antenna can be rotated through a vertical plane from the ground by means of a rope over a grooved pulley on the horizontal pole. The rope has a point of attachment to the pulley to prevent slippage, and a 1½ turn wrap to permit complete once-around rotation of the antenna. Detents accurately locate the horizontal and vertical antenna positions. The bearing block for the horizontal pole is mounted on a roller-equipped carriage on a bakelite track up the side of the mast. The height of the carriage is controllable by means of a rope over a pulley on top of the mast. Thus, the antenna can be easily lowered for length adjustment to suit the signal frequency, and raised for field intensity measurement to any selected height between five and twenty feet. A venetian blind type of rope latch holds the antenna at the selected height, and is an effective safeguard against accidental dropping of the antenna assembly.

The VHF field intensity meter antenna is aimed broadside at the receiver test location. The mast is held erect in its stand by means of a hinge pin at the bottom, and a locking pin at the top of the stand. The locking pin is removed and the mast lowered when not in service to avoid possible overturning and damage in windstorms. The mast and accessories are readily transportable piecemeal when the stand and carriage are removed.

Sources of Incidental Radiation

The source of incidental radiation which thus far has given greatest concern in frequency-modulation and television receivers is the frequency conversion oscillator. Other sources must be considered also, however, particularly in television receivers. These include:

1. The kinescope high-voltage supply, whether of the radio-frequency oscillator or kick-back type.
2. The horizontal sweep, which is prolific of harmonics of the sweep frequency, 15.75 kilocycles, and operates at a high energy level.
3. The intermediate-frequency video, and intercarrier-sound amplifiers. A 4.5-megacycle intercarrier beat is generated whether used for sound or not.

Induction and radiation fields of significant magnitude may be set up by any of the foregoing sources via the antenna and lead-in, the chassis and components, the power cord, or any combination thereof.

Measurement Procedure

VHF Measurements

In preparation, all surplus metallic objects, such as tools and servicing equipment, are removed from the measurement area. The test receiver is placed in the center of the turntable so that its front panel occupies a vertical plane parallel to the external antenna at the top of the tower. The line cord is plugged into the power outlet in the center of the turntable, any excess cord being bundled up. The field intensity meter antenna is adjusted to the length proper for the frequency of measurement, and raised to the selected height (ordinarily 20 feet) above ground. The field intensity meter is tuned to the signal to be measured, due care being taken that this signal is correctly identified and adequately interference-free. In the case of local oscillator radiation measurements, interfering signals are avoided by suitable adjustment of the receiver fine-tuning control. The field intensity meter is tuned to a "clear" frequency as near as possible to the frequency to be measured. The receiver is then tuned, using a ten-power target-spotting telescope for observation of the field intensity indicating meter. The field intensity meter indication is then maximized by rotating the receiver and its antenna by means of the turntable remote control rope.

Field intensity values are measured for horizontal polarization using each of the two possible polarities of the receiver's antenna lead. Vertical polarization field intensity is taken for one polarity only, since the vertical field is not materially affected by the polarity of the receiver antenna connections. Present practice is to make conversion oscillator radiation measurements as follows:

1. With the external antenna and lead-in connected.
2. With a dummy antenna having a resistance value equal to the characteristic impedance of the antenna transmission line specified by the receiver manufacturer for use with the receiver, or with a 300-ohm dummy antenna in the absence of such specification. (The external antenna and lead-in are removed from the measurement vicinity for measurements 2 and 3.)
3. Occasionally measurements are also made with the built-in antenna connected, when the receiver is so equipped.

Low- and Medium-Frequency Measurements

Use of the open-field test installation in the low- and medium-frequency ranges thus far has been limited mostly to small-scale experimental investigation. Measurements have been confined to vertical polarization orientation of the field intensity meter antenna, which is legitimate to at least the high-frequency end of the broadcast band. In both the aviation beacon band and broadcast band, the problem of avoiding interference from extraneous signals is excruciatingly severe when attempting to survey a virtually continuous spectrum of signals such as is generated by a television receiver sweep and kick-back high-voltage-supply circuit. While some relief is afforded by conducting tests during off hours, the aviation beacon channels, and many of the broadcast channels, are occupied on an around-the-clock basis.

Consideration must also be given the type of antenna used for pick-up in conjunction with the field intensity meter, since the 100-foot listening-post position, which may be considered typically spaced as regards a practical broadcast-band interference condition, is in a predominantly induced rather than radiated field at frequencies lower than about 1500 kilocycles.

Results to date indicate the desirability, if not necessity, of using a self-contained, battery-operated field intensity meter in the low- and medium-frequency ranges in order to circumvent power-line coupling. Much work remains to be done before a satisfactory operational procedure can be said to have been established for this frequency spectrum.

THE SELECTIVE ELECTROSTATIC STORAGE TUBE*

By

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Summary—*This paper describes an electrostatic storage tube developed for high speed registry and read-out of digital information consisting of bivalued signals. It has storage capacity for 256 signals.*

The selection of the internal address of the stored signals is by means of a fixed matrix formed by two orthogonal sets of spaced parallel bars. A uniform electron emission, available in all windows of the matrix, is suppressed everywhere except in a particular window, by applying appropriate control voltages to the bars. A novel system of connections between the bars allows control of many windows by relatively few external leads. Bivalued address signals are applied to these leads. This does not require circuits having an accurate amplitude response.

The storage is obtained by the two stable potentials which tiny floating metallic elements, located in register with the windows, assume under continuous electron bombardment. The signal to be stored is applied by capacitive coupling to all elements and brings the selected one to the desired stable potential. The reading signals are sizeable electron currents passing through a hole in the storing elements under the control of the element's potential. A visual display of the stored information is obtained also.

The tube has many ideal characteristics: indefinitely long storage time; random access to any element for writing and reading in a few microseconds; no erasure needed before registry; and possibility of indefinitely repeated readings from any element.

The characteristics of the tube and requirements of the associated circuits are given. A theory of the connections of parallel bars for combinatorial selection is included.

INTRODUCTION

Some ten years ago electronic circuits were developed to perform arithmetical computations accurately at tremendous speeds by operating on bivalued or "on-off" signals which express the numbers in digital form, either in the binary or some coded decimal numeration system. It became possible to perform extremely long sequences of accurate computations in short enough times to attack by numerical methods many scientific, technical and military problems which could not, in practice, be solved by former methods. Several successful machines have already been built in this country and abroad, and many are under construction. In fact, electronic computation has been in-

* Decimal Classification: 621.375.2.

troduced to business machines and, to believe some prophets, is likely to produce a revolution in clerical work.

The inner memory of an electronic digital computing machine is a storage device into which bivalued signals can be registered in a very short time, stored for an indefinite period and read off on very short notice. It is the essential component required for automatic operation because numbers resulting from one computation can be stored to serve as a basis for a subsequent one, and thereby make possible long sequences without undue waste of time for handling intermediary results. The memory can also store coded instructions describing the arithmetical operations to be performed for a particular problem and enable the machine to execute them in a fixed sequence, or one which is conditioned by the progress of the computation. In this way a machine consisting of an input device, an inner memory, a control, a single arithmetical unit and an output device, is essentially universal because it can solve a great variety of problems by merely inserting a proper program of instructions in it. These instructions, or orders, will contain in general the location or address within the memory of the numbers to be operated on, the desired operation, and the address to which the result is to be stored.

The usefulness of a computing machine depends, evidently, on the storage capacity of its memory, as this limits the size of the program and the number of possible intermediary results. Access to the desired address of the storage should take a time shorter than, or comparable to, that taken by the most frequent arithmetical operation in order to avoid wasting most of the time in mere "filings" and "look-ups." The requirements of large capacity and short access time are mutually contrary in storage devices in which access to all bits of information is in time series, such as in a moving magnetic tape, rotating drum or sonic delay line. With area (or volume) storage, this is not generally the case as it is possible to have direct access to any point without motion of the entire pattern of information. In short, the essential component of an information handling machine is a large capacity memory with rapid random access to its elements.

The storage of electric charges on a multitude of small areas of a surface has been utilized for years in television pickup tubes, such as the iconoscope, and other beam deflected storage tubes and is one of the fastest storage systems known. The electrostatic storage tube described in this paper is based on a novel conception of area selection. It is by the means of a fixed matrix formed by two orthogonal sets of parallel wires rather than by the deflection of a beam. Electrons bombard the matrix uniformly and are stopped in all windows except a particular one, as illustrated in Figure 1. This is accomplished by applying proper control voltages to the selecting bars. It is obvious that such a go, no-go, control of positions materially fixed by the matrix provides a greater certainty of selection than is possible by controlling precisely the deflection of a beam. Moreover, the bivalued signals coding the address of the information are directly convertible to the control voltages of the tube without resorting to accurate amplitude conversion, unnatural to digital computation which deals with quantized quantities only. These advantages are obtained by the use of a large number of control bars which would be impractical if it were not for the possibility of connecting them into a relatively small number of groups. This possibility arises from the fact that a

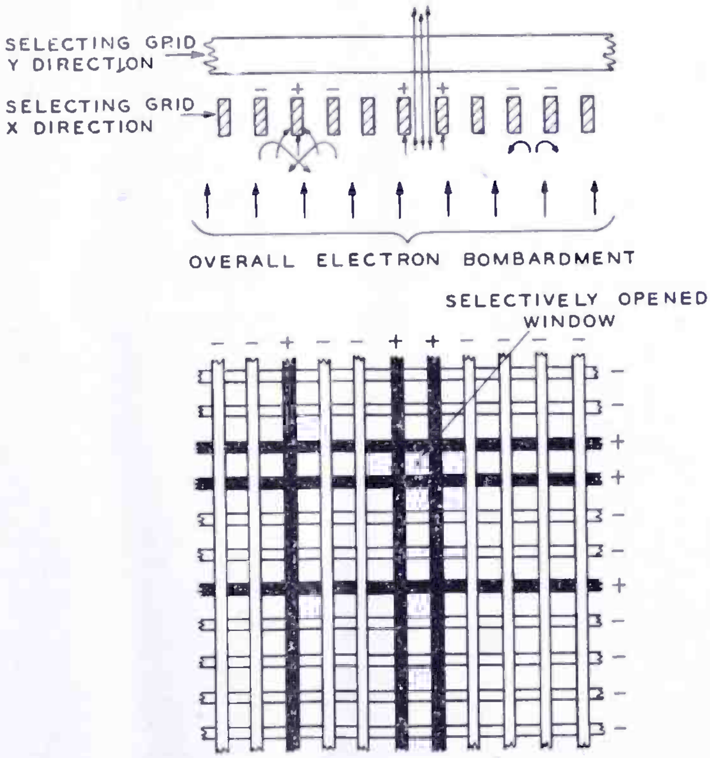

Fig. 1—Principle of selection.

quadruple coincidence of positive voltages on the bars making up a window is necessary to let electrons pass through that window. The number of elements is, therefore, proportional to the 4th power of the number of controlling leads. The resulting economy of control leads may be appreciated by considering that about a million elements can be controlled by one hundred leads.

The storage mechanism, like the selection, is inherently quantized in this tube. The storage is obtained by the two naturally stable potentials which electrically floating metallic storing elements assume under constant electron bombardment: the potential at which electrons are repelled from the element and one of the potentials for which the bombarding primary electrons produce an equal number of secondaries. The storage time is indefinite, because the constant current holds forcefully the equilibrium potential in spite of the leakage due to the imperfections of the insulation of the element's supports. The signals to be stored are applied by capacitive coupling to all elements and bring the selected ones to the desired stable potentials. The reading currents are sizeable electron currents obtained by the "grid-action" of the storing elements. These currents are either present or absent, so that readings are bivalued also and require no amplitude discrimination. These currents produce incidentally a monitoring visual display.

The selection principle and early results of the research work were described in several patents[1] and reported by the author at several meetings[2,3] and in one brief published note.[5] Later results were briefly reported at several conferences.[6,8] Several geometrical arrangements with larger storage capacities were tried at first. The present paper is confined to the description of the latest tube with a capacity of 256 elements which has been developed to practical usefulness. It includes, in the appendix, a theory of connections of parallel bars for combinatorial selection, as this aspect of the tube is the most novel.

It is believed that this tube — the only truly random access storage device operating with bivalued inputs and outputs — will be very useful for the inner memory of computing machines. It may also have

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[1] U. S. Patents 2,442,985, 2,494,670 and 2,519,172. Other patents pending.

[2] Jan Rajchman, "The Selectron — A Tube for Selective Electrostatic Storage", Symposium for Large Scale Computing Machinery, Harvard University, January 8, 1947. The Proceedings of this Symposium have been published by the Harvard University Press.

[3] Jan Rajchman, "The Selectron — A Tube for Selective Electrostatic Storage". Several talks were made on this subject but not published: Society of Sigma Xi, Princeton Section, February 13, 1947; I.R.E. National Convention, New York City, March 4, 1947; I.R.E. Tube Conference, Syracuse, New York, June, 1947.

[4] A. V. Haeff, "A Memory Tube", *Electronics*, Vol. 20, No. 9, p. 80, September, 1947.

[5] Jan Rajchman, "The Selectron — A Tube for Selective Electrostatic Storage", *Mathematical Tables and Other Aids to Computation*, p. 359-361, Vol. II, No. 20, October, 1947.

[6] Jan Rajchman, "Recent Progress on the Selectron", oral progress reports presented at Association for Computing Machinery, Oak Ridge, Tennessee, April 19, 1949; I.R.E. Tube Conference, Princeton University, June 20, 1949; Second Symposium on Large Scale Computing Machines, Harvard University, September, 1949.

[7] S. H. Dodd, H. Klemperer and P. Youtz, "Electrostatic Storage Tube", *Electrical Engineering*, Vol. 69, No. 11, p. 990-995, November, 1950.

[8] Jan Rajchman, "Recent Experiences with the Selective Electrostatic Memory Tube", Conference on Electron Tubes for Computers, Atlantic City, December 11, 1950.

use in the broader field of automatic handling of coded information where temporary storage with rapid access to a moderate amount of information is required.

Description of the Tube

The present model of the selective electrostatic storage tube has 256 storing elements. Figure 2 is a photograph of the tube. It is 3 inches in diameter and 8 inches long. The diametral and axial cross sections of the tube are shown in Figures 3 and 4.

Eight elongated cathodes of rectangular cross-section (.020 × .040 inch) are located in a diametral plane of the tube. Between and parallel to the cathodes are a set of nine nickel-coated copper bars of square cross-section (approximately .1 × .1 inch), referred to as vertical bars, the axis of the tube being vertical in normal operation. The vertical selecting bars are connected into 6 groups V1, V2, V3, V4 and V′1, V′2 as shown in Figure 5. On either side of the plane of the cathodes and V-bars there is a set of 18 parallel nickel bars of square cross-section (approximately .05 × .05 inch) at right angles to the V set. These two sets of horizontal selecting bars sandwich the cathodes and V bars as do all other electrodes of the tube, the tube being symmetrical with respect to the plane of the cathodes. The 36 H bars are connected into 12 groups: H1 to H4 and H′1 to H′8, as shown in Figure 5. The spaces between the V or H bars are gates for the flow of electrons which can be selectively opened or closed. Since there is one more bar than intervals between bars, 9 vertical bars are required to form 8 gates. The two sets of horizontal bars are divided by the central cathode supports, making it necessary to have 36 bars for 32 gates.

On either side beyond the horizontal bars there is a collector made of two flat metal plates perforated with round holès in register with thè windows formed by the V and H bars. The collector mask plate, with relatively small holes (.040 inch), is backed by the collector spacer plate with larger holes (.120 inch). (See details in Figure 6.) Adjacent to the collector plates there are two mica sheets, the front and back micas, perforated to match the collector holes. These mica sheets hold between them 128 metallic storing elements which are insulated from all electrodes, i.e. electrically floating. These elements are tiny bodies of revolution made out of steel on an automatic screw machine and nickel plated. The conical heads of these elements protrude in the holes of the collector spacer plate which shields them completely from one another. The heads have a .020-inch hole in the center. The enlarged diameter in the middle allows the elements to be held without individual riveting. Adjacent to the back mica is a metal plate, the writing plate, with matching holes into which protrude the tails of the storing elements.

Beyond the writing plate is another plate, the reading plate, with the same pattern. Still beyond, is a Faraday cage made of two perforated plates spaced .300 inch apart and closed on all four sides by

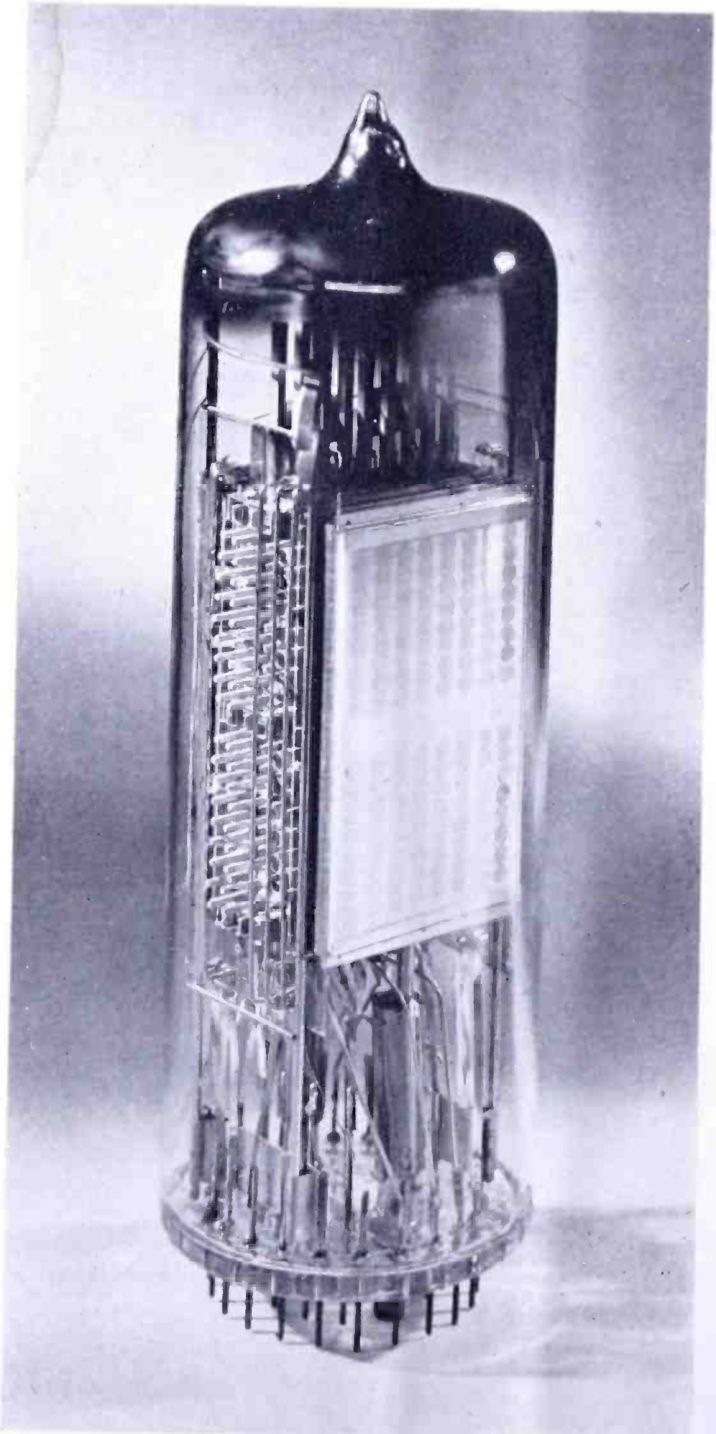

Fig. 2—Photograph of Selective Electrostatic Storage Tube

metallic strips. A glass coated with a fluorescent material, willemite, backs the outer plate of the Faraday cage. In the central plane of the cage there are 9 vertical wires located between the columns of holes of the plates. These reading wires are insulated from the cage by supporting mica strips and are connected together. The corresponding lead is shielded, even through the stem of the tube.

The construction of the tube is facilitated by several subassemblies: the cathode and V bars; the collector, storing elements and writing plate; the Faraday cages; and the side micas.

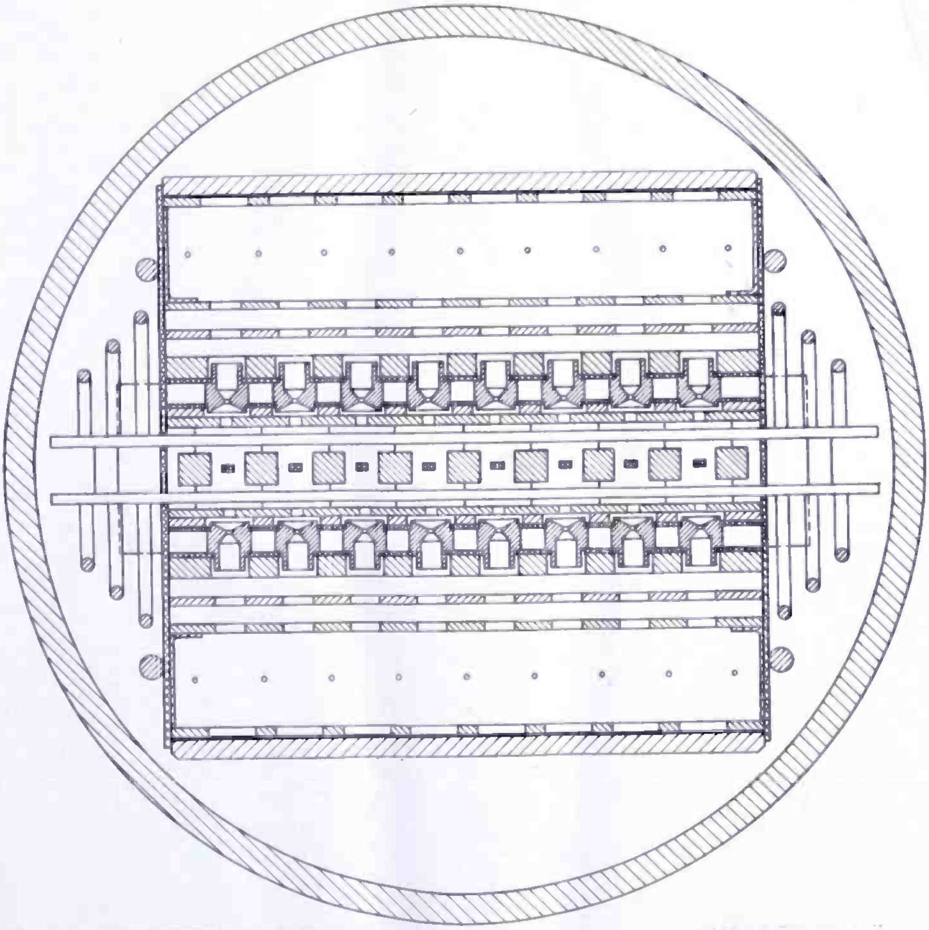

Fig. 3—Diametral cross section.

The cathodes and V bars are supported by small ceramics. The ceramics are held in a row between pairs of U-shaped strips. The eight cathodes are mounted in three such rows of ceramics and are sprayed together to insure uniform emissivity. A hairpin heater wire and a reinforcing straight tungsten wire are inserted subsequently in each cathode.

The mask and spacer collector plates, the front and back mica plates holding 128 storing elements and the writing plate form a tight

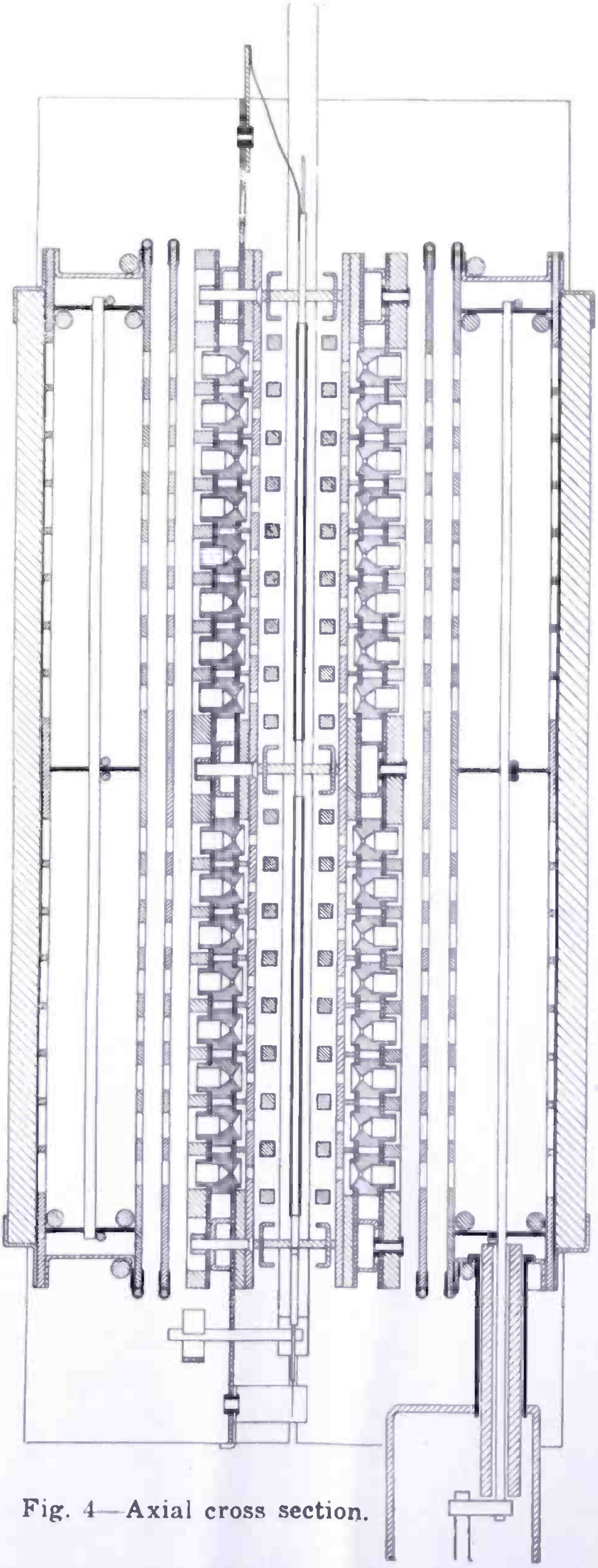

Fig. 4—Axial cross section.

subassembly which is riveted together at the upper and lower ends and in the center. To insure insulation between the collector and writing plate, there are two separate sets of rivets engaging the back mica in different locations where appropriate clearance holes are provided in the plates not held. The back mica of one of the two subassemblies is longer and carries terminals for the heaters and cathodes.

The side micas hold together two partial assemblies, each consisting of: 18 horizontal wires and their 6 connecting wires, one collector-writing plate subassembly, one reading plate and one Faraday cage.

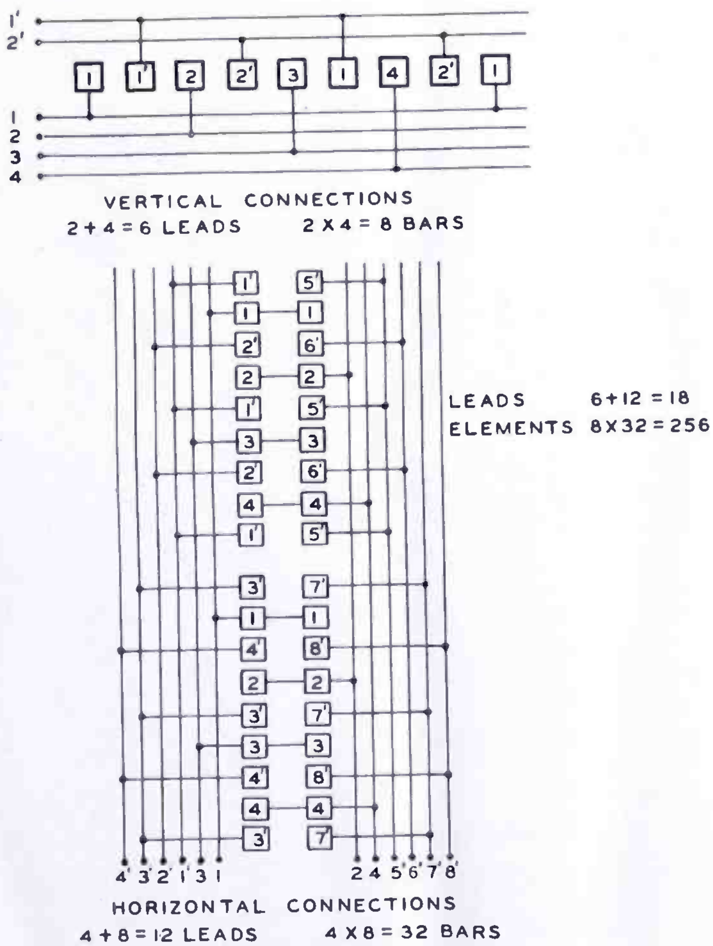

Fig. 5—Connections of selecting bars.

The two partial assemblies and the central cathode-V bars subassembly are tied together to produce the final mount. The cathode ceramics insure accurate equal spacings of the two halves of the collector on either side of the plane of the cathodes. Four rods connected to the two Faraday cages support the structure on the stem. The connections to the stem have been designed to be as direct as possible to avoid accidental short circuits. The stem has 34 pins: 22 on an outer circle comprising the 18 selecting leads and 4 for the supporting rods; and 12 on an inner circle including 2 coaxial output leads and all other 10 necessary connections.

The operation of the tube depends almost exclusively on the geometry of the electrodes, for while it depends also on reasonable thermionic and secondary emissions, it requires no exceptional emissivities or

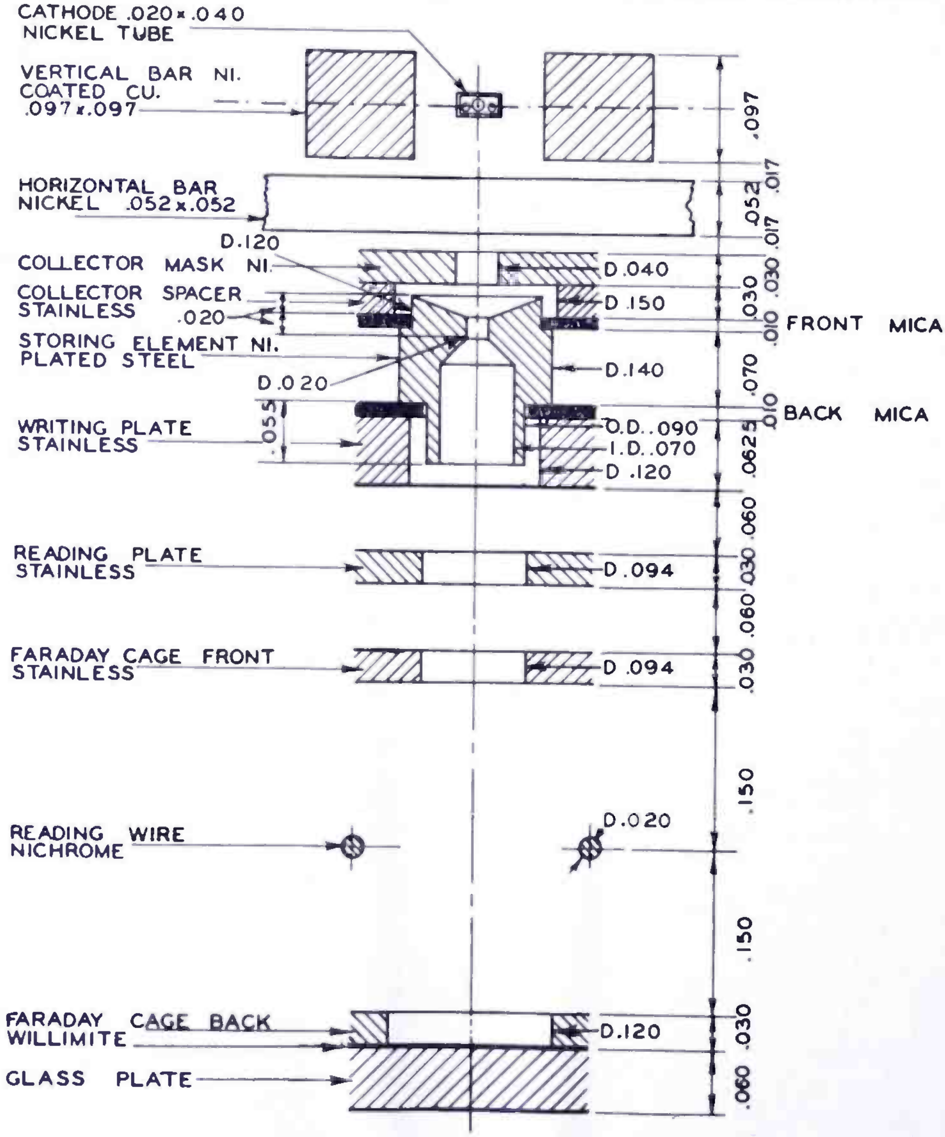

Fig. 6—Detail of one electron channel.

surface uniformity. The significant dimensions are shown in Figure 6 for a single electron optical channel. The clearances between parts and tolerances of dimensions are within reasonable shop practice. The mounting requires some precautions of cleanliness, particularly to avoid lint. No explicit activation of the secondary emission is necessary. The exhaust and cathode activation schedules are similar to those of a small transmitting tube.

Holding State

The operating condition is obtained by applying fixed direct-current potentials to the cathode, collector and Faraday cage and direct-current bias voltages to the selecting V and H bars, the writing and reading plates and the reading wires, as indicated in Figure 7. In the quiescent state of the tube, storing information previously written-in, all the selecting bars are at their more positive or opening potential which is the bias potential designed to be the zero potential of the cathode. In this condition electrons emitted from the cathodes are formed into 256 beams which are focused through the centers of the collector holes and are directed on the storing elements. Since these elements are not

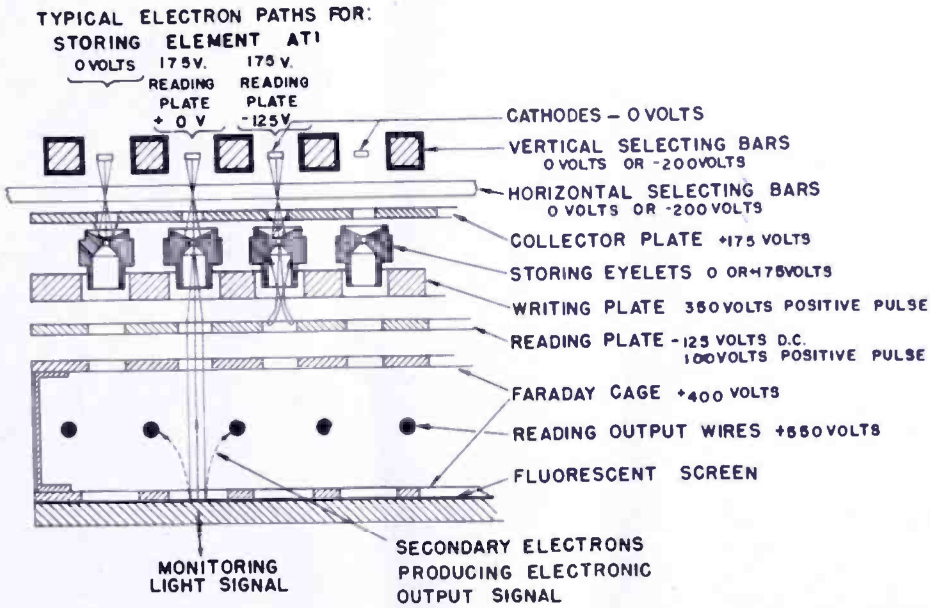

Fig. 7—Principle of operation.

connected anywhere and are electrically floating, except for ohmic leakage of the mica supports, their potential will adjust itself so that the algebraic sum of currents to them is exactly zero. It turns out that there are two naturally stable potentials for which this is the case.

This can be understood by examining the actual current to the storing element as a function of a forced variation of its potential as shown in Figure 8. When the element is more negative than the cathode, it repels any incurring electrons and receives only small positive leakage currents through the micas in contact with the collector and writing plate as well as some ions, present however perfect the vacuum, as shown exaggeratedly by the load line of the figure. As the element is made more positive, the incurring electrons produce a negative current much greater than the ohmic and ionic contributions and the net current passing through zero near the cathode potential, becomes negative. The zero-current potential is about half-to-one volt negative with respect to the cathode potential, because there are enough electrons with sufficient initial thermal velocities to overcome the resulting retarding potential and compensate the leakage current. At more positive potentials, secondary emission from the storing element tends to cancel the bombarding primary current, being a loss of negative charge. The two become equal at the so-called first crossover. For still more positive potentials, the secondary emission becomes greater than the primary bombardment and a positive current is obtained. Finally when the element exceeds the collector potential, the secondary emission is suppressed, due to the

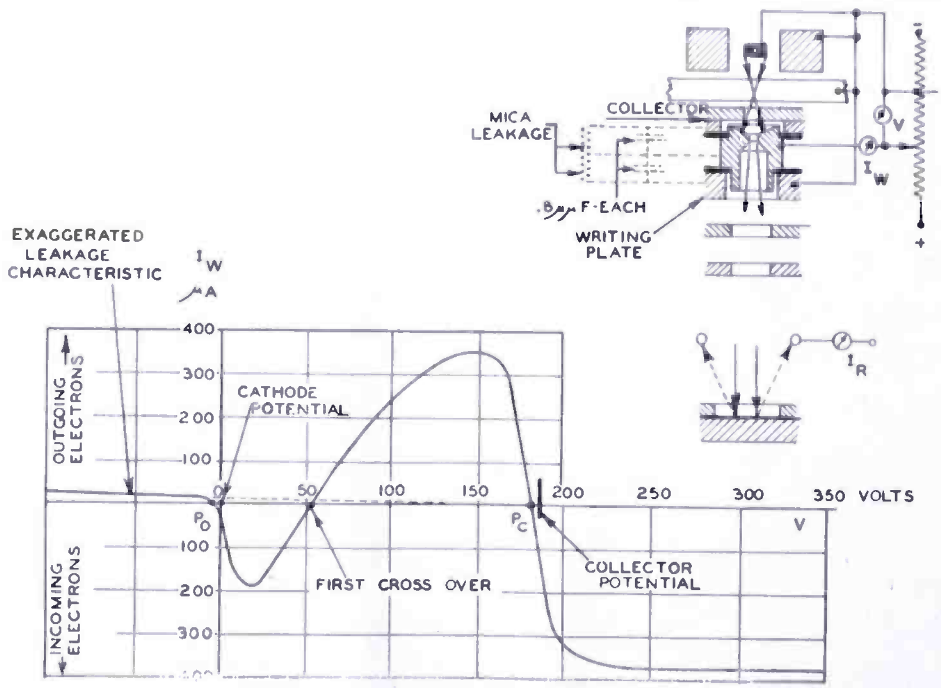

Fig. 8—Current-voltage characteristic of storing element.

retarding field between the element and the collector, and the full negative primary current is obtained. The current is zero at a potential of about one volt lower than the collector potential at which the resulting accelerating field opposes the retarding effect of the space charges due to the relatively large operating current density of the secondary electron current, to an extent just sufficient to obtain a secondary emission ratio of one. The zero-current potentials near the cathode and the collector are stable for a floating element because any deviation results in currents tending to restore equilibrium. The first crossover point, on the other hand, is unstable. The positive restoring current below cathode potential due to leakage and ions is usually negligibly small, so that for any potential up to the limiting half-to-one volt below cathode potential, there is essentially zero current. Consequently, the equilibrium potential may well be considerably below cathode potential in actual dynamic conditions prevailing during the operation of the tube.

It is clear from the above that any pattern of stable potentials (conveniently referred to henceforth as cathode or negative, and collector or positive potentials), once established on the storing elements, will remain indefinitely, as long as power is on the tube, without any deterioration whatsoever, by virtue of the holding action of electron currents present on all elements which counteract detrimental ohmic or ionic currents.*

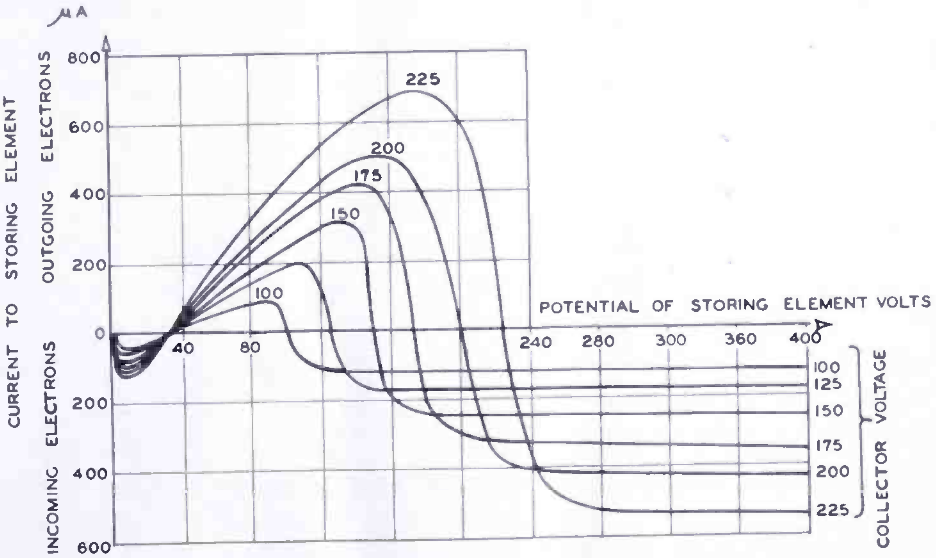

Fig. 9—Current-voltage characteristics for different collector voltages.

A family of current-voltage characteristics of the storing element was determined by direct measurement of a test element to which a lead was connected especially. Several remarks are of interest in connection with these curves shown in Figure 9 for different collector voltages. The negative current below first crossover, which has its greatest value at about 10 volts, insures the stability of the lower equilibrium potential (and could be used for charging the element as will be explained). The angle of the conical head of the element, designed to maximize this current, was chosen so as to force the low velocity electrons to approach the surface as normally as possible. The first cross-over potential is a sensitive measure of the secondary emis-

* This principle of storage has been adopted in beam deflected storage tubes, by A. V. Haeff in the Naval Research Laboratory tube, and by S. H. Dodd, H. Klemperer and P. Youtz in the Massachusetts Institute of Technology tube. See References 2, 4, 7, page 56.

sivity, and while for the particular test element it is 33 volts, it may be found to have any value between 30 and 100 volts. The maximum positive current flows at a potential considerably below collector potential, at which the secondary emissivity is higher, because an appreciable accelerating field is required to collect the space charge limited emission. The space charge accounts also for the lower-than-collector, zero current potential as mentioned above. While most secondary emission is suppressed for a retarding potential of 10 volts, the full negative current is obtained at 20 volts above collector. At higher voltages still, the primary current is practically independent of the potential of the element because this potential has no effect on the field near the cathode which is well shielded by the collector. These asymptotic values of the current to the storing element increase rapidly with collector voltage, not only on account of the enhanced space-charge-limited cathode emission, but also because the percentage of wasted current to the collector becomes smaller.

SELECTION

The registration of the incoming information into the tube or writing-in and the subsequent interrogation or reading from the tube, are made to a single or a few elements at a time and require the selection of the storing elements to which access is desired. This is accomplished by applying a negative pulse to all the V and H bars except one in each of the four groups V, V′, H and H′. The bars are connected in such a way that one and only one gate in each of the V and H directions will have its two limiting bars remaining at the bias-cathode potential, while all other gates will have one or both limiting bars at the pulsed negative potential. This can be verified by examining Figure 5. For example, if the element defined by V3V2′H4H5′ is selected, there are gates such as V2V2′ or H4H6′ in which one bar is pulsed negatively and the other remains at zero, other gates such as V2V1′ or H1H6′ in which both bars are pulsed, but only in the selected gates V3V2′ and H4H5′ both limiting bars remain unpulsed. When a V or H bar is sufficiently negative, it cuts off almost entirely the emission from the corresponding adjacent cathodes or cathode regions and the small remaining part is deflected and does not reach the hole in the collector. Of course when both sides of a gate are negative, the current is cut off even more because a negative potential barrier is formed through which no electrons can pass. It follows, therefore, that only the particular selected window, with its four bars at zero bias potential, will still have its original current while all others will be cut off.

The actual cutoff characteristics of the selecting bars have been determined by measuring the current passing through a collector hole as a function of the potential of one selecting bar, the three other surrounding bars being at zero bias. The ratio of this current to the full operating current of the opened window is shown in Figure 10 for different collector voltages. The current to most unselected elements is reduced, of course, by a much greater proportion since two, three or four bars contribute to that reduction. It can be estimated, on the basis of the writing and reading mechanisms described below, that

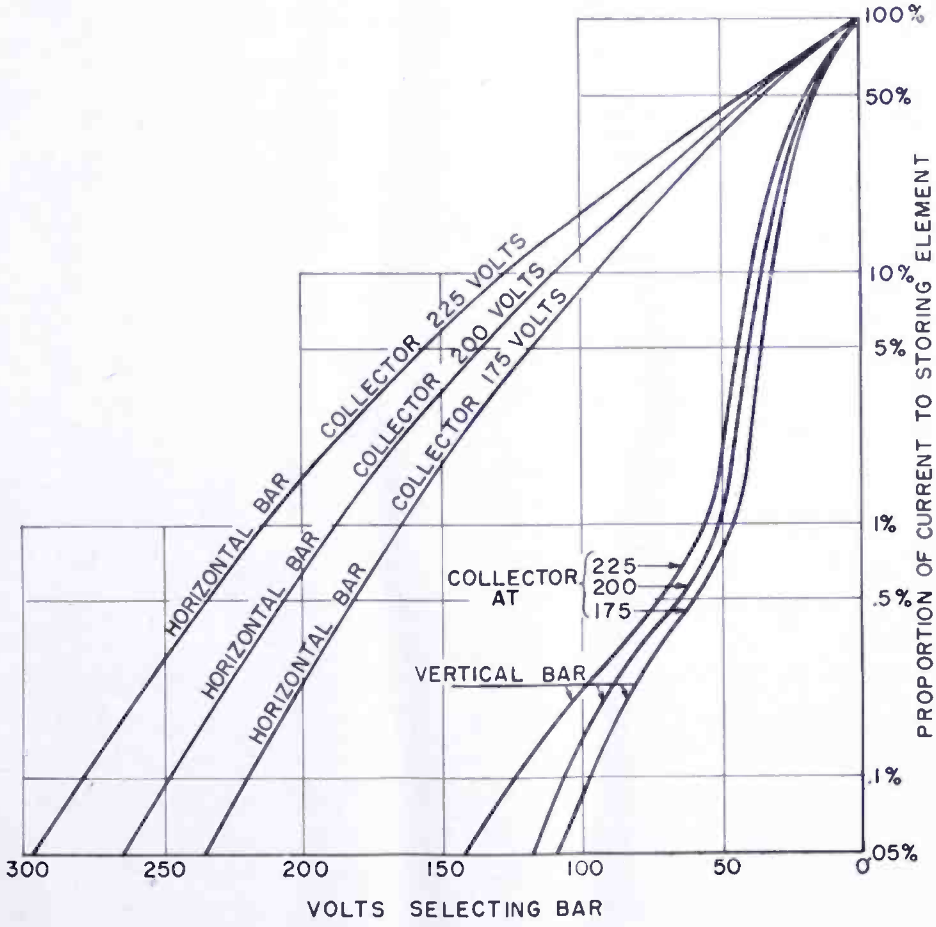

Fig. 10 — Cutoff characteristic of selecting bars. Voltages shown are negative.

there is completely negligible cross coupling between elements when less than 1 per cent of the full operating current is the maximum reaching any unselected element. This occurs for a selecting pulse on the horizontal bars about equal to the collector voltage, and for a quarter of that voltage on the vertical bars. For sake of uniformity of the driving circuits, equal selecting pulses may be used on all bars. A pulse amplitude of 200 volts will be adequate for a wide range of collector voltages. Since the bars are either at zero bias or negative, they draw no current. Their load is purely capacitive and varies from 14.5 to 28.5 micromicrofarads depending on the particular bar group. (See Table II, which lists capacities.)

The electron optical system formed by the cathode, selecting V and H bars, collector and storing element is reasonably efficient, as about 80 per cent of the electrons emitted from practically the entire length of the cathode are actually focused on the storing elements and only 20 per cent are wasted on the collector. It is designed to operate at zero bar bias voltage, which is not only most convenient for the select-

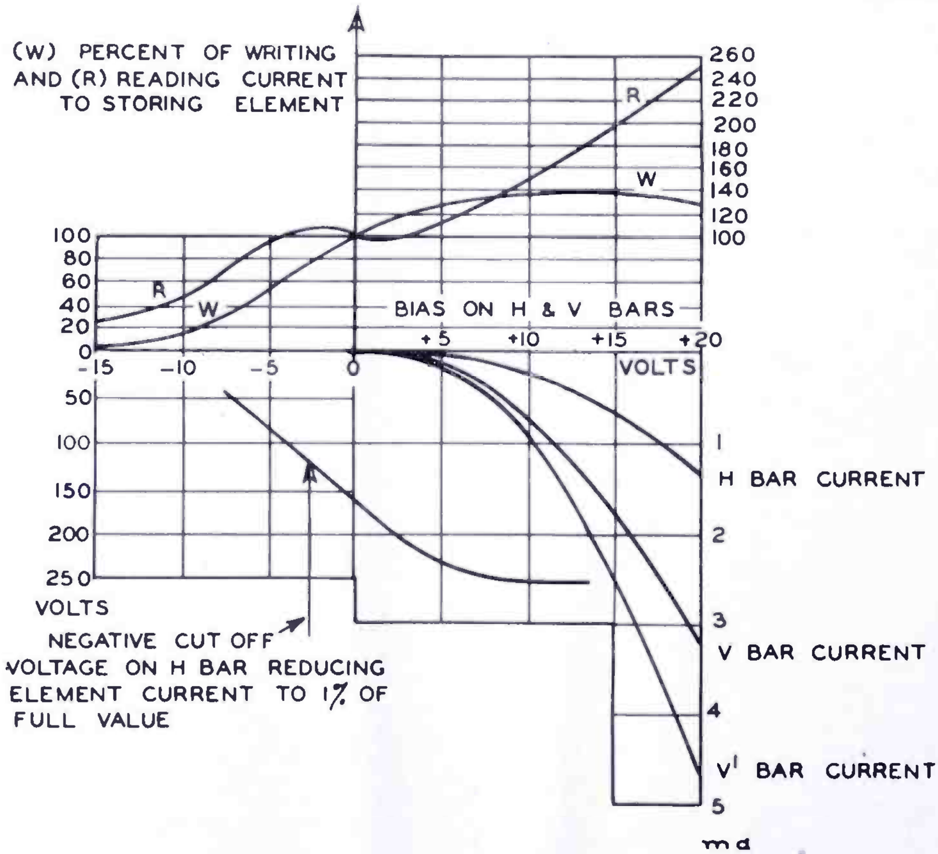

Fig. 11—Effects of selecting bar bias.

ing driving circuits, but results also in a two-electrode system in which the focus is independent of the magnitude of the collector voltage applied (except for some space charge effects). The effects of different bias voltages are shown in Figure 11. For negative biases, some saving in the required selecting cut-off pulse is obtained at the expense of writing and possibly reading currents. For positive bias, greater writing and reading currents are obtained, at the expense of increased selecting cut-off voltage and considerable currents to the vertical and horizontal bars. Since these currents differ greatly between bar groups, particularly well regulated driving circuits must be used to prevent bias dissymmetry to which the optics are very sensitive.

The principle of selection of this tube is based on the idea that both sides of a gate have control on the passage of electrons through it and that, therefore, combinatorial systems of connections are possible by connecting each side of the gate to appropriate sides of other gates. In fact, since this is done in both directions, a fourth power relation exists, in general, between the number of necessary connection groups and the number of controlled windows. Since a seal through the vacuum envelope and an external circuit is required for each connection group, the economy in the number of these groups is of great practical importance. In the present tube 18 leads control 256 windows, but a more spectacular result of the fourth power relation would be a tube with 128 leads controlling 1,048,576 windows. The combinatorial principles of area selection, the chief novel characteristic of this tube, are analyzed at some length in the appendix.

WRITING

To register an incoming information bit* into a particular storage element, that element is first selected by interrupting the current to all other elements except to it, as explained above. The selected storing element is then brought to the desired potential by a combination of the electronic current remaining on it—which may be keyed during the selection time—and the displacement current resulting from the pulsing of the writing plate to which all elements are capacitively coupled.

The writing method providing the shortest access time is as follows (See Figure 12): At the instant at which the selecting pulse has reached its most negative value and the element is truly selected, a positive pulse is applied to the writing plate of sufficient rate of rise and amplitude to cause the corresponding positive displacement current, by overriding the electronic holding currents, to raise the storing element potential by an increment equal to the cathode-collector potential. This brings an element originally at cathode potential near the collector potential, and one originally at collector potential to nearly twice that potential. If positive registry is desired, the writing pulse is made to decay sufficiently slowly to cause the displacement current to the element to be smaller than the maximum positive net electron current from it. Therefore, an element, brought from cathode to collector potential by the pulse rise, will be charged positively during

* One bivalued signal.

this decay and remain locked at collector potential. If negative registry is desired, the writing pulse is kept at its maximum value for a plateau time sufficiently long to allow an element, brought from collector potential to twice that potential, to be charged back to collector potential by the incident primary electrons. The writing pulse is then made to drop as sharply as it rose, resulting in a negative displacement current which overrides the holding positive electron current and brings the storing element to cathode potential where it remains locked. It is apparent that this writing procedure leaves unchanged the potential of the selected element which was originally at the potential to which it was driven. Therefore no erasure is necessary before writing, since the slowly decaying writing pulse will leave the element at collector

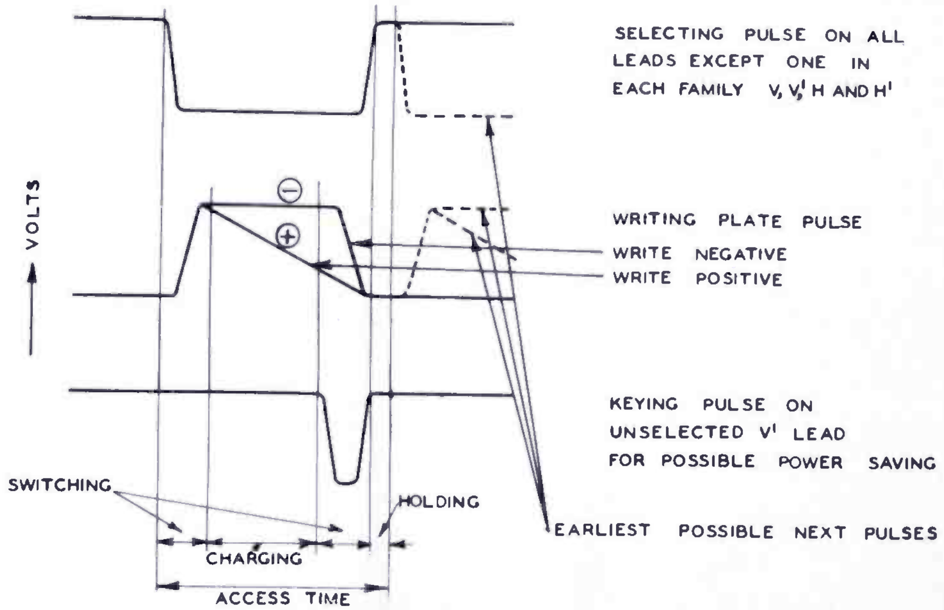

Fig. 12—Writing by writing-plate pulse modulation.

potential, and the square pulse will leave it at cathode potential, regardless of the initial condition of the element.

The minimum time t required to charge the storing element to the V volts between cathode and collector can be computed easily because the charging electronic current, i, is essentially constant. For negative charging, i is simply the constant value between V and $2V$ (see Figure 8), while for positive charging it is the maximum positive current flowing at a potental slightly below V to which the element is driven by the decaying writing plate pulse. At the operating collector voltage of 175 volts, these currents are about equal, both approximately 350 microamperes, because the effective secondary emission ratio is near 2. The capacity C of the storing element is made up of .8 micromicrofarad to the collector and of .8 micromicrofarad to the writing plate, giving a total of 1.6 micromicrofarads to be charged. The simple relation $Q = CV = ti$ gives a minimum writing time of $t = .8$ microsecond. The actual charging time which is the length of the decay or of the plateau of the writing pulses must be made longer than this minimum to allow for variations in current and capacity between elements as well as a reasonable safety factor. Times of 2.5 microseconds were actually found satisfactory for prolonged operation.

The sharpness required in the rise of the writing plate pulse and in its decay when it is square, can be estimated by computing the duration for which the displacement current to the storing element is just equal to the opposing electronic holding current. For the decay, this is precisely the minimum charging time, .8 microsecond, computed above. For the rise, the positive writing-in to an element orginally at cathode potential is the only critical case. Since the holding current for potentials below first crossover is at most a fifth the asymptotic charging current present above collector potential (see Figures 8 and 9), the limiting duration of the rise is 4 microseconds. The actual rise and decay of the writing plate pulse must occur in much shorter times than these limits in order to insure that the displacement current will be the controlling factor. Rise times of 1 microsecond and decay of .2 microsecond were actually found to be safe.

The power requirements of the circuits driving the writing plate so sharply are appreciable. The capacity division of the element between the collector and writing plate makes it necessary to use an amplitude of the writing pulse equal to about twice the cathode-to-collector potential, i.e., approximately 350 volts. Furthermore, each writing plate is a capacitive load of 112 micromicrofarads, or a total of 224 micromicrofarads with both plates in parallel. The driving power can be reduced considerably by keying-off the electron current to the selected element at the appropriate pulse decay (and rise) times in order to suppress the opposing holding action of this current, as shown in Figure 12. When this is done, these times may be lengthened at will. The consequent saving in writing plate power is obtained at the expense of additional keying circuits and some lengthening of the access time. The keying can be accomplished on any one of the four selecting bars surrounding the element, V, V', H or H', but most conveniently on the V' bar, since there are only two families of V' bars, $V'1$ and $V'2$. The keying pulses, applied to both bars, will be effective on the selected one while the other carries the full length selecting pulse. A first keying pulse could be used during the rise of both positive and negative write-in pulses but is not very important because the negative current below first crossover is very small (omitted on the figure). A second pulse is really important during the decay of the negative writing-in pulse which otherwise must be so rapid. This second pulse is detrimental to the positive writing-in because it diminishes slightly the charging current, but this effect is so small that it is simpler to introduce no polarity differentiation and always have it present.

Another writing method utilizes a standard shape writing plate pulse and keying of the current to the selected element to control the polarity of registration. The pulse has a rapid rise (or relatively slow one with power saving keying) followed by a plateau and a relatively slow decay equal respectively to the duration of the former square and decaying pulses (e.g., each 2.5 microseconds). At the end of the

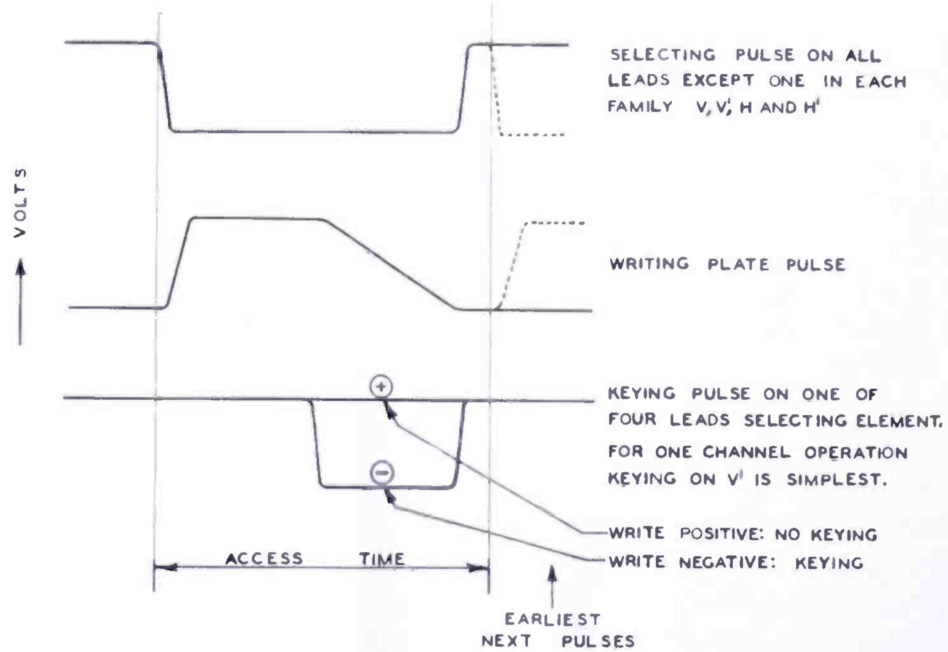

Fig. 13—Writing by selecting-bar current keying.

plateau time, the storing element will be at collector potential, either through the initial displacement current, or by electronic negative charging from twice the collector potential. If positive registry is desired, the current is not keyed-off and the element is charged positively during the decay and remains locked at collector potential. If negative registry is desired, the electron current to the selected element is keyed-off during the decay of the writing pulse and the displacement current brings the element to cathode potential (see Figure 13). This writing procedure is longer than the previous by one charging time (practically 2.5 microseconds). Its chief advantage is to allow simultaneous writing into two, four or eight elements. This can be accomplished as follows: When all the bars of one of the *V'*, *V*, *H* or *H'* families are left at zero potential while only one group of bars in each of the three other families is selected to remain at zero, several windows remain open: two for the family V', four for the family V or H and eight for the family H'. The several windows in the family not participating to selection may be keyed separately by the corresponding bar groups. For example, in the octet $V'1V2H4$, the eight selected windows may be keyed separately by the controls $H'1$ to $H'8$. By keying or not keying these individual controls during the decay of the writing pulse, negative or positive registry will be obtained in the corresponding storing elements. For writing-in purposes, the tube can, therefore, be considered as having one input channel with 256 storing elements, or two channels with 128 each, or 4 with 64 each, or finally as having 8 channels with 32 elements each.

Several other writing systems can be imagined in which negative and positive writing pulses are used to control the polarity of registration. Such systems are not as fast because negative charging depends on the relatively smaller current below first crossover, rather than the large current available above collector due to the transposition of negative charging into the higher-than-collector potential region (see Figures 8 and 9).

In all writing systems, immediately after the end of the writing pulse, the selection pulses on the bars end and current is re-established to all storing elements. During the selection time all storing elements, except the selected one, do not receive the benefit of the holding currents and retain their charges in the measure of the perfection of the mica supports and vacuum in the tube. The leakage current from the storing element is in part to the collector and in part to the writing plate. These parts flow in opposite directions for negative or zero biases of the writing plate. The resulting compensation is never perfect because the maximum permissible negative bias of —60 volts (which does not reduce the reading current passing through the element) is insufficient to make up for the difference in leakage through the front and back micas arising from their difference in temperature. At zero bias, which is the convenient recommended value, the natural storage time was always found to exceed 20 milliseconds. As this is very long compared to the few microseconds selection time intended in most applications, the unselected elements, after a violent potential excursion due to their coupling to the writing plate, will regain almost exactly their original potentials. The exact equilibrium potentials will be reached almost immediately thereafter by virtue of the stabilizing currents.

The input access time, or time between successive registrations to any two elements of arbitrarily selected addresses, is spent between switching, actual charging of the storing element and holding information in all elements. This is illustrated in Figure 12. The switching time extends from the origin of the selecting pulse to the beginning of the actual charging and from the end of charging to the end of the selection time. It is of the nature of "red-tape" and depends on the design and power of the driving circuits. A practical limit to switching time was found to be less than 2 microseconds. The actual charging time is inherent to the storage tube, has a theoretical limit of .8 microsecond and a practical value of 2.5 microseconds, as mentioned above. The holding of information need not be done for the long interval of natural storage of at least 20 milliseconds, but is inherent in the routine of most switching circuits which include a passage through the holding state between successive selections. Sufficient holding action is obtained when this passage is as short as feasible with any practical circuit, i.e., a fraction of a microsecond. The minimum operating input access time is, therefore, about 5 microseconds, half being spent in actual charging and half in switching and holding. Of course appreciable economy in the driving circuits can be realized with somewhat longer access times.

READING

The reading signal is derived from the electron current passing through the central hole in the storing elements. Part of the electrons aimed at the element are directed at that tiny hole (.020-inch diameter). When the element is positive, near collector potential, these electrons pass through the hole by virtue of their inertia, but when the element is negative, near cathode potential, it exercises "grid action" and the electrons are repelled and do not pass through the storage element. The presence or absence of the current through the element is, therefore, an indication of the state of the element.

Approximate electron paths resulting from the different conditions are shown in Figure 7. Figure 14 shows the reading current passing through the element as a function of the element's potential as determined from a test element to which a connection was made especially. The complete cutoff of the reading current at the lower equilibrium potential, which is a fraction of a volt negative with respect to cathode, is due principally to the depth (.020 inch) of the central hole and the length of the element's tail. The thermal electrons with initial energies sufficient to overcome the retarding potential of the element are stopped either by having too oblique a direction, or by the potential barrier developed by the space charges of returning electrons within the relatively large region in the element where the potential and fields are very low. For more positive potentials of the storing element the current through the hole increases slowly and at collector potential of 175 volts it is equal to about 35 microamperes, or one-tenth of the current directed to the head of the element. There is, of course, some variation in the reading currents from element to element, usually in the range of 20 to 40 microamperes. The reading current increases with collector voltage somewhat faster than the primary current because of improved current concentration at higher voltages.

In the quiescent state of the tube, with current present on all elements, the current which passes through the holes of all positive elements may be considerable, possibly as much as 256 times the reading current of a single element. This large irrelevant current is prevented from reaching and overloading the output circuits by the

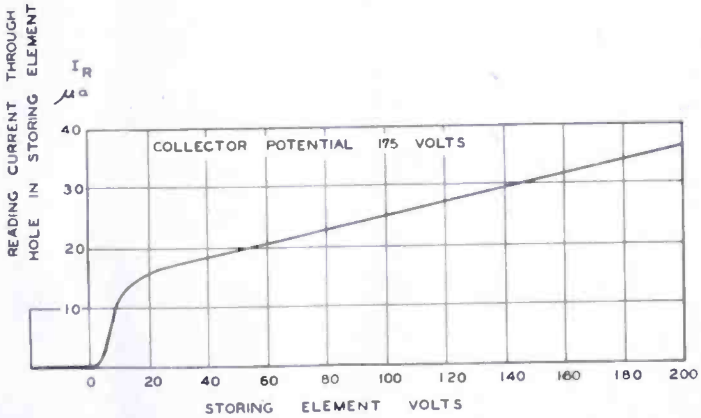

Fig. 14 — Reading current versus element potential (grid action).

reading plate which is biased at a negative potential, about —125 volts. When reading is desired, the interrogated element is selected by applying a negative potential to all selecting bars except the four defining it, as explained above. Immediately thereafter a positive pulse is applied to the reading plate which allows the current through the selected element—if any—to proceed to the output electrodes. The current penetrates the Faraday cage through the front plate, strikes the willemite coated on the glass plate backing the rear plate, where it produces an incidental light signal and causes the emission of secondary electrons. This secondary current is collected by the nine wires located within the Faraday cage, and constitutes the reading output signal. The selecting and reading plate pulses, as well as the resulting output pulse, if any, are shown in Figure 15.

The electrostatic shielding of the output electrode is almost perfect because there is negligible field leakage to the reading wires through the holes of the front plate of the Faraday cage, and the connecting lead is completely shielded, even through the stem of the tube. For convenience of wiring, the part of the shield sealed through the glass is insulated from the cage and may be at any desired direct-current potential, such as ground. Consequently, there are no capacitive pickups in the output circuits which may have resulted from the steep and large selecting and reading plate pulses, and the output signal is exclusively due to the 35 microamperes of reading current. The output voltage depends only on the desired speed of response. Since the capacity of the output electrode is about 20 micromicrofarads on each side of the tube, several tenths to one volt can be obtained for a time constant of about one microsecond, while 5 or 10 or more volts are available for time constants in the tens of microseconds. The reading access time, or time between successive interrogations of elements of arbitrary address, is spent between switching, reading and holding. Considera-

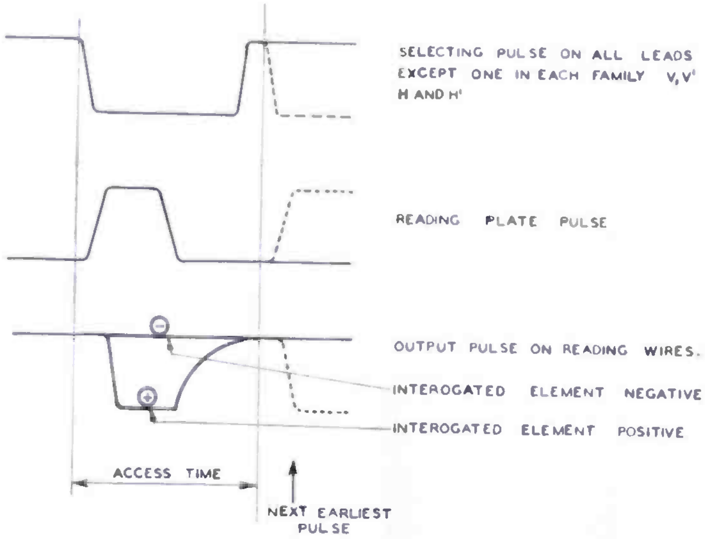

Fig. 15 — Pulses in reading.

tions concerning switching and holding times made previously for writing apply to this case. These times depend only on the circuit capabilities, and amount to 2.5 microseconds for reasonable assumptions. The reading time determines the obtainable output voltage and may be very short if adequate amplification is provided. The practical minimum reading access time is, therefore, about 3 microseconds. When it is made 5 microseconds, or equal to the writing access time, 2.5 microseconds are available for reading and an output of about half a volt is obtainable.

It is obvious that the interrogation of any element can be repeated indefinitely since it is derived from a current controlled by the potential of the storing element but playing no role in maintaining that potential. It is clear also that writings and readings to any element can be interlaced in any arbitrary manner. Moreover, since holding occurs automatically in the waste time following switching-between-elements, no holding routine need be provided. In fact the tube possesses truly random access since input or output access to any arbitrarily selected element is independent of any previous history.

The control characteristic of the reading plate of Figure 16 shows the percentage of output reading current as a function of the reading plate potential. This current reaches saturation when all the current passing through the storing element, passes also through the reading plate and none is reflected. It is apparent that the potentials of the writing plate and Faraday cage have no influence on this saturation value but determine the reading plate cut-off voltage. In order to avoid a reading signal due to the pulsing of the writing plate—from its zero bias to about 400 volts—the reading plate must have a sufficient negative bias of —125 to —150 volts, depending on the Faraday cage

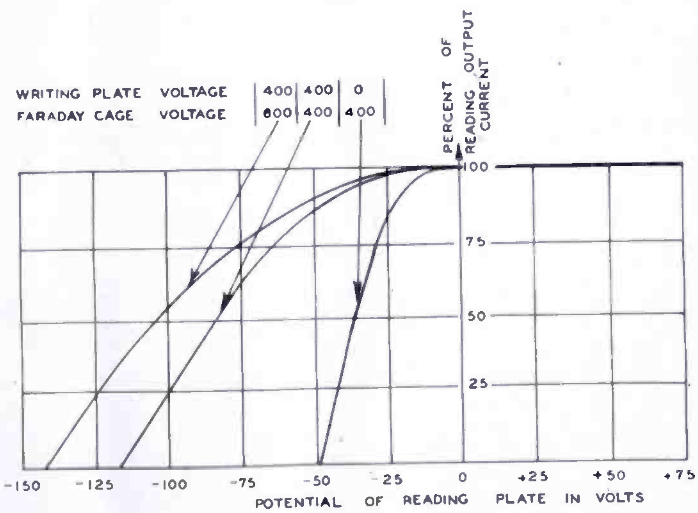

Fig. 16 — Reading plate control characteristic.

potential. In that case a reading plate pulse of 100 volts is required to obtain the full reading current. If the reading output due to the writing plate pulse is gated out by an external circuit, a reading plate voltage of only 30 volts is required with —50 volts bias. The reading plate draws no current, even when it is positive, because the reading currents are perfectly focused through its holes. It presents, therefore, a purely capacitive load to the driving circuit, which is 52 micromicrofarads on each side of the tube or 104 micromicrofarads total.

The current which is repelled by the reading plate when it is negative, is focused through the holes of the writing plate towards the back of the storing elements. While some of this current returns to the cathode or is reflected in the cathode region to strike the front of the elements, most of it strikes in the back of the elements, where secondary emission is suppressed for lack of collecting field. Consequently, the equilibrium potential shifts slightly to a negative value at which additional secondary emission from the front compensates the back bombarding current. This effect is negligible in practice because the reading current is only a tenth of the writing current.

The current voltage characteristic of the reading wires is shown in Figure 17, and resembles the desirable characteristic of a pentode since the current is independent of voltage for all values higher than about 150 volts with respect to the Faraday cage potential. The current to the reading wires is supplied by the secondary emission of the current entering the Faraday cage. This current comes mostly from the phosphor, but also, due to imperfect focusing, from the hole's sides of the front and the face of the back plates. The secondary emission from the phosphor is necessarily equal to the primary current striking it, since the net current to the insulated surface is zero at equilibrium, and, in general, it will be collected in part by the Faraday cage and in

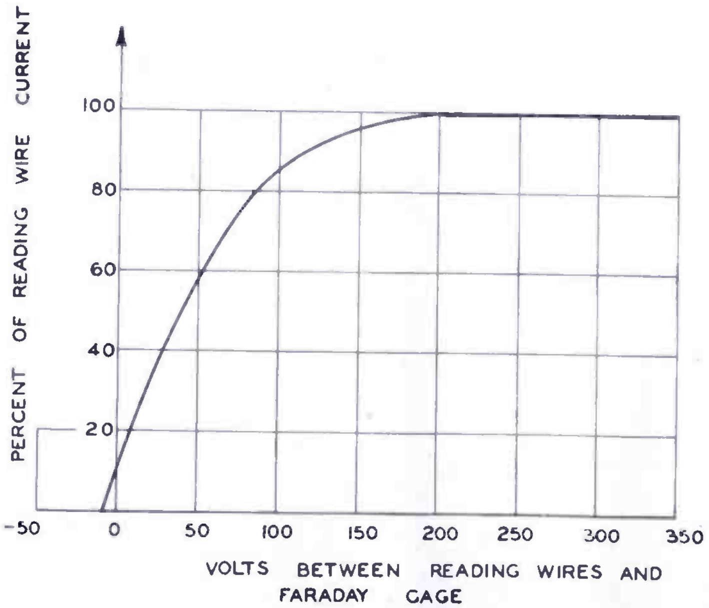

Fig. 17 — Reading wires current-voltage characteristic.

part by the reading wires. As the potential of the reading wires is gradually increased with respect to that of the cage, the equilibrium potential of the phosphor (or equilibrium distribution on the bombarded area) rises and a greater proportion of the secondary electrons is collected by the wires. The relatively high potential difference of 150 volts is necessary to saturate this division process. It is interesting to note that the phosphor, like the storing element, may assume the cathode equilibrium potential, particularly when low voltages are used on the Faraday cage. Usually only the center of the phosphor areas is dark, i.e., at cathode potential, while the outer ring is at a higher potential and gives light. In any case, the reflected primaries are almost totally collected by the reading wires, and this condition results in no detrimental effect other than spoiling the aesthetic value of the monitoring light signal.

When the reading plate is set above its cutoff voltage, the current passing through all positive elements is allowed to strike the fluorescent screen and a pattern of the stored information is obtained. This display, convenient for checking the operation of the tube itself, is most useful for monitoring the computing or other information-handling-machine in which the tube is the central information store. The intensity of the light depends on the potential of the Faraday cage, and is limited only by possible burning of the phosphor. Satisfactory indications in normal room illuminations are obtained at 350 volts and voltages up to 800 volts are allowable. The potential of the reading (and writing) plate determines the focusing of the indicating spots, and may be adjusted at will when the tube is being viewed.

Characteristics, Circuits and Applications

The operating characteristic values of voltages and currents of the tube are listed in Table I. These values were found to be reasonable averages. Table II is a list of electrostatic capacities of pulsed electrodes with respect to all other electrodes of the tube.

When the tube is operated with a single access channel to its 256 storing elements, the writing and reading plates and the reading wires on the two sides A and B of the tube are connected in pairs, and all 18 selecting leads are used separately.

Simultaneous writing and reading access to two storing elements can be obtained as a natural result of the symmetry of the tube which may be considered as two tubes with 128 storing elements each. By connecting the selecting leads H' in pairs, 1′ and 5′, 2′ and 6′, 3′ and 7′, and 4′ and 8′, it is apparent that two elements are selected simultaneously in similar locations on opposite sides A and B of the tube (see Figure 5). The two input signals to be registered modulate separately the shapes of the pulses on the writing plates A and B. The reading plates A and B are pulsed together and the output signals are detected separately on the reading wires A and B.

As was mentioned previously, simultaneous writing into more than two elements is possible with modulation by bar keying. However, simultaneous reading is possible on at most two channels since there are only two reading wire outputs. This restriction is not significant in applications requiring parallel writing in many channels and serial reading from one or two channels.

In most applications of the tube, binary numbers will be used to specify the address of the information to be stored or detected. Therefore, a conversion is required between the binary signals and the inputs on the selecting leads. This can be accomplished by matrix circuits which are particularly simple in this case since the families of selecting leads are powers of two: two leads in V', four leads in V and H and eight leads in H'. These circuits must keep the leads at their zero bias for the holding state of the tube, and upon a command to select, must impress negative pulses on all leads except the ones selected by the binary inputs. Resistance, crystal or multigrid matrix circuits of various kinds can be designed for this gating operation (see Figure 18).

A circuit for one of the 2 × 4 matrices illustrates partially one solution (see Figure 19). The address of the information is assumed to be in a register. Two flip-flops of the register control four double control-grid tubes (e.g., 6AS6), so that one and only one conducts for any combination of flips or flops. The plate resistance of each double-grid tube is also connected to the plate of a paralleling triode (e.g., ½ 6J6). The grids of all four triodes are connected together and biased to cathode potential. Consequently, the four plates are at their relatively lower potential. This causes the bar-driving tubes (e.g., 6L6) to be non-conducting and leaves all the selecting bars at their bias potential, as required for the quiescent state of the storage tube. To select, a commanding pulse is applied to the paralleling triode grids and renders them non-conducting. This leaves no current on the three unselected plates of the double-grid tubes which will take their positive (+B) potential and produce thereby selecting pulses on the corresponding bars. The fourth plate, on the other hand, will have current from the selected double-grid tube and, therefore, no pulse will be produced on the corresponding selected bar. Circuits, more ingenious than this illustrative example, can be designed with smaller tubes and reduced power.

In the memory devices required for most computing machines, it is desirable to store under a given address, a group of bivalued signals—or "words" representing in coded form, numbers, letters, or command symbols—rather than merely a single signal. This can be accomplished by paralleling as many tubes as there are signals in the word (or half that number when two channels are obtained from each tube). The selecting leads of same identity of all tubes are connected together. A common matrix controls the selecting buses through suitably powered driving amplifiers (see Figure 20). Access is simultaneous to the identical address in all tubes and information is stored or detected in parallel to or from all tubes (or tube channels) through separate writing and reading circuits. Each individual writing circuit provides, on the writing plate, a triangular or square pulse depending on the polarity of the signal to be registered. The reading circuits amplify the output signals from the individual tubes. The size of these circuits depends greatly on the desired speed of response. For 5-microsecond

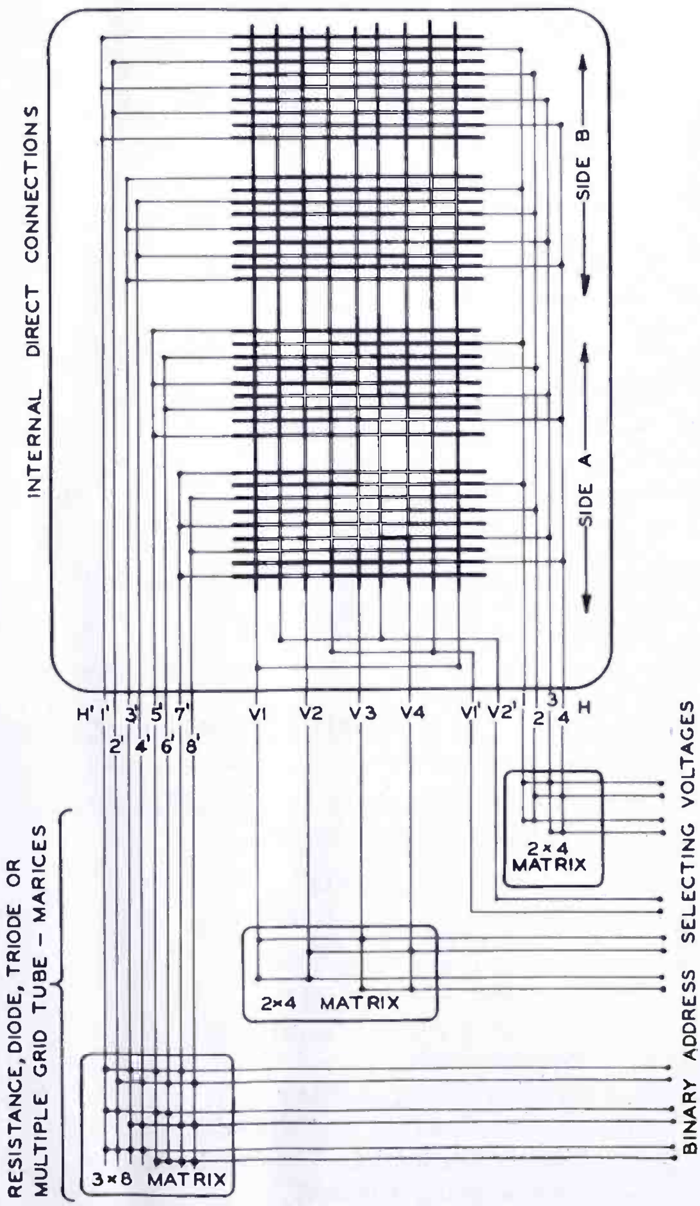

Fig. 18 — Internal connections and external matrices.

access time, reasonable designs require 4 tubes for the writing and 2 for reading.*

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* From circuit developments by I. E. Grosdoff in connection with memory unit including a score of these storage tubes.

Table I — Operating Characteristics

Heater Voltage	50	Volts
Heater Current	.7	Ampere
Cathode Current: all gates open, collector 175 volts	100 ± 10	Milliamperes
one gate open, collector 175 volts	350	Microamperes
all gates closed	0	Microamperes
Bias selecting, V and H Bars	0	Volts
Cutoff voltage: for single H bar—to 1 per cent of open gate value—	Collector voltage	
for single V bar—to 1 per cent of open gate value—	¼ Collector voltage	
Current to V and H bars	0	Microamperes
Collector voltage: averaging operating	175	Volts
Collector current: all gates open	Cathode current	
all gates closed	0	Milliamperes
Writing plate: Bias	0	Volts
Current	0	Microamperes
Writing Pulse	(Twice collector) 350	Volts
Reading plate: Bias, operating	—125	Volts
Bias, monitoring, spot focusing	—30 to +30	Volts
Current	0	Microamperes
Pulse	+100	Volts
Faraday Cage: Voltage	300 to 900	Volts
Current (with reading wires at specified voltage)	0	Milliamperes
Reading Wires: Voltage above Faraday cage	150	Volts
Current, per positive element	15	Microamperes min.
Current monitoring condition	10	Milliamperes max.

Table II — Electrostatic Capacities

All capacities are averages, in micromicrofarads, to all other electrodes:

Vertical Selecting Bars	
V′1, V′2 and V1	28.5
V2, V3, and V4	14.5
Horizontal Selecting Bars	
H1, H2, H3, H4	28.5
H′1, H′3, H′5, H′7	22.5
H′2, H′4, H′6, H′8	16.5
Writing Plates — Each Side	112
Reading Plate — Each Side	56
Reading Wires — Each Side	20
Storing Element to Collector	.8
Storing Element to Writing Plate	.8
Total Storing Element	1.6

Performance Tests

About 50 tubes were built in the laboratory according to the design described above, with minor modifications from tube to tube. The tubes were first tested with direct-current or simple-pulse techniques. Uniform characteristics of selection and monitoring display were observed in all tubes, as these depend on built-in geometrical properties. The uniformity of cathode emission and of the secondary emission from the storing elements was gradually improved with experience. In the latest tubes there is approximately 20 per cent variation of primary current available at the elements. The variation of secondary emission which is more difficult to measure is certainly somewhat greater. The phosphor screens were found to be uniform in light output.

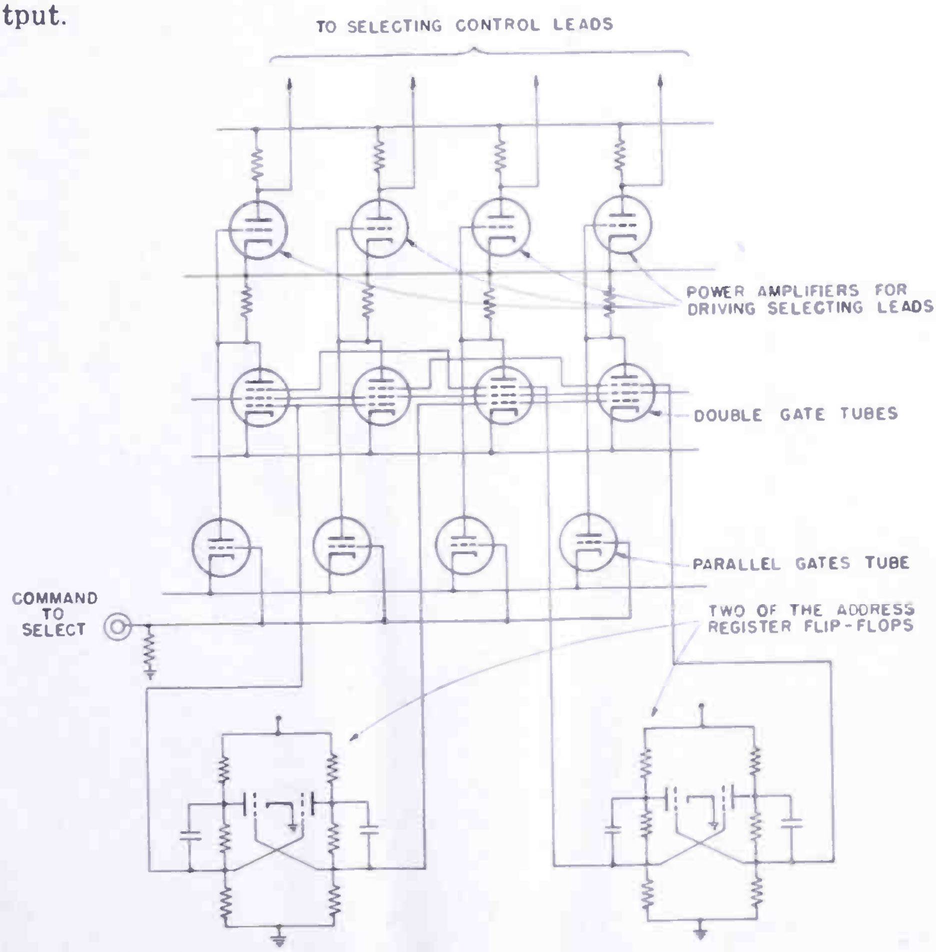

Fig. 19—Typical matrix for four selecting leads.

A test unit was built in which the tubes would be in dynamic conditions as similar as possible to those of actual use in a computing machine. The test system consists of setting some arbitrary pattern of information in one tube, interrogating the elements of that tube in succession and registering the results in the corresponding elements of a second tube. The initially stored pattern will, therefore, appear in both tubes No. 1 and No. 2 as shown in Figure 21. The pattern of

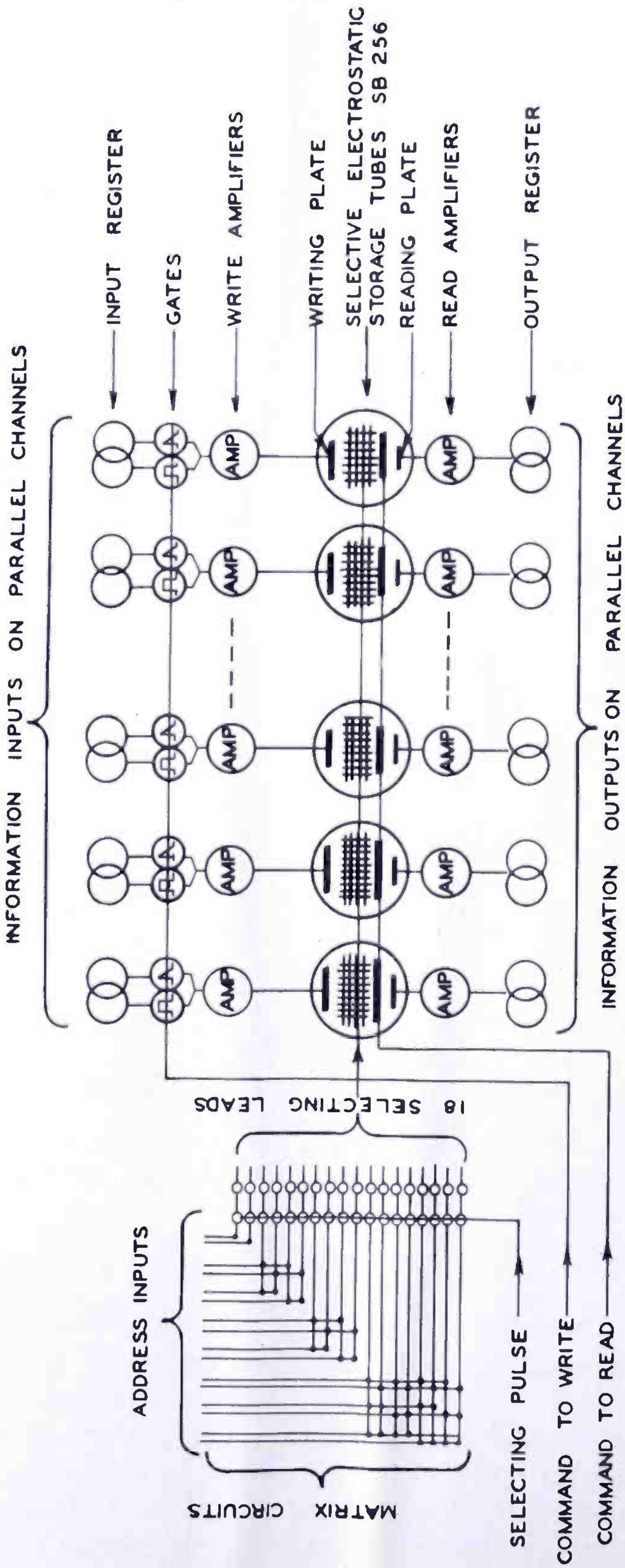

Fig. 20—Typical use of selective storage tubes in parallel arrangement.

tube No. 2 is then read off element by element and is registered with reversed polarity into tube No. 1, so that a pattern in which all initially positive elements are negative and vice versa, appears in tube No. 1. The test consists of letting this back and forth transfer proceed automatically at high repetition rate and observe whether the initial pattern remains unspoiled. Runs of seventeen hours without failure have been observed. The occasional transient failures were probably due to the circuits, as their number seemed to decrease with circuit improvements.

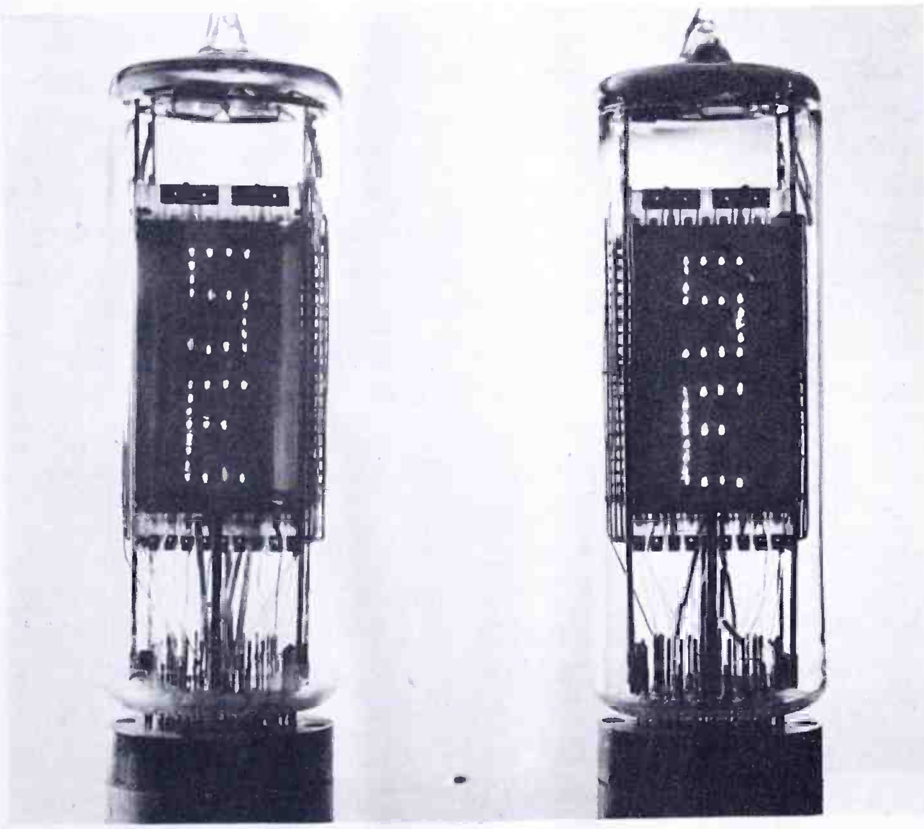

Fig. 21—Pair of tubes in life test.

The life of the tube depends principally on cathode emission and secondary emission of the storing elements. The required emissivity of the cathode is considerably lower than in radio receiving tube practice. Therefore, it is very likely that adequate emission will be available for a very long time. The secondary emission of the nickel-plated steel elements is influenced by the barium oxide evaporated from the cathodes as well as the intense electron bombardment which has an oxide reducing action. These two mutually compensating effects are likely to produce a fairly stable secondary emitter.

Some empirical data, accumulated to this date, supports these expectations. A pair of tubes after 1500 hours of operation had still their initial cathode emissions. The secondary emission, after a drop in the initial 30 hours, leveled off to a steady value. Similar observations were found in shorter runs of several hundred hours on a dozen tubes.

CONCLUSIONS

After several years of research following the original conception of the novel principles of selection and storage, the tube has been developed in the laboratory to practical usefulness. The first steps towards manufacturing have been taken, and a series of development tubes referred to as C7761 have already been built. The Laboratories have undertaken to build a complete memory unit including 20 selective storage tubes, type C7761, with associated circuits and power supplies. The halves of the tube are used separately to provide 40 parallel information access channels. The device is to operate with an access time of five microseconds.*

There is no doubt that a tube with superior performance will result eventually from manufacturing experience. Engineering of a full-size memory unit has already resulted in improved circuits. Early completion of the unit will also reveal valuable operational experience and further data on the life of the tubes.

The considerations on the future possibilities for better tubes based on the same principles will become more realistic in the light of these experiences, but some aspects may be speculated upon at the present time. The versatility and usefulness of computing and information handling machines is primarily determined by the capacity and speed of its random access memory. To make up a capacity sufficient merely to justify the existence of the machines, scores of tubes are already necessary. For the desired larger capacities, the relative merits, of a few large tubes or many smaller ones, must be appraised by weighing the technological difficulties of building many elements into a single tube against the increase in wiring, circuit and servicing complications resulting from the use of many tubes. The optimum capacity with the present techniques is most likely above the present 256 elements, probably 512 and possibly even 1024. Numbers of elements other than powers of two are possible, of course, and it may be practical for decimal machines to have some simple multiple of a power of ten.

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* Circuit developments by I. E. Grosdoff.

The selective electrostatic storage tube described in this paper is the only truly random access storage device operating with bivalued inputs and outputs. To the author's knowledge, it is also the fastest memory for computers available at the present time. Some increase of speed is still possible by improving the electron optics, the secondary emissivity of the storing elements or the controls of the tube.

The tube can be used as an information storage tank in any machine handling digital information. Designed for the high speed inner memory of computing machines, it can be used also in other parts of these machines, particularly for the auxiliary memories associated with the arithmetical or control units.

ACKNOWLEDGMENT

The author wishes to express his appreciation to V. K. Zworykin for his guiding interest in this work and to acknowledge the contributions of many members of the RCA Laboratories Division. P. G. Herkart gave invaluable help in tube technique problems throughout the project. G. W. Brown, now with Rand Corporation, made essential original contributions to the theory of selection. M. H. Mesner was responsible for all circuits and tests during a major part of the project. M. Rosenberg assisted enthusiastically in all phases of the work throughout the project. J. A. Briggs helped in problems of mechanical design. J. E. Dilley and two Scandinavian students, Erik Stemme and Per Hals, were most helpful in circuit design and tests. The skill of H. E. McCandless in assembling the mounts contributed greatly to the success of the project.

The author appreciates also the interest which Prof. John von Neumann and his associates from the Institute for Advanced Study have shown in the early part of this project.

APPENDIX

THEORY OF CONNECTIONS OF PARALLEL BARS FOR COMBINATORIAL SELECTION

The selective storage tube uses, for area selection, a matrix of control electrodes which stop the electrons from a uniform electronic bombardment in all but a single location, or a predetermined number of locations, rather than deflecting an electron beam to the chosen location as is done in conventional cathode ray tubes. It is obvious that such a go, no-go control at positions materially fixed by the matrix provides a greater certainty of selection than is possible by controlling precisely the amplitudes of deflection of a beam. In addition, several or all locations may be selected simultaneously. These advantages are obtained at the expense of a large matrix of control electrodes. It turns out, however, that these electrodes may be simple straight parallel bars, which can be connected into a number of groups so relatively small that the system is practical even when the number of selected elements is very large, e.g., a few millions.

The system is based on the idea that the space between two adjacent bars can be considered as a gate for the passage of electrons, which is opened only when both limiting bars are at some relatively more positive potential, and closed when at least one is at a sufficiently more negative potential. Consider a loop of N parallel bars, separated by N spaces or gates, like the pickets of a closed fence. Because of the dual coincident control on each gate by its limiting bars, it should be possible to connect the N bars into G groups to produce gate selection by making two selected groups positive and the remaining negative, in such a way that N be proportional to G^2. This possibility would be obvious if the two bars controlling each electron gate did not affect also the two adjacent gates, as for example, the two control grids of a converter tube. It is a purpose of this appendix to show that the row of N gates formed by N bars can, in fact, be connected as economically as the $2N$ control elements of a row of N independent coincident devices in spite of the restriction imposed by the fact that there is a "fixed connection" of the two halves of each bar which connects necessarily adjacent gates. This restriction limits only the choice of the geometrical location of the gates with respect to the group codes to which they belong, and is irrelevant in storage devices for handling coded information, since identification of each element is by its selecting bars only and the "picture" formed by the elements plays no role.

First consider the simple case of a row of N coincident double grid tubes, where N is not prime:

$$N = P \cdot Q. \tag{1}$$

As coincidence is similar to the intersection of lines, it is convenient to think of all the grids grouped into one connection as lying on a line, the intersections of the lines being the coincident devices. The G groups of grids may be divided into two families of P and Q, represented by a set of P lines intersecting a set of Q lines, as on the symbolic diagram of Figure 22, each intersection corresponding to one tube.

$$G = P + Q. \tag{2}$$

Since each P (and each Q) group has Q (and P) grids, the $2N$ grids are accounted for. For a square pattern, $P = Q = G/2$, and $N = G^2/4$. Instead of selecting the two groups to determine a tube, one from the P and one from the Q family, the two can be chosen in more ways among the whole set G, provided that:

$$N = \frac{G(G-1)}{2}. \tag{3}$$

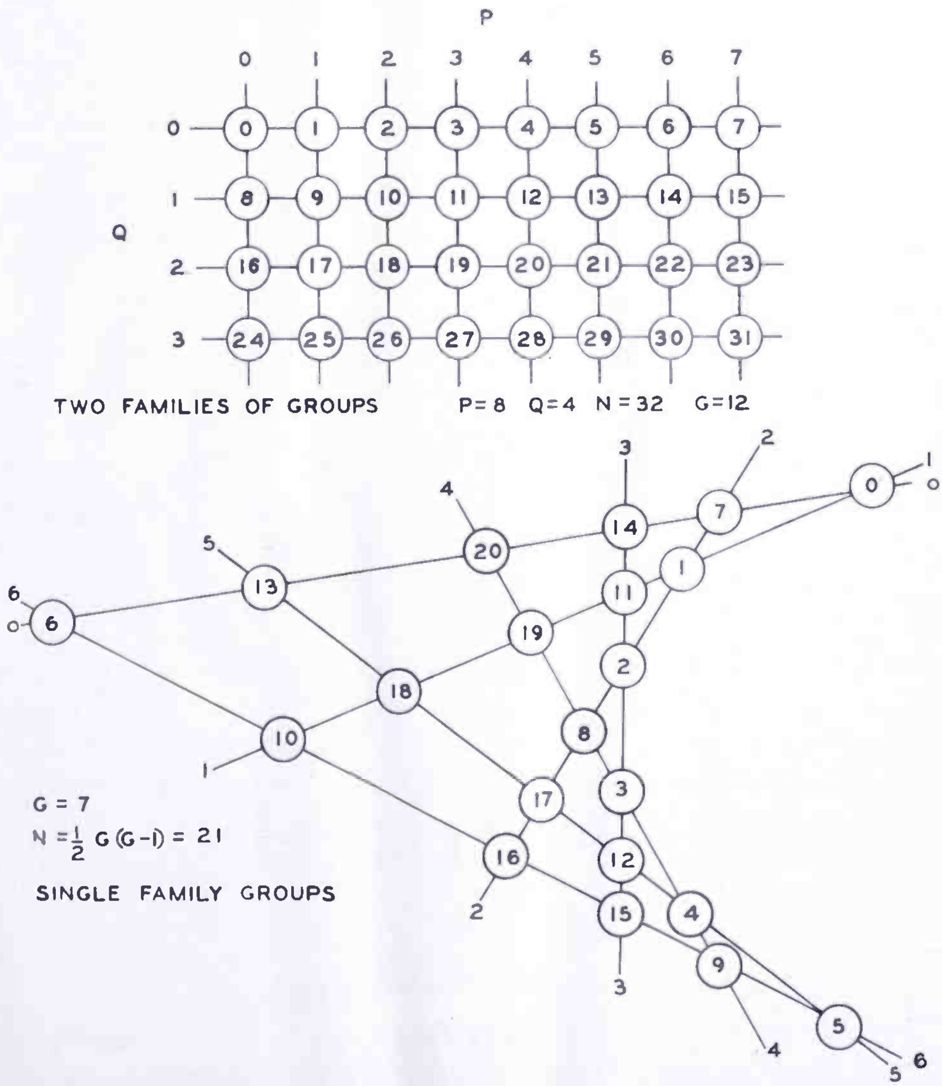


Fig. 22—Symbolic diagram of connections for single and two families of groups.

This is illustrated in Figure 22 where each of the G lines intersects all remaining $(G-1)$ lines. This one-family system is more economical than the two-family system, since N tends to $G^2/2$ for large G's.

In the case of either the one- or the two-family system, the N independent devices may be permuted arbitrarily on the N intersections so that there are $N!$ possible different systems of connections. In particular, the two-family system can be connected so that the selected element progresses monotonically (shown in Figure 22) along the row as the values of p are successively 0, 1, 2, . . . p, $(p-1)$, for $q=0$, and again through the cycle 0 to $(p-1)$ for $q=1$, etc., up to $q=Q-1$, so that the ordinal number n of the selected element is

$$n = P \cdot q + p. \tag{4}$$

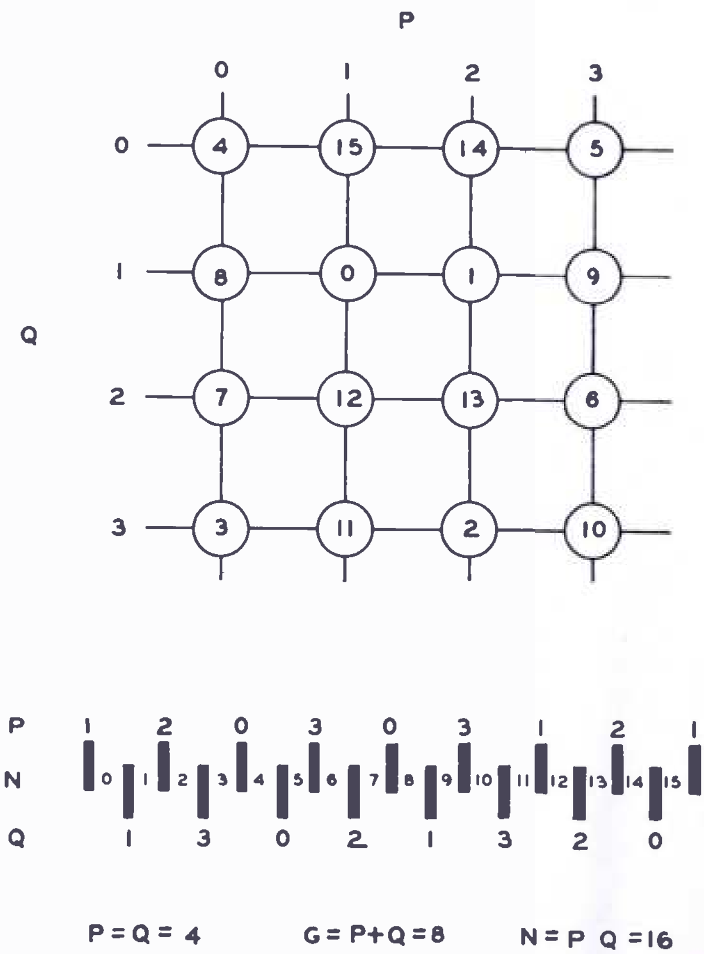


Fig. 23 — Bar connections in two-family system.

Consider now the electron gates formed by the bars. In a similar symbolic diagram the lines represent again the connections between bars and the intersections represent the gates. However, the bars are not on the intersections as were the grids, but between them. Typical connections of 16 bars ($P=Q=4$, $N=16$) for a two-family system are shown in Figure 23. A general method for finding a two-family connection system is as follows: Start at any intersection out of the $N=P \cdot Q$ intersection formed by the P and Q lines, call it gate $n=0$

(for example, $p = 1$, $q = 1$, Figure 23). Then along the q line (or p) containing gate 0, choose any intersection p (or q) for gate $n = 1$ (e.g., $p = 2$, $q = 1$, $n = 1$). The first q chosen ($q = 1$ in the example) is the group number of the bar separating gates $n = 0$ and $n = 1$. Since gate $n = 1$, as any gate, must be limited by a q and a p bar, proceed now on the p line containing gate 1 and choose any intersection for gate 2 ($q = 3$, $p = 2$, $n = 2$). Continue similarly for gates 3, 4 . . . up to $(n - 1)$, alternately moving along Q and P lines in such a way as to never choose an intersection already numbered. There are many trial routes which will end up by filling the whole matrix. Many others will end in an impasse on a filled P or Q line, there being still vacant points elsewhere. It will be shown now that if P and Q are even, there is at least one successful route or connection system such that any group in the P (or Q) family has a bar adjacent to a bar of every group in the Q (or P) family, once and once only.

The N bars may be numbered by the same numbers $n = 0, 1, 2, 3$ . . . $(n - 1)$ used for the gates, with a chosen sense of rotation. The bars with even n's will be assigned in succession to the groups of index p [$p = 0, 1, 2, \ldots$ to $(P - 1)$] as follows: $n = 0$ to $p = 0$, $n = 2$ to $p = 1$, $n = 4$ to $p = 2$, $n = 2p$ to p, up to $n = 2P - 2$ to $p = P - 1$. This cyclic assignment is done $Q/2$ times, creating P groups containing $Q/2$ bars each, accounting thereby for $PQ/2$ even numbered bars. This procedure can be expressed by the formula:

$$p = \tfrac{1}{2} R_{2P}(n) \text{ when } n \text{ is even or } [R_2(n) = 0] \qquad (5)$$

where the notation $R_a(b)$ is used for the remainder of the division of b by a.

The odd n's will now be assigned to the Q groups as follows: To $n = 1$ assign $q = 0$; to $n = 3$, $q = 1$; to $n = 5$ assign $q = 0$ again; and to $n = 7$, $q = 1$ again etc.; . . . n to $q = 0$ when $R_2\left(\dfrac{n-1}{2}\right) = 0$ and n to $q = 1$ when $R_2\left(\dfrac{n-1}{2}\right) = 1$ . . . up to $n = 2P - 1$ assigned to $q = 1$. To the second $2P$ cycle of the n bars assign the pair $q = 2$ and $q = 3$, in a similar way: the bar $n = 2P + 1$ to $q = 2$; $n = 2P + 3$ to $q = 3$; $n = 2P + 5$ to $q = 2$; $n = 2P + 7$ to $q = 3$ etc.; . . . $n = 2P + R_{2P}(n)$ to $q = 2$ when $R_2\left(\dfrac{n-1}{2}\right) = 0$ and to $q = 3$ when R_2

$\left(\frac{n-1}{2}\right) = 1$, up to . . . $n = 4P - 1$ to $q = 3$. The third cycle will be connected similarly to the pair $q = 4$, 5, the fourth to $q = 7$, 8 etc., with the last pair being connected to $(Q - 2)$ and $(Q - 1)$. Each group Q contains $P/2$ bars, and the N bars are divided into $Q/2$ cycles of $2P$ bars each. This accounts for the $PQ/2$ odd numbered bars. The assignment of the q bars can be expressed formally:

$$q = \frac{n - R_{2P}(n)}{P} + R_2\left(\frac{n-1}{2}\right) \text{ when } n \text{ is odd or } [R_2(n) = 1]. \qquad (6)$$

It is evident that this system of connections satisfies the required conditions. Indeed, for any selected set of p and q, there is a cycle of $2P$ bars corresponding to the pair of q's to which the particular q belongs, since all values of q were assigned, and in that cycle, one of the bars of the selected group must be adjacent to a bar of the selected p group, since the bars of the q group are adjacent to a bar of all p groups.

Figure 24 shows an example of the two-group family system, according to the teachings of the above proof, for $P = 8$, $Q = 4$, $G = 12$, $N = 32$, obtained by applying relations (4) and (5) for all values of n. This particular system corresponds to the zigzag line in the symbolic diagram which orders the gates as closely as possible to the monotonic order of Figure 22, obtained with the independent double gate system. A completely monotonic order is impossible since there must be alternate P and Q segments. There are, of course, many other possible combinations. Just how many less than the $N!$ connections, possible with the independent double control devices, are still possible with the row of bars, is an interesting problem of combinatorial analysis. This problem amounts to determining the number of ergodic paths in a mesh of points when walking is subject to the restriction of making consecutive steps in alternate directions.

When the bars are shown in a straight row, rather than a closed loop, the extreme bars must be considered adjacent (bars 0 and 31 on Figure 24), a condition realized in practice by an additional bar on one end connected to the bar on the other end. Such additional bars are necessary also when there are gaps in the row of gates. This is the case in the tube described, at the location of the central cathode-supports.

The economy in the number G of vacuum seals and external circuits of the two-family group system of N gates can be measured by the ratio of merit N/G or $PQ/P+Q$. This ratio is the greatest when $P-Q$ is the smallest. For example, when $N=32$, $N/G=2.67$ for $P=8$ and $Q=4$ and only $N/G=1.78$ when $P=16$ $Q=2$. The most efficient system is, of course, when N is the square of an even number,

SYMBOLIC SCHEMATIC

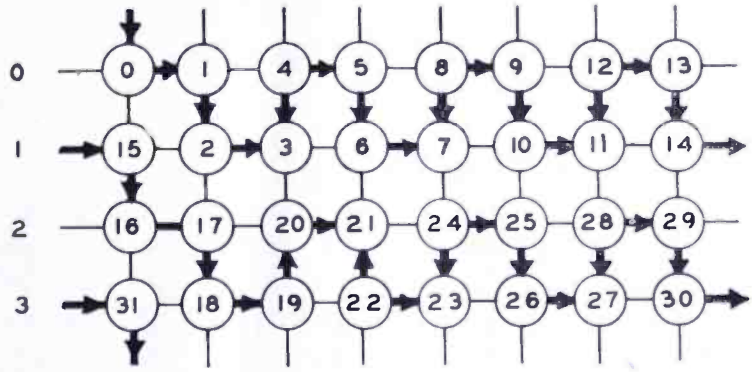
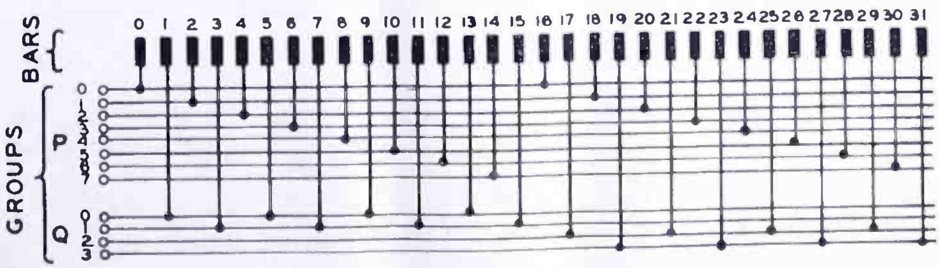


TABLE OF INDEX VALUES

n	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25	26	27	28	29	30	31
p	0		1		2		3		4		5		6		7		0		1		2		3		4		5		6		7	
q		0		1		0		1		0		1		0		1		2		3		2		3		2		3		2		3

CONNECTIONS



N = 32 P = 8 Q = 4 G = P+Q = 12

Fig. 24—Connections of 32 bars according to the two-family of groups system.

then $P=Q$ and the ratio of merit is $P/2$. It is obvious also that the advantage of the system of connections grows with its size.

There are more ways to choose two groups out of G than one group out of each P and Q, where $G=P+Q$, as was mentioned in the case of independent coincident devices. It turns out, as with the two-family system, that the single-family of groups system is possible with the bars when G is odd and also that there is no loss of economy in the connections but merely a restriction on possible choices of location of the gates. The number of gates must be of the form of Equation (3). Figure 22 was drawn so as to illustrate the case of N bars (as well as a special case of N independent double coincident devices). A system of connections is obtained by drawing G lines, each line intersecting the $(G-1)$ others, and "walking" from any origin, so as to never take two consecutive steps on the same line. There are many, but less than $N!$ such walks which pass through all points. A method certain to yield a successful connection system, when G is prime, is to make a list of the group numbers, $G = 0, 1, 2 \ldots (G-1)$, according to the order of the bars belonging to them, as follows: Start with any group, e.g., 0, and make additions modulo G, first by adding 1 G times, then

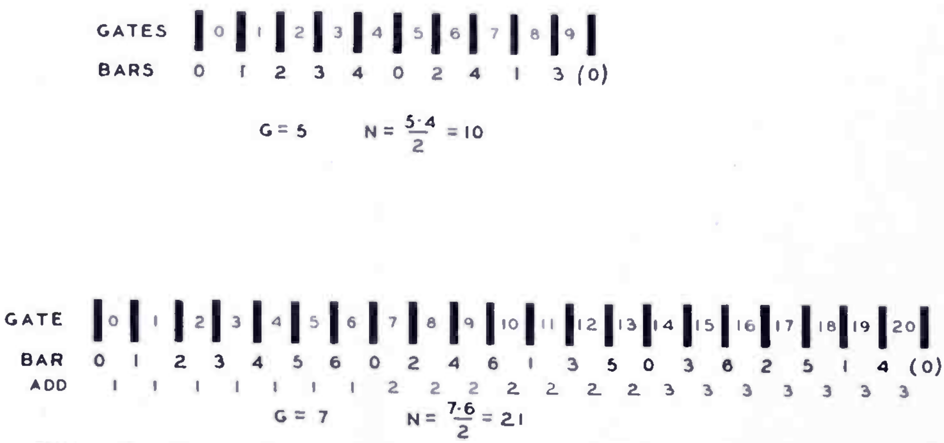


Fig. 25—Connections of bars according to single-family system.

2 G times, then 3 G times, etc., up to $\dfrac{G-1}{2}$. All $\dfrac{G(G-1)}{2}$ bars will have thus been numbered and each group is a neighbor with all others, once and once only. Figure 25 illustrates two examples of this procedure, one for $G = 5$ and one for $G = 7$. The second example is also illustrated on the diagram of Figure 22. The single family of groups system is more economical — asymptotically by a factor of two — than the two-family groups system as mentioned before. However, it is inconvenient in most circuit applications in that the selection of one group influences the selection of the second.

Area selection of elements is obtained by two systems of straight gates at an angle to each other through which electrons pass in series. In the tube described in this paper, there are two systems of straight parallel bars normal to each other. It is possible also to form the selecting matrix with a coaxial pile of rings and equiangular straight bars parallel to the axis, or by two concentric sets of parallel helical bars rotating in opposite directions. In fact, such geometries were used in the early tubes. It is clear that the number of windows E created by the intersections of the gates is simply the product of the number of gates in each direction, while the total number of groups L, i.e., controlling leads, is the sum of the leads used in each direction, $L = G_1 + G_2$. For the case of the two square two-family systems of $N = G^2/4$ gates each, the total number of windows is:

$$E = \left(\frac{L}{4}\right)^4 \tag{7}$$

The merit ratio between the number of windows or selected elements and controlling leads is, therefore, $L^3/256$, showing that the advantage grows tremendously with size. For $L = 16$, E is 256 but for $L = 128$ the number of elements, $E = 1{,}048{,}576$, surpasses a million. Similar compound formulas may be derived for the single family system.

The fourth power relation comes about naturally from the four bars limiting each window. Each electron passageway is equivalent to a four-control grid tube. Of course, to a higher number of control electrodes corresponds a still higher power relation. With n control electrodes for each electron channel, the relation between the number E of selectable elements and the control leads L is:

$$E = \left(\frac{L}{n}\right)^n \tag{8}$$

Successive rows of aligned parallel bars have the particular restriction to the combinations of connections mentioned before, arising from the fact that each bar is adjacent to two gates. It turns out that, in spite of this, connections yielding the full advantage according to relation (8) are possible. The only limitation is again in the choice of the geometrical location of the gates which is irrelevant in storage tubes. As an illustration, consider the case of the binary connection system, where

$$E = 2^n$$

elements are controlled by n pairs of control leads

$$L = 2n.$$

Each pair is one push-pull input, one or the other lead being the more positive. Figure 26 shows an example of 64 gates controlled by three successive rows of bars. Two binary inputs, or two pairs, are assigned to each row, so that in each row ¼ of the gates are open and ¾ closed. The opened quarters in successive rows are interlaced so as to leave only one open channel. On the example of Figure 26, the inner row provides a division in geometrical halves with the control of one pair and opposed quarters with the control of the next pair. The middle row again divides geometrically the gates into opposed eighths and sixteenths. However, in the last row such ordered geometrical division is not possible, because each bar is necessarily a neighbor to two gates. In fact, this is the only restriction to the obvious way of pyramiding control grids in halves, quarters, eighths, etc.

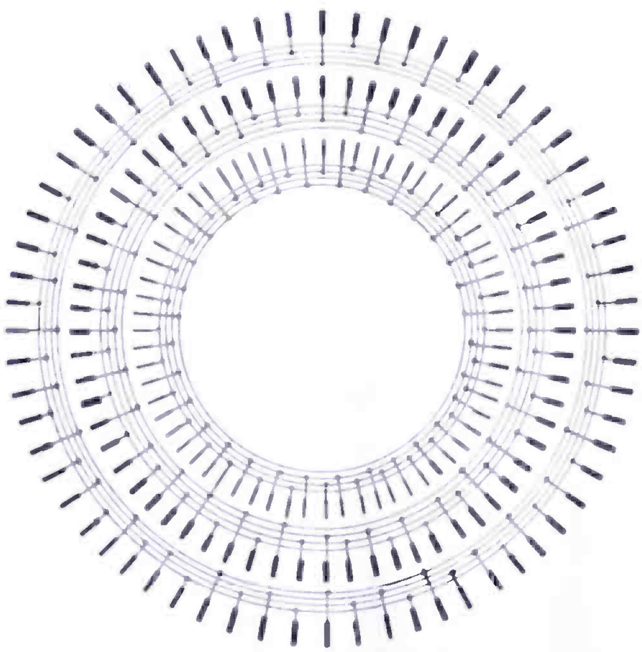


Fig. 26 — Connections of 3 successive rows of bars in binary system.

A binary address for the information is used often in a storage device in computing or information-handling machines, because of its inherent economy. A tube with direct binary address requires half as many successive rows of bars as there are digits. For a fairly large capacity, this number of rows may be appreciable. Instead, the tube can be made with a two-family group system, the numbers of groups P and Q being powers of two.* External circuit matrices are used to convert the binary address into the appropriate bases. This is

* G. W. Brown was the first to suggest the one and two family systems after analyzing the author's original conception of a purely binary system. See U. S. Patents 2,494,670 and 2,519,172.

done in the tube described in this paper. Such external converting matrices may be common to many tubes used in parallel, providing simultaneous access to identical addresses in all tubes. For this reason, the over-all economy of tube and circuits favors the simpler tubes with the group-of-two connection system.

Several windows may be opened simultaneously by the simple expedient of connecting several groups together. This is equivalent to several smaller tubes connected in parallel, and requires separate input and output channels for each separate region within the tube. It is also possible to open all windows simultaneously by making all bars positive, as is done for information-holding purposes as mentioned above.

INVESTIGATION OF ULTRA-HIGH-FREQUENCY TELEVISION TRANSMISSION AND RECEPTION IN THE BRIDGEPORT, CONNECTICUT AREA*

BY

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Summary—Information gained in the operation of experimental ultra-high-frequency television station KC2XAK in the Bridgeport, Connecticut area is presented. Over a period of a year the performance has been comparable with that of commercial very-high-frequency stations in economy and reliability.

Experience gained in home receiver installations, including the method of rating and analyzing reception, is described.

The dual automobile caravan assembled for field measurements throughout the service area is described and various types of measurements and analyses utilized up to the present time are outlined and reported. The information accumulated from this project is compared with that of other similar ultra-high-frequency projects conducted in other areas and found to be in very close agreement.

There are included comparisons of ultra-high-frequency and very-high-frequency picture ratings at 120 locations utilizing simultaneous transmissions of 217 watts effective radiated power at 67.25 megacycles and 13.900 watts effective radiated power at 530.25 megacycles from the Bridgeport station. The picture ratings for the very-high-frequency channel were superior to those of the ultra-high-frequency channel.

Compared with smooth earth theoretical propagation, in which terrain factors are not taken into consideration, it was found in the Bridgeport area that:

(a) At 10 per cent of the receiving locations the actual measured field intensity is below theoretical by 35.5 decibels or more.

(b) At 30 per cent of the receiving locations the actual measured field intensity is below theoretical by 29 decibels or more.

(c) At 50 per cent of the receiving locations the actual measured field intensity is below theoretical by 23.5 decibels or more.

INTRODUCTION

IN A previous paper[1] there was given a brief review of research in ultra-high-frequency (UHF) propagation and a description of KC2XAK, the UHF experimental television broadcasting station in the Bridgeport, Connecticut area.

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* Decimal Classification: R310 × R583.16.

[1] R. F. Guy, J. L. Seibert and F. W. Smith, "Experimental Ultra-High-Frequency Television Station in the Bridgeport, Connecticut Area", *RCA Review*, Vol. XI, No. 1, p. 55, March, 1950.

This paper contains data gained from station operations and field measurements which have been carried on during 1950.

KC2XAK Station Operation

Regular test pattern and picture program transmission was inaugurated on January 11, 1950, five days per week, from 9:00 A.M. until late evening. This schedule is still being followed on Tuesdays through Saturdays, additional hours being added as required for special phases of the investigation.

Program transmissions have been conducted about 340 hours per month throughout the year.

Inasmuch as the Bridgeport-Stratford station was intended to be a typical plant such as might be located in the future in many United States cities, its planning, construction, and operation, have conformed to the routine established for commercial stations. The apparatus was factory built to commercial standards, was delivered to the operating company engineers who built the station, and a small staff consisting of only two technicians and a caretaker, was assigned to operate it.

Emphasis has been placed on maintenance of fixed operating schedules and reliability with good apparatus maintenance and exacting performance. Only a few minor apparatus changes were made by the headquarters staff engineers since completion of the plant. The KC2XAK UHF transmitting apparatus may be considered to be nearly in the same category as commercial apparatus of standard types in economy, reliability and fidelity. During some months a perfect record of program continuity was experienced without failure of any kind.

The circuit adjustments for the TTU-1A transmitter were equal in ease of handling and stability to those of commercial very-high-frequency (VHF) transmitters. Because of the high operating frequency and the nature of the circuits, the UHF transmitter requires less special circuit adjustment for the broad band operation required in television. Tube performance has been very good.

a. Frequency Stability

The frequency stability of the KC2XAK transmitter has been very satisfactory as evidenced by the frequency measurements which follow. Of particular interest is the close correlation between sound and picture frequencies. The measurements shown are the number of cycles by which the carriers deviated from the assigned frequencies which are 4.5 megacycles apart. Special controls to insure these close relationships were built into the determining control circuits to maintain the 4.5-megacycle carrier separation in order that intercarrier type receivers could be used.

Date		AURAL		VISUAL	
		Riverhead	KC2XAK Monitor	Riverhead	KC2XAK Monitor
January	17			—4470	—4250 (before final adjustment)
	19	+ 180	+ 200	0	— 900
	20	— 300	0	— 300	— 800
	27	+ 300	+ 400	— 210	— 750
February	1			—2400	—2200
	11	+2400	— 200	+2250	— 400
	16	+1920	+ 100	+1620	0
	22	+ 90	+ 100	+ 180	+ 200
	25		— 200	— 690	— 300
	28	+ 480	+ 300	+ 240	+ 400
March	1	+ 900	+ 700	+ 690	+ 700
	2	+ 900	+1400	+ 600	+ 600
	3	+ 300	+1100	0	— 100
	4		+1000	— 300	+ 200
	7	— 540	+ 600	— 990	— 300
	8	— 600	0	— 900	0
	9	+2820	0	+2400	0
	10	+ 90	— 500	— 150	—1100
	11	+ 90	— 300	— 210	— 800
	14	— 600	+ 300	— 840	— 700
	16	+ 690	+ 700	+ 390	+ 400
	17	+ 840	0	+ 600	0
	18	+ 600	+ 300	+ 300	— 100
	21	+1200	+ 800	+ 600	+ 400
	22	+ 510	+ 400	+ 180	+ 100
	23	+ 300	+ 200	— 300	— 50
	24	+ 360	+ 300	— 150	— 200
	25	+ 990	+1000	+ 450	+ 500
	28	+ 300	+ 400	— 60	— 200
	29	+ 600	+ 800	+ 120	+ 500
	30	— 210	— 300	— 450	— 600
	31	+ 60	— 100	— 180	— 300
April	1	— 90	+ 200	— 360	— 200
	4	—1500	—1200	— 990	—1000
	5	+ 240	+ 400	— 210	+ 100
	6	+ 360	+ 800	+ 60	+ 400
	7	— 450	+ 100	— 810	— 300
	8	— 300	+ 400	— 600	— 500
	11	— 340	— 300	— 930	— 600
	12	— 120	+ 200	— 420	— 200
	13	— 300	+ 200	— 660	— 200
	14	— 720	— 200	—1200	— 450
	15	+ 300	+ 300	— 150	— 200

Date		AURAL Riverhead	AURAL KC2XAK Monitor	VISUAL Riverhead	VISUAL KC2XAK Monitor
	18	— 300	— 350	— 660	— 800
	19	+ 60	— 300	— 300	— 700
	20	— 270	— 200	— 630	— 400
	21	— 300	— 400	— 300	— 400
	28	— 180	— 200	— 540	— 200
May	9	+ 360	— 100	— 90	— 100
	16	+ 420	+ 700	+ 60	+ 100
	25	— 540	— 300	— 900	— 300
	31	+ 180	+ 200	— 330	+ 100
June	8	+ 360	+ 200	— 90	— 200
	15	+ 300	+ 200	— 60	— 200
	22	—1200	—1100	—1110	—1000
	24	+ 300	+ 200	+ 300	+ 200
July	4	— 420	— 200	— 570	— 200
	14	— 150	— 50	— 270	0
	22	+ 240	+ 50	+ 210	+ 50
	28	— 150	— 50	— 300	— 50
August	5	— 300	— 200	— 360	— 200
	11	0	+ 100	— 360	+ 100

KC2XAK operates on 530.25 megacycles for the visual carrier, 534.75 megacycles for the aural carrier.

The accuracy of the Riverhead frequency monitoring service at these frequencies is plus or minus 150 cycles. The KC2XAK frequency monitor is an experimental model and has been described.[1]

b. Program Feed From WNBT to KC2XAK

In order to evaluate the performance of the system represented by the KC2XAK transmitter, the radio path and the UHF receivers in the Bridgeport-Stratford service area, it is very desirable that the test pattern and pictures fed into the KC2XAK transmitter be of high definition and free of transients and noise. If mediocre test patterns or pictures are used to modulate the transmitter, the received pictures will also be mediocre, regardless of how faithfully the UHF system performs. The result would be inaccurate reporting of the performance of the UHF system.

The pictures and sound used to feed KC2XAK are those used by station WNBT, New York which operates on Channel 4 with antenna atop the Empire State Building. Early in this experimental program the pictures were received on Channel 4 and demodulated in a receiver at Bridgeport, and thence fed into the input of the UHF transmitter.

Thus the pictures from the studio were modulated at WNBT, demodulated at Bridgeport, again modulated at Bridgeport and again demodulated in the UHF receivers in the area.

Vestigial sideband transmission and reception introduces a small amount of distortion. The double transmission and reception consequently introduced an element of picture degradation which, while small, was undesirable in a carefully controlled field test of UHF television transmission. Occasional interference received on Channel 4 made the feed to the UHF transmitter poor at times. It therefore was desirable to take suitable steps to obtain a more perfect input picture, uninterrupted by Channel 4 interference.

Experiments were conducted utilizing a relay link, operating on 2000 megacycles, the transmitter being located in the Empire State Building with the parabolic transmitting antenna located outside on the parapet at the 85th floor. At KC2XAK the parabolic receiving antenna was located at the 180-foot level of the tower. To minimize losses, a 7/8-inch coaxial transmission line was installed between the receiving antenna and the receiver in the transmitter building. The picture from the studio was fed directly into the 2000-megacycle relay transmitter in New York without going through the WNBT VHF transmitter. The sound continued to be provided by Channel 4 reception. The effective radiated power of the relay is 1216 watts, based upon a transmitting antenna gain of 27.85 decibels over a dipole.

The picture quality through this relay was and has remained excellent and reliable and the relay is practically free of external disturbances of any kind. The airline distance involved in the relay operation is 55 miles. The input pictures at KC2XAK are, as nearly as can be observed, as good as those leaving the studio, and the pictures at good UHF receiver locations are comparable to those through VHF stations at good receiver locations. The signal-to-noise ratio of the 2000-megacycle relay is between 26 and 30 decibels peak-to-peak signal to peak-to-peak noise measured at the video output circuit.

A request for special temporary authority to utilize this 2000-megacycle relay was filed with the Federal Communications Commission on January 19, 1950. The grant was received on January 25, and the relay was placed in operation on the same day. The Channel 4 receiving system remains intact as a program channel standby facility.

c. Apparatus Changes Since Installation

Since the transmitter was originally placed in operation, various minor improvements have been made as might be expected with a prototype transmitter, but in no instance has it been necessary to make any changes of consequence. The most significant minor change may be of interest.

As initially installed, the frequency tripler stages which drove the aural and visual power amplifiers operated at an input frequency of one third of the output carrier frequencies, 176.75 megacycles and 178.25 megacycles, which fall within Channel 7. Extremely minute fields from the circuits produced interference to reception of Channel 7 within a fraction of a mile, despite double shielding and other precautions provided in the transmitter to confine such fields. Considerable time was devoted to efforts to eliminate these minute external fields with incomplete success. Elimination of this problem was accomplished by changing the triplers to doublers and correspondingly changing preceding multipliers from doublers to triplers.

Home Receiver Installations

The evaluation of the service potential of a UHF television station requires, among other things, observations of service in private homes with typical receivers. In this project there was included a program for the manufacture of 50 UHF converters for attachment to VHF receivers, and also 50 complete receivers equipped to receive VHF or UHF. These receivers and converters were manufactured specially for the Bridgeport-Stratford project.

a. The VHF-UHF Receiver

This complete receiver consists of a standard television receiver to which there was added internally, a specially built UHF tuner. Described in simple terms, the tuner consists of a high-pass 500-megacycle filter to reject lower frequencies, a radio-frequency amplifier, a mixer-oscillator, a 132-138-megacycle intermediate-frequency amplifier and a second mixer-oscillator which produces the 21-27-megacycle signals which are fed into the standard receiver intermediate-frequency circuits. This UHF-VHF receiver has been described elsewhere.[2]

b. The UHF Converter

The UHF converter is a unit designed to be externally connected to a standard VHF television receiver. In design, this unit differs from the UHF tuner previously described mainly in that it converts the KC2XAK signals to either Channel 12 or 13 and delivers them to the VHF receiver terminals so that these signals may be tuned in

[2] T. Murakami, "An Experimental Ultra-High-Frequency Television Tuner", *RCA Review*, Vol. XI, No. 1, p. 68, March, 1950.

exactly as a Channel 12 or 13 station would be. This UHF converter has been described elsewhere.[3]

c. Selection of Homes

This phase of the problem involved the establishment of a "bank" of names of individual home owners who resided in areas which would be of interest and who were interested in cooperating in the project. To facilitate the establishment of receiver installations where they would contribute the maximum amount of information to the overall project, a bank of several hundred names was sought and compiled.

Maps were prepared showing the areas within which KC2XAK might provide service based upon theoretical considerations and experience gained elsewhere. The listed potential test locations were then shown on the map by color coded pins. Pins, also color coded, showed the homes in which receivers or converters were installed. In general, an effort was made to distribute the receivers uniformly throughout the area. The locations of homes actually used are shown in Figure 1. The map does not include homes in which installations were tried and removed because of the absence of usable field intensities.

The rather considerable amount of publicity which attended the purchase of the station property, the zoning hearing and other advance activities stimulated a large amount of local interest which resulted in many voluntary offers of participation in the making of observations in private homes.

In general, an effort was made to utilize homes in which the residents had had some previous experience in viewing television reception, but this was by no means a requisite. The home picture ratings used in the analyses herein were made independently by engineering personnel during visits to the homes. This method of rating was considered necessary to insure uniform and accurate standards of ratings and conclusions.

Experienced service men installed and serviced the receivers. They were carefully selected for technical knowledge, diplomacy and experience. An installation crew consisted of two men and a service automobile equipped in the usual manner. During the initial phases of the home installation work, three crews were employed.

In about half of the homes of the volunteer participants there were no existing television receivers or antennas, and it was therefore necessary to install not only a UHF antenna but also a VHF antenna

[3] W. Y. Pan, "Some Design Considerations of Ultra-High-Frequency Converters", *RCA Review*, Vol. XI, No. 3, p. 377, September, 1950.

to permit reception of the commercial stations as well as KC2XAK. By being able to receive VHF stations, the installation could contribute more information.

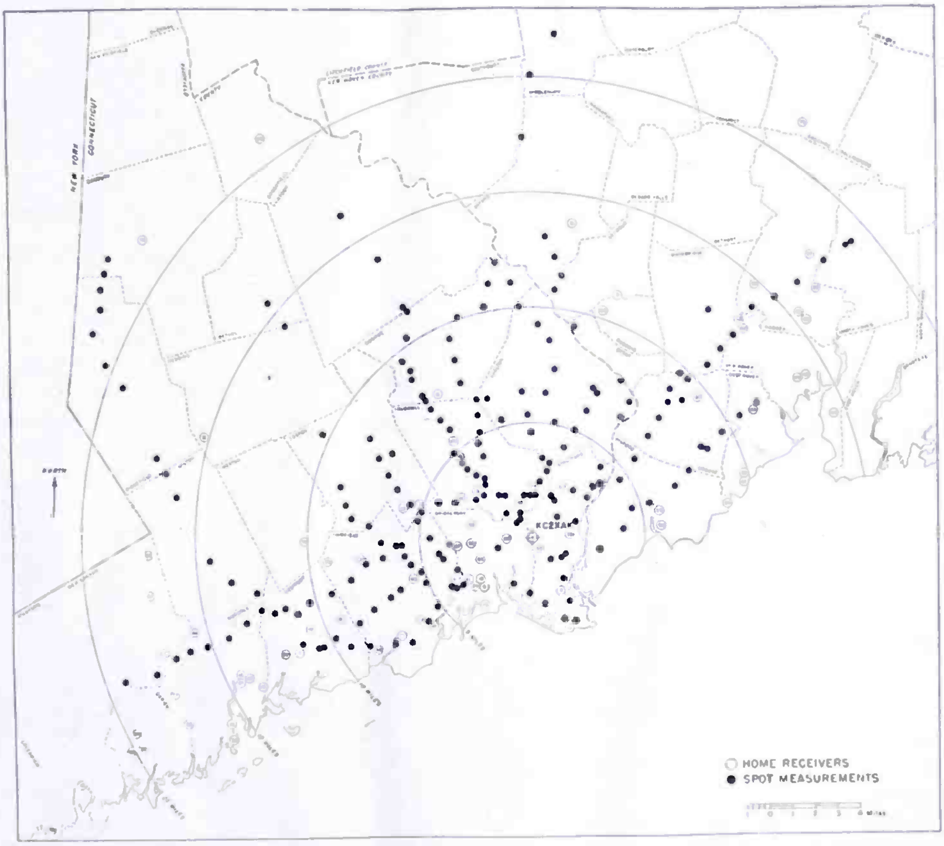


Fig. 1—Map of Bridgeport area showing locations of home receivers and spot measurements.

d. Receiving Antenna Types

Four distinct types of receiving antennas were used for UHF reception. These consist of the following:

1. Fan dipole.
2. Rhombic.
3. Stacked vee.
4. Parabola.

Fan dipole—The fan dipole has unity gain, represents the cheapest type of UHF antenna to manufacture, is the lightest in weight, the least objectionable as regards appearance on a roof and has the least directivity of any type used. However, its lack of gain limits its use to installations within a five- or six-mile radius. It is substantially useless in eliminating reflections and ghosts.

Rhombic—The rhombic antenna has a gain of 3.6 decibels. This antenna provides noticeably better pictures than the fan dipole where the latter is inadequate. However, the best feature of this antenna has been its sharp horizontal directivity pattern. Where ghosts were experienced, this antenna completely eliminated them in all but one case which was considerably improved. The rhombic antenna was not generally used due to its higher cost of manufacture, additional weight and the relatively long time required for assembly and mounting. Figure 2 shows the design.

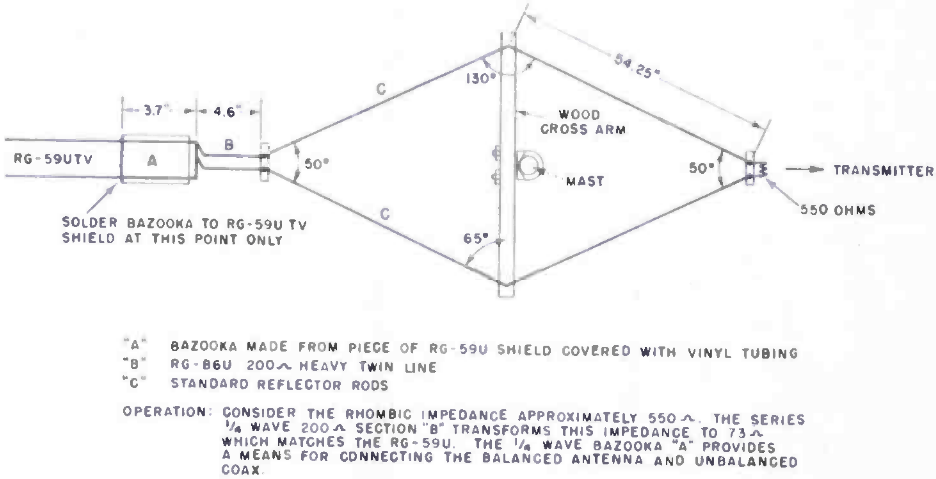


Fig. 2—Schematic diagram of rhombic antenna.

Stacked vee—The stacked vee antenna has a gain of 5.7 decibels and was the most generally used because it is cheap to construct, easy to install, has substantial gain and has good directivity characteristics. This antenna will also perform well on VHF if the elements are cut to 52 inches. This antenna, in common with the rhombic, has a sufficiently large aperture to render it relatively uncritical as to placement in obtaining the best local location. This antenna has been preferred above all other types on the KC2XAK project. Figure 3 shows the design.

Parabola—The parabola antenna has a gain of 7.5 decibels, the highest of any available. It is the most expensive to manufacture and the heaviest. In some instances, it has been out-performed by the Stacked Vee with a larger aperture. This is ascribed to the sharp vertical directivity of the parabola. It is not always possible to find the optimum location for the parabolic antenna. Its directivity pattern is very broad in the horizontal plane. Figure 4 shows the design.

The breakdown of antenna types used is shown below.

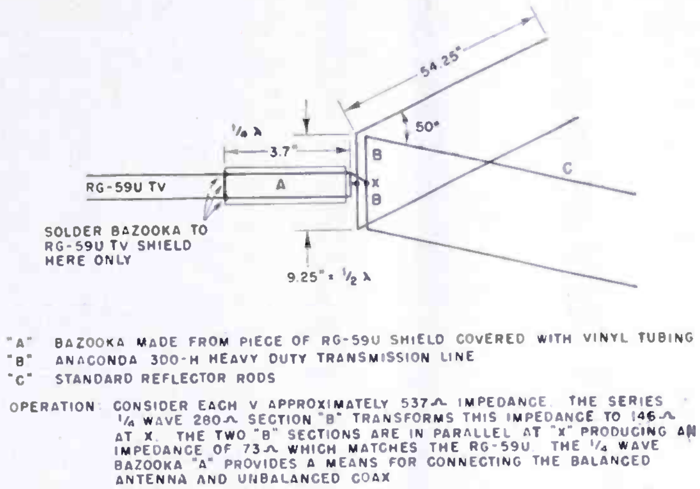
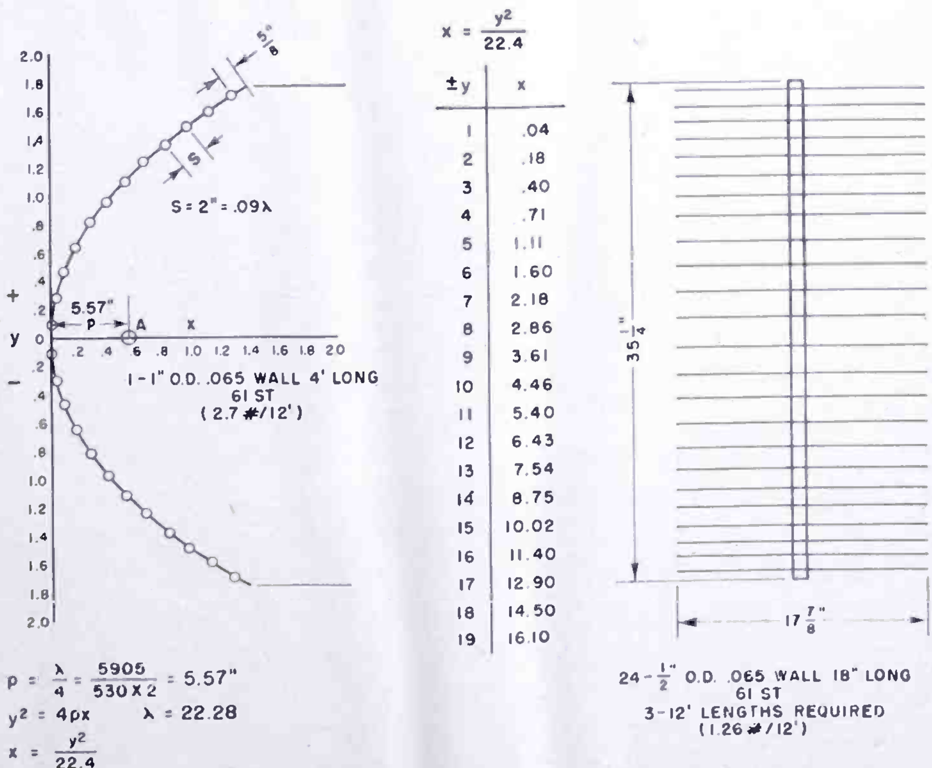


Fig. 3 — Schematic diagram of dual vee antenna.

Stacked vee	53 per cent
Fan dipole	29 per cent
Parabola	12 per cent
Rhombic	6 per cent

e. Importance of Good Receiver Installations in UHF

During the UHF field test at WNBW, Washington, the transmission lines used in homes to connect the antenna to the receivers consisted



±y	x
1	.04
2	.18
3	.40
4	.71
5	1.11
6	1.60
7	2.18
8	2.86
9	3.61
10	4.46
11	5.40
12	6.43
13	7.54
14	8.75
15	10.02
16	11.40
17	12.90
18	14.50
19	16.10

Fig. 4—Dipole and cylindrical parabolic reflector.

of "bright picture", flat, unshielded 300-ohm type. This line was vulnerable to losses when it became wet from rain or sleet. In addition, the routing of the line over roofs, gutters, around pipes, etc., made it vulnerable to standing waves or reflections, loss of energy, and interference from sources of noise. While the losses were quite low under ideal conditions, it was decided to use a different type line for the Bridgeport project. RG59U, a coaxial type, consists of a single inner conductor surrounded by molded insulation which, in turn, is surrounded by copper braid forming the second conductor. While this line has higher losses than the bright picture line, it is weather-proof, free of interference from nearby noise sources and can be routed conveniently without being affected by other objects in the immediate vicinity. For this reason it was selected for the Bridgeport project. Since this line has 72 ohms impedance, unbalanced, the receiver and converter input circuits were designed to terminate correctly that impedance. A balun is used to match balanced antennas to unbalanced lines and receiver input circuits, impedance transformation being included when required.

Contrary to the smooth-earth theory, the service range of a UHF station is markedly smaller than for a VHF station. This became evident after the initial receivers had been installed and observations completed. It was also evident from investigating the service potentialities of a UHF television system that a number of small receiver installation losses accumulated into a very significant total when converted into the number of transmitter kilowatts and corresponding dollars of cost required to compensate for them. For example, a receiving antenna with 3 decibels less gain than could reasonably be obtained, a transmission line loss 3 decibels higher than necessary and a receiver noise factor 4 decibels higher than necessary, add up to 10 decibels. To compensate for these losses, the transmitter power would have to be increased *ten times*. At Bridgeport this would mean an effective radiated power of 139, instead of 13.9 kilowatts. Expressed in different terms, such as service radius, it could mean that, with such receiver installations throughout, the service area would be but a fraction of the potential area.

Since the purpose of this project was to evaluate the service potential of a UHF television broadcasting system, it was deemed essential to have knowledge and control of all of the factors involved. Accordingly, each home installation having less than optimum picture ratings was revisited by an engineering team, to insure that the conditions were optimum for reception. It should be borne in mind that careless installation and the use of receivers with characteristics below the tolerances for good practice in alignment and noise factor, exact a penalty in service area shrinkage or spotty reception. It is unlikely that the standards for home installations maintained on this project would be equalled in commercial practice so in that respect the analyses reported herein may be optimistic.

The work at each receiver installation revisited consisted of three primary phases, as described in sections f, g, and h which follow:

f. Measurement of Receiver Noise Factors

Shipment and handling may cause changes in the performance of the receiver components and the overall receiver itself. To eliminate the unknown variable of receiver noise factor, it was deemed necessary on this project to measure in the field all UHF receivers and converters used, label them and correct those found to be outside the factory tolerances.

During the period from April 20 to May 9, 1950, the television receivers and converters used to check reception of KC2XAK were brought in and noise measurements were made on them, using facilities specially established for the purpose.

The noise factor limits established by the receiver design engineers were 18 decibels for the receivers with the built-in tuners, and 22 decibels for separate converters. Some of the units which were borderline cases were brought within limits by substituting tubes or crystals. A few other units which failed to respond to this treatment were sent to the factory for correction.

g. Antenna Locations

A careful reinvestigation was made of the location of receiving antennas where possible, because frequently large changes in field intensity take place within distances of feet or even inches. With few exceptions it was found that the improvement due to this reinvestigation was very small or zero, because of the care which had been exercised on the original search.

h. Use of Low-Loss Transmission Line

The loss of RG59U is about 9.5 decibels per 100 feet at 530 megacycles. Other forms of lines of fairly recent development provided much lower losses per 100 feet but had other characteristics which were not sufficiently well-known to warrant their use. In homes having short transmission lines the reduction of loss through substitution would be small and in some cases negligible. But in other homes the expected improvement appeared promising. Accordingly, transmission lines of the following types and characteristics were investigated.

Type	Nominal Impedance (Ohms)	Loss per 100 feet at 530 megacycles (Decibels)	Cost per foot (Cents)
Flat Twin	300	3	3
Amphenol tubular Twin for receiving	300	3	7
Amphenol tubular Twin for transmitting	300	2.5	9.6
ATV225	225	7.6	21
K-111	300	8.0	21
RG59U	75	9.5	12.8

Fig. 5 — Types of transmission lines used for receiver installations.

Figure 5 is a photograph of the lines enumerated. The losses indicated for each transmission line type assume unity standing-wave-ratio. Losses will naturally increase as the standing-wave-ratio departs from unity. In multichannel receivers such losses may be as high as 2 decibels above the rated loss. In addition, losses may occur if unshielded lines are improperly routed.

The costs shown were the prevailing prices during the summer of 1950.

Tests were made with amphenol tubular twin line designed for transmitting purposes. Baluns were built to match this line to the receiver. Initial tests took place during a period when the line was covered with ice for two days during which little or no changes were noted in the picture rating or the receiver input voltage. Upon the conclusion of this test, amphenol line having the same general characteristics but designed for reception rather than transmission was tried and found to give the same results, with a few exceptions where placement of the line was fairly critical. In this respect, it was superior to the flat bright-picture wire and represented the best line available for the purpose. Upon the conclusion of these tests, the amphenol tubular receiver line was installed at installations having less than optimum picture ratings. With the usual line length of about 55 feet, some cases resulted in upgrading the signal by one of the four steps of rating ("poor", "fair", "good", and "excellent").

i. Effect of Foliage

It is not yet possible to give comprehensive data on the effect of tree foliage inasmuch as this project had not advanced sufficiently to obtain adequate measurements with the trees defoliated. But the effect was observed to be quite marked when the signal had to traverse certain wooded areas. At one receiver located in Waterbury, twenty-five miles from the transmitter, a picture was received during the winter months which was only fair because of insufficient signal intensity. With the growth of foliage in the late Spring, the picture disappeared and could not be restored by rechecking the installation. At another installation in Norwalk, fifteen miles distant, a noise-free picture with the trees defoliated was degraded to good or fair when the foliage appeared. Some other instances were noted where these effects occurred to a lesser degree due to fewer trees, or because the picture was so far above marginal excellence that a reduction of signal strength could be tolerated before the effect became noticeable. The average height of the receiving antennas above ground is about 45 feet.

j. Receiver Tuning

Many service calls have been prompted by frequency drift in the experimental receivers and converters, most observers having remarked on the necessity of having to retune to maintain satisfactory sound. The separate converter was somewhat better in this respect than the combination receiver. This difficulty of maintaining correct receiver sound channel tuning is largely due to the need for precise tuning in the system used in television sound. Also, the Bridgeport units do not contain very recent improvements in oscillator techniques.

k. Facts of General Interest Concerning Receiver Installations

Upon arriving at a home for an installation, the two man crew divided the work. One man installed the receiver in the preferred location while the other man went to the roof and dropped a transmission line temporarily to the man indoors. Inter-phones were used for communication so that as the man on the roof explored the area with a fan dipole mounted on a 12-foot pole, the other man observed the picture on the receiver. When the optimum location for the fan dipole was determined by the man at the receiver, the antenna location was noted. This operation was repeated in various areas on top of the house. In the majority of cases, it is found that where the signal is received directly from the transmitter, rather than by reflection, the antenna position is less critical and may permit an attachment to the chimney in conventional VHF fashion.

More often than not, it is found that the fan dipole does not produce sufficient pick-up. In such cases, a stacked vee is substituted. If this proves inadequate, the parabola is substituted. If the improvement is insufficient, an attempt is made to increase the height of the antenna by adding an extension to the pole. Where signals are received by reflection, the optimum position of the antenna may be very critical and the optimum height may or may not be high. A distance of a few inches vertically or horizontally or both may make a large difference in picture quality. This effect is very much more pronounced on UHF than on VHF.

Where UHF and VHF antennas are both being installed, it is customary to explore for optimum location with the UHF antenna since it is much more critical to the location. Ordinarily, the horizontal position of the VHF antenna is secondary in importance to its height. In many installations the UHF antenna is mounted on the VHF supporting pole because there is no choice. In other cases, where the VHF pole location is unsatisfactory for UHF, another pole is mounted in the best position for UHF. Upon completion of the search for the optimum antenna location, and assuming that a usable picture is received, the antenna is mounted and the transmission line is permanently routed to the receiver.

In strong signal areas close to the transmitter, indoor antenna installations have been tried but with only fair results. In any UHF indoor antenna installation the antenna must be located in such a manner that persons cannot walk in front of it. Unless this is done the signals will fluctuate, often falling to unusable values, and reflections appear in the pictures. Furthermore, operating the house wiring switches occasionally changes reception due to changes in UHF currents flowing in the associated circuits. Where rooftop antenna locations are not available even in high signal strength areas, best performance is obtained by mounting the antenna outside a window, rather than indoors.

Multipath reflections have not been as troublesome on UHF as on VHF and where encountered have been eliminated with a few exceptions by the use of the most directive types of antennas. The greater directivity obtainable on UHF antennas of practical size makes it possible to select a signal coming from a single direction to better advantage than with VHF antennas. A UHF antenna of comparable mechanical size can be made to have better directional characteristics than a VHF antenna, as might be expected.

No interference from diathermy has been received or reported in the KC2XAK service area, despite the fact that one location is directly across the street from a medical building with active diathermy equipment in use throughout the day.

Although ignition interference appears on VHF receivers in the Bridgeport area, it is barely noticeable on UHF to an experienced observer and would probably go unnoticed by the layman viewer. Ignition interference has been observed only under the most unfavorable circumstances on UHF from the local station.

Summer static produces no interference to UHF reception although local lightning flashes may momentarily produce impulses sufficient to cause a picture to roll a few frames from destruction of vertical synchronizing.

Heavy rain such as is encountered during a cloudburst will attenuate UHF signals noticeably as will a very intense snow storm. Receivers located so that trees intervene towards the transmitter, experience a substantial drop in signal intensity when the trees are wet from rain. At one location twelve miles distant from the transmitter, the voltage dropped 40 per cent, gradually recovering to normal after a few hours of sunshine.

The first UHF receiver installation was made on January 17, 1950, 12 miles from KC2XAK in a home at Westport, Connecticut. A stacked vee antenna with a rotator was installed 55 feet above the ground and connected to the receiver through ATV-225 twin shielded transmission line, 125 feet long, having a loss of approximately 9.5 decibels. The receiver was a Model 9T246 RCA unit, modified by a built-in converter. The description of this installation is included because it is typical of many others where steps were taken to improve the picture. The picture originally received was weak and very noisy. A rhombic antenna replaced the double V with similar poor picture rating. RG59U transmission line was then substituted for the ATV-225 type, both having a similar loss characteristic. No improvement was noted. The amphenol flat 300-ohm line was then installed, having a loss of approximately 3.75 decibels. As expected, a much better picture resulted with a rating of "fair".

However, the picture became very poor when this line was used in wet weather. Consequently, Amphenol twin-transmitting line was substituted, having a loss of 3 decibels. This line produced equally good pictures without noticeable vulnerability to wet weather as evidenced by voltage measurements at the receiver when the line was coated with ice and rain. A further improvement was noted in that the latter line was less critical of placement in routing. Subsequently, some modifications were made in the receiver which improved the noise factor by 6 decibels, producing a very good picture. Some months after this experience, when the foliage came out, the picture deteriorated considerably.

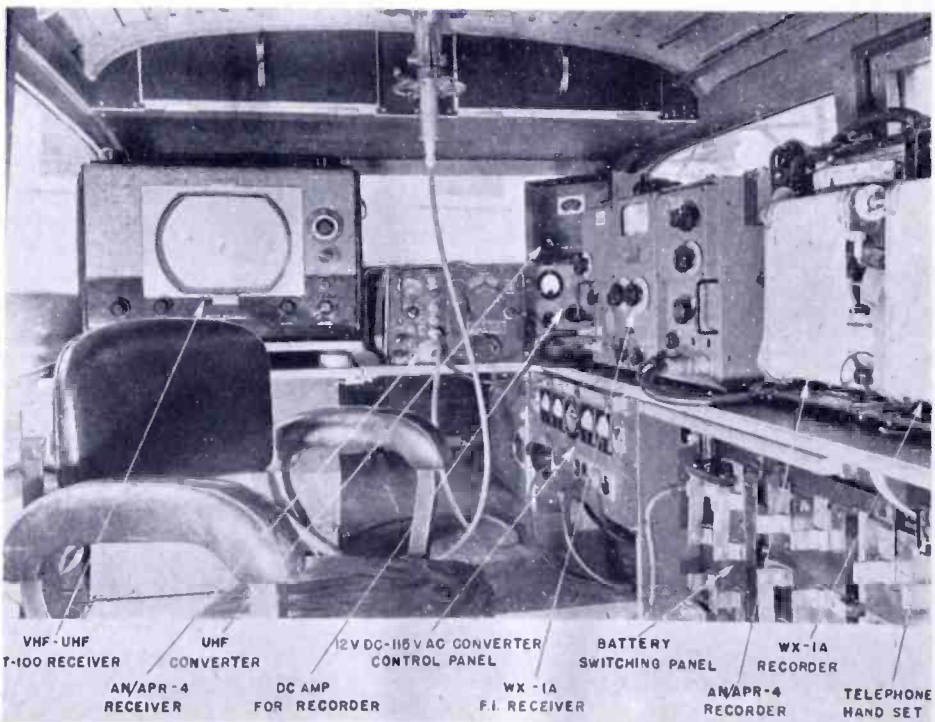


Fig. 6—Interior of station wagon showing mobile equipment used in making field measurements.

Ground Propagation Measurement Technique, Data and Analysis

a. Description of Measuring Equipment

All of the equipment used for mobile field measurements was built into a 1948 station wagon as shown in Figure 6. This equipment includes at the left an RCA T-100 receiver with built-in tuner for UHF. This was used for picture quality ratings for all spot measurements. Next is a modified AN/APR-4 receiver and specially built direct-current amplifier which is shown connected to one of the Esterline-Angus recorder units. This receiver employed a 10-foot elevation antenna mounted on the roof of the station wagon. Mounted on top of the direct-current amplifier is a selected UHF converter which is used in conjunction with the WX-1A Field Intensity receiver and a 30-foot elevation antenna. The output of this converter appears on Channel 12 to which the WX-1A receiver is tuned. The output of the WX-1A receiver is connected to the second Esterline-Angus recorder. Both recorder drives are coupled together and attached to the car speedometer drive mechanism. The gear ratio in the drive mechanism provides 8 inches of tape recording per mile of travel.

Two types of power supplies were required to operate this equipment. The WX-1A receiver required 6 volts direct current. The T-100, AN/APR-4, UHF converter, and direct-current amplifier required 115 volts alternating current. Two heavy-duty batteries were mounted behind the switching panel and each of these, as well as the regular car battery, was connected to a double-pole, double-throw switch. This switching scheme provided both a 6-volt supply for the WX-1A receiver and a 12-volt supply for a rotary converter used to generate the required alternating-current voltage. The controls for this converter are mounted on the panel to the left of the battery switches, and permit an independent adjustment of output voltage and frequency. A Leece-Neville generator satisfactorily replaced the standard battery charging generator to provide a higher rate of charge.

A mobile telephone was installed in the field car so that it was possible to keep in touch with the transmitter at all times and to make frequent checks on transmitter power, modulation, and barometric pressures for resetting of the field car altimeter, etc.

The antennas used were of the single-bay super-turnstile type connected to the receivers with RG8-U cable. This type antenna was selected because of its omnidirectional pattern, eliminating the necessity for frequent reorientation of the antenna. The 10-foot elevation antenna was mounted on the roof of the station wagon and the 30-foot elevation antenna was mounted on the top section of a 40-foot aluminum extension ladder which was hinged on a truck bed as shown in Figure 7. With this arrangement the upper antenna height could be varied from 30 feet to 46 feet above the ground as desired. The station wagon was normally towed by the truck, using a standard tow bar.

b. Methods of Measurement

After a route for measurement was selected, calibration of the equipment was checked with a Hewlett-Packard Model 610-A Signal Generator. The starting point for most radials was about one mile from the transmitter. At this point the field intensity was measured using the 10-foot antenna and AN/APR-4 receiver for one setup and the 30-foot truck antenna and converter—WX-1A receiver combination for the second setup. The signal was measured for 30- to 46-foot heights in 2-foot increments. The recorders were then coupled to the speedometer drive and the mileage and elevation were noted. The caravan proceeded along the measuring route for a distance of approximately one mile, taking continuous recordings for both 10- and 30-foot antenna heights. At this point all spot measurements were repeated as outlined above. This procedure was followed for the entire route.

Fig. 7—"Caravan" used in making field measurements.

Whenever obstructions were encountered under which the 30-foot antenna could not pass, the ladder was lowered and then raised again as soon as the obstruction was cleared. It was necessary to lower the ladder often, in some instances four or five times in a one-mile interval. Many of the routes were so covered with overhanging branches or wires that continuous recordings with the 30-foot antenna were impossible. In these cases, one-mile interval measurements were made. A crew of six men was required for the caravan, four on the truck to manipulate the antennas and two in the station wagon to make the measurements and record the data.

In general, two types of measurements were utilized. First, continuous and spot measurements of the aural carrier were made using the 10-foot and 30- to 46-foot antenna heights. The second type was spot measurements using a 30-foot antenna, but in this case the antenna was moved back and forth to determine the maximum and minimum signal for both visual and aural carriers and to observe the standing-wave pattern at each location. Figure 8 is a map of the Bridgeport area showing all of the routes traveled to make these measurements. Approximately three hundred miles of roads were traveled. Tape recordings were made over approximately one hundred fifty miles, and over five hundred spot measurements were made.

All field intensities were computed on the basis of the peak visual signal. Since the aural carrier was used for some measurements and the visual carrier with test pattern modulation was used for others, it was necessary to apply certain correction factors to the measured values in order to convert them to the equivalent of peak visual signal.

c. Field Intensity Variations With Distance Compared With Theoretical Predictions

The measurements described in Section "b" were taken along the N-0° E, N-7.5° E, N-47° E, N-243° E, N-255° E, N-270° E, N-305° E, and N-326° E radials which are indicated in Figure 8. The results obtained are presented in Figures 9 through 16. In the upper portion of each of these figures, the field intensities obtained at various points along the radial have been plotted as a function of distance, as well as the theoretical field-intensity curves for each of the radials in question, computed on the basis of smooth-earth theory at 530 megacycles.[4,5] The nulls which should theoretically appear at distances close to the transmitter due to cancellation by the ground reflected wave are not shown. Also, no correction has been made in the theoretical prediction to take into account the effects of the vertical directivity of the transmitting antenna, which were small.

In the lower portion of each figure is shown the corresponding profile for each radial. It will be observed that in most cases there is a fair agreement with the irregularities indicated on the profile and the corresponding field intensity measurement for that sector. The

[4] K. A. Norton, "The Calculation of Ground-Wave Field Intensity over a Finitely Conducting Spherical Earth", *Proc. IRE*, Vol. 29, p. 623, December, 1941. See also FCC Report 39920, March 18, 1940.

[5] F. W. Smith, "UHF-TV Field Intensity Calculations", *Electronics*, Vol. 23, No. 10, October, 1950.

terrain and road distributions in the Bridgeport area are highly irregular. As a consequence, it was difficult to follow the radials exactly and a number of measurement points were slightly off the radial. These are indicated as circles on the radial plot.

It can be seen that a majority of the measured field intensities were well below those predicted theoretically for 530 megacycles by the smooth earth theory.

It is pertinent to analyze the field measurement data statistically to determine the extent to which the individual measurements deviated from the theoretical value as predicted by the smooth earth theory at 530 megacycles. It can be seen that in some cases, well within five

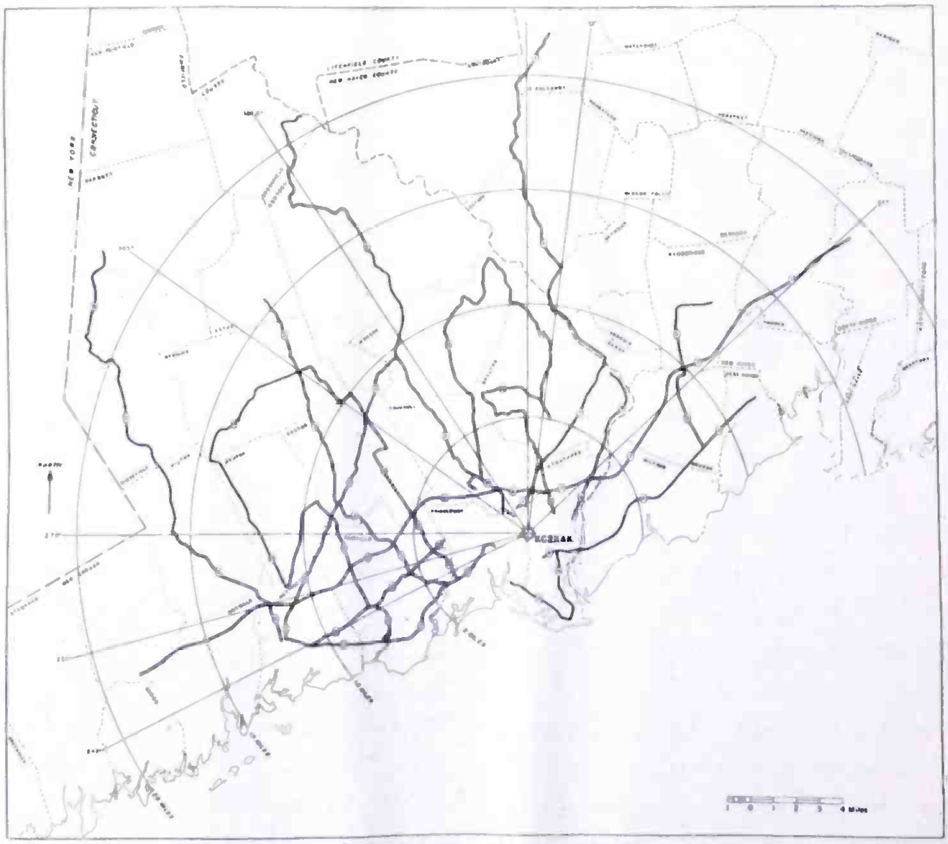


Fig. 8—Map of Bridgeport area showing radials and recording routes.

miles of the transmitter, the measured field intensities were 35 to 40 decibels lower than those predicted theoretically, illustrating the reductions in signal strength which can occur in regions of irregular terrain such as are encountered in the Bridgeport area.

The data in Figures 9-16 were combined and analyzed statistically, yielding the results shown in Figure 17. Here, the abscissa indicates the percentage of receiving locations having field intensities higher than the ordinate, and the ordinate indicates the extent of the deviation from smooth earth theory. It will be observed that at 10 per cent of the measurement points, the signal was below theoretical by at least 35.5 decibels. For 30 per cent of the locations, the signal was below theoretical by at least 29 decibels and for one half of the locations it was at least 23.5 decibels below theoretical.

In computing theoretical field intensities, the height above average terrain was based upon the 2- to 10-mile average as described in the

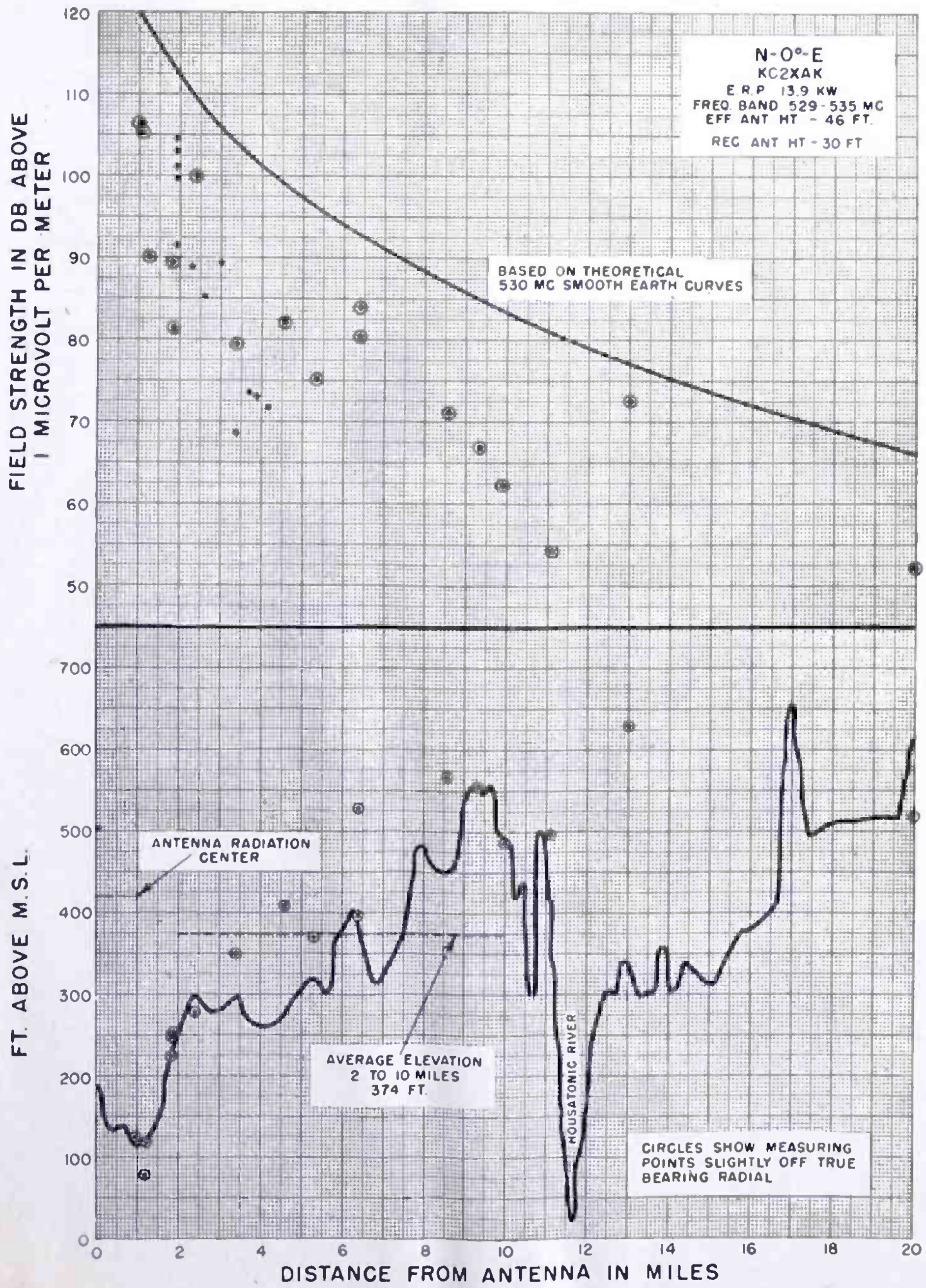


Fig. 9—Field intensity measurements, North radial.

FCC Standards of Good Engineering Practice for VHF Television. This is less than an ideal method but was adopted for lack of a superior method. An analysis was made which compared the same measured values with the theoretical values of the free-space field, which do not include terrain considerations or antenna heights. This analysis showed that the departure of measured values from theoretical values for 50 per cent of the locations was only 2.5 decibels below the value derived

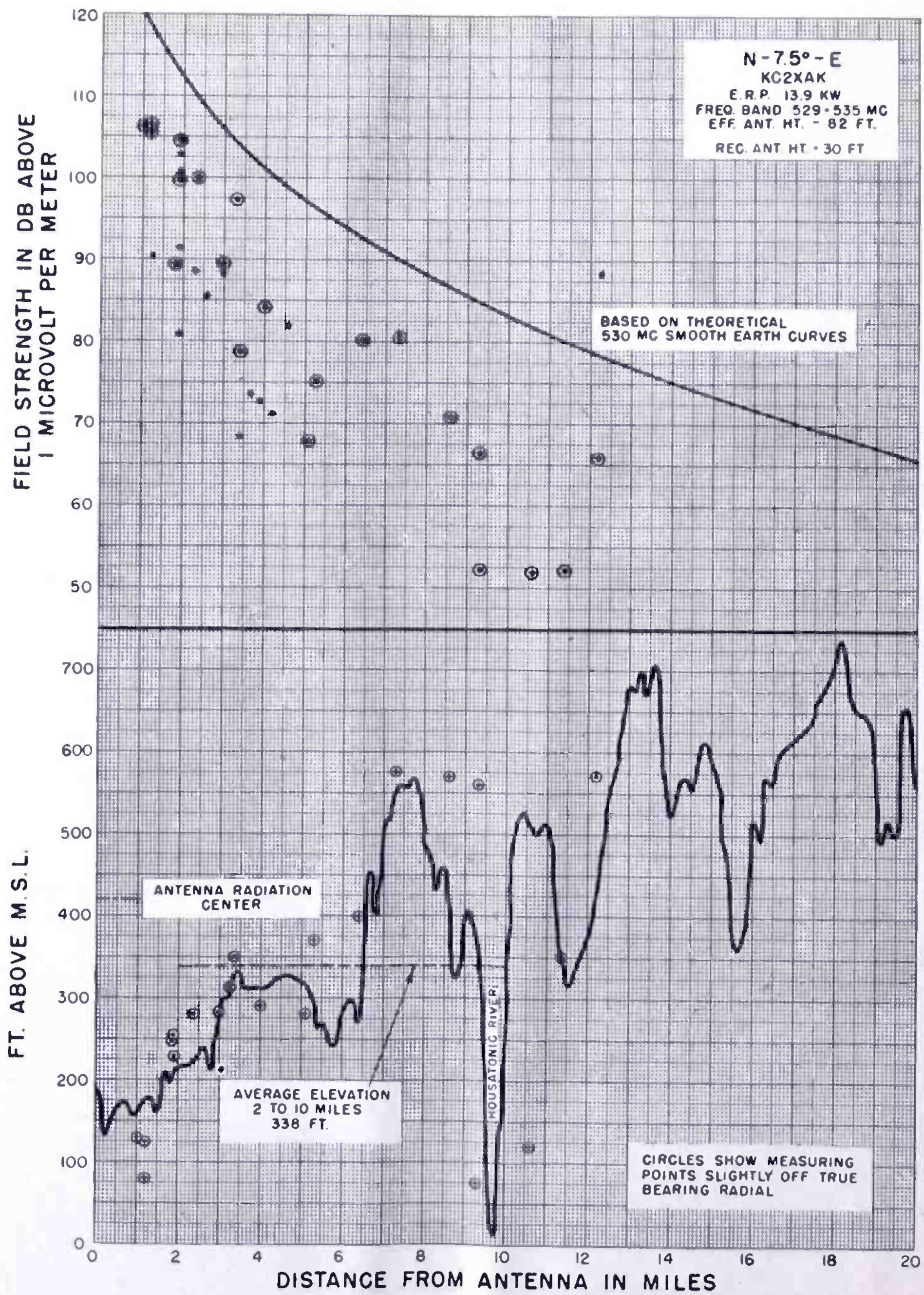


Fig. 10—Field intensity measurements, N-7.5°E radial.

from the use of the above mentioned 2- to 10-mile method and that the statistical distribution was approximately the same.

Figure 18 shows the departure from theoretical field intensities with distance for the Bridgeport project. It should be noted that the statistical deviations from the theoretical are relatively constant with distance.

In Figure 19 is shown a plot of the maximum field intensities

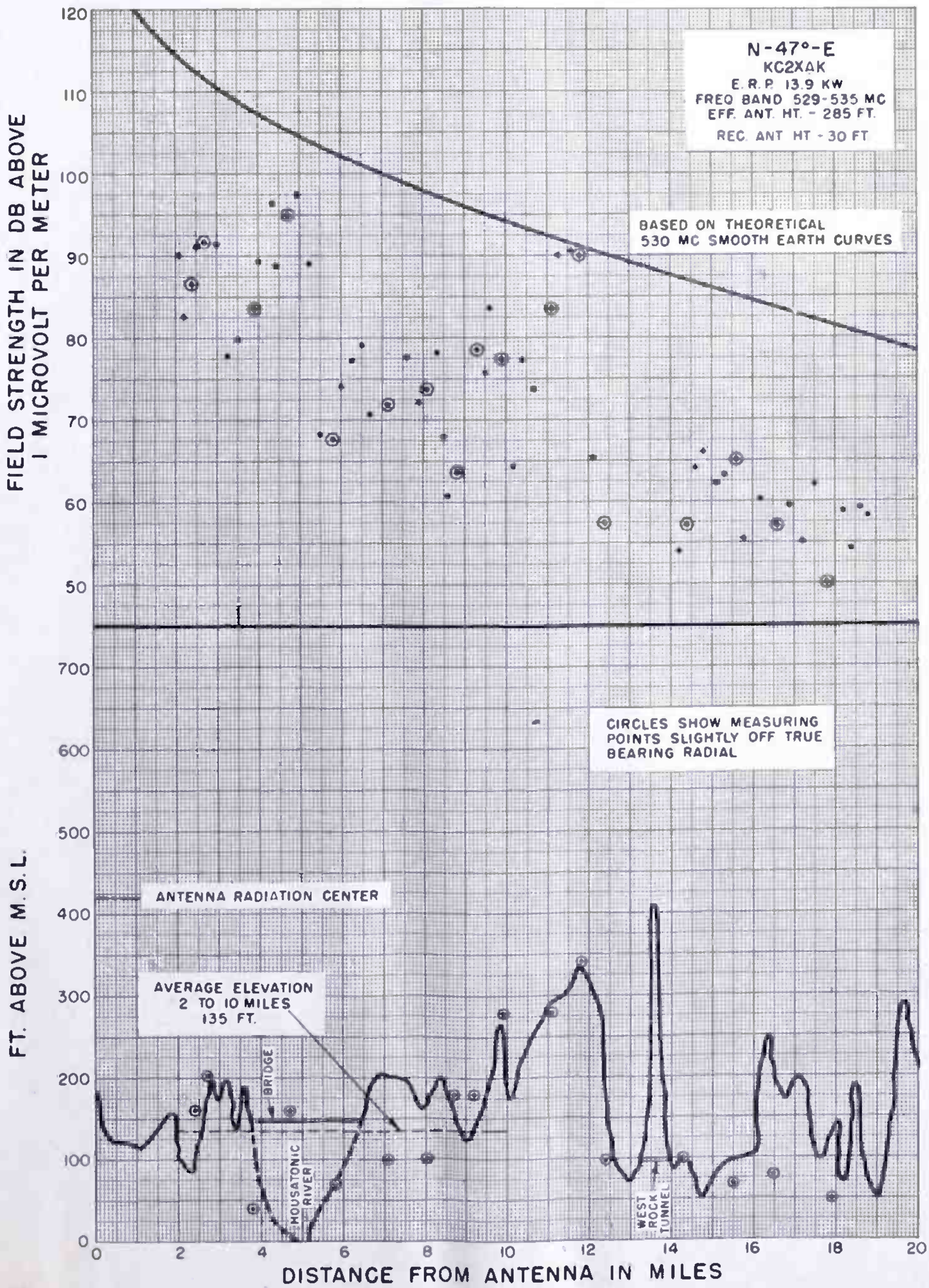


Fig. 11—Field intensity measurements, N-47° E radial.

obtainable at spot locations which were situated randomly in the Bridgeport area. At each location, the position of the receiving antenna was shifted several feet in all directions in order to establish the maximum field intensity which could be secured. The extreme variability of field intensities encountered in practice is indicated by the fact that the field intensities measured at two locations equally distant from the transmitter vary as much as 40 to 1.

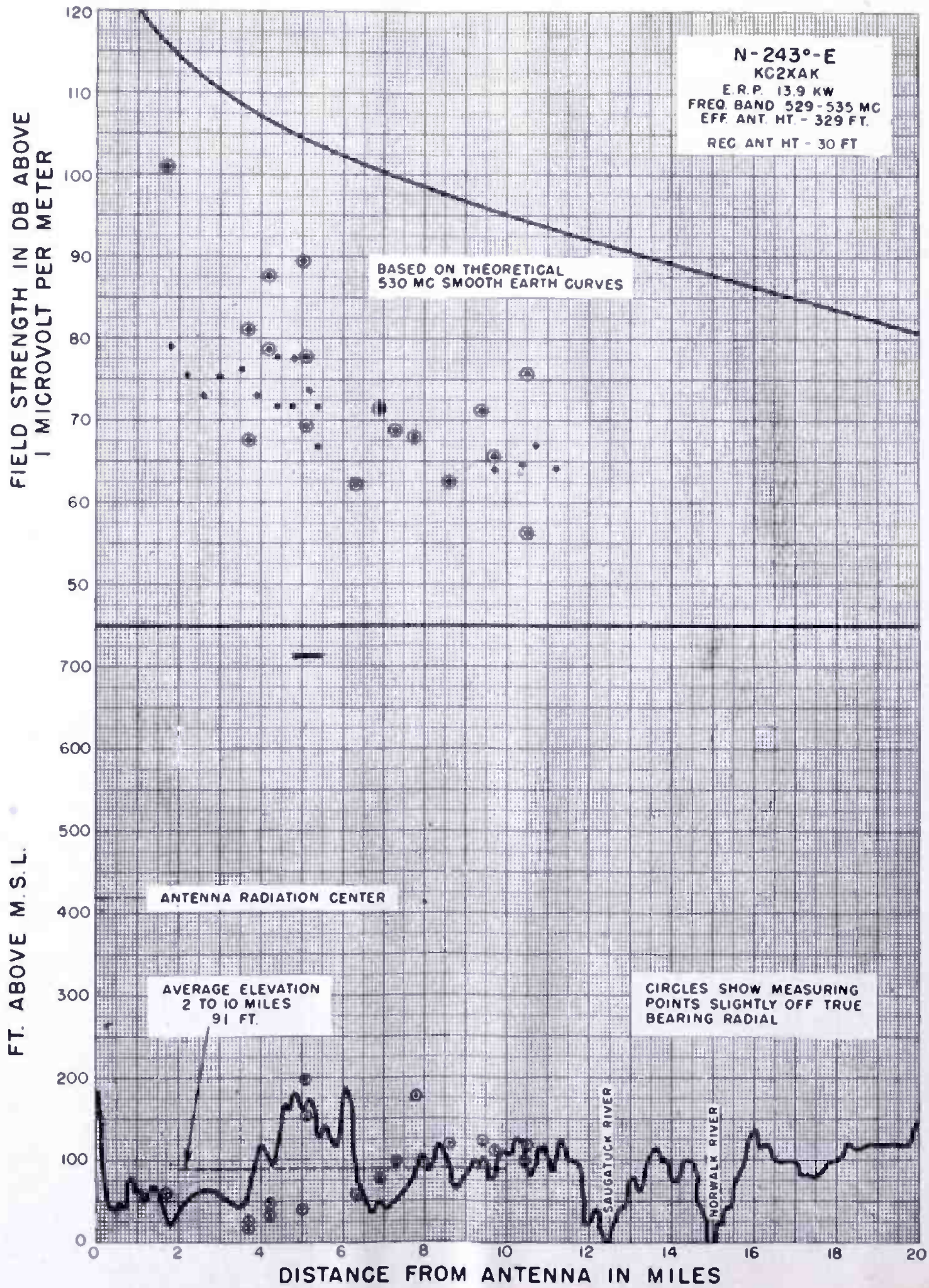


Fig. 12—Field intensity measurements, N-243°E radial.

d. Distribution of Picture Quality and Receiver Terminal Voltage with Distance

The simplest and perhaps the most revealing method of evaluating the service range of a UHF television station is the observation of pictures at typical receiving locations distributed throughout the area under study, because it integrates all factors to give directly the information sought. It is a method of analysis which is easily under-

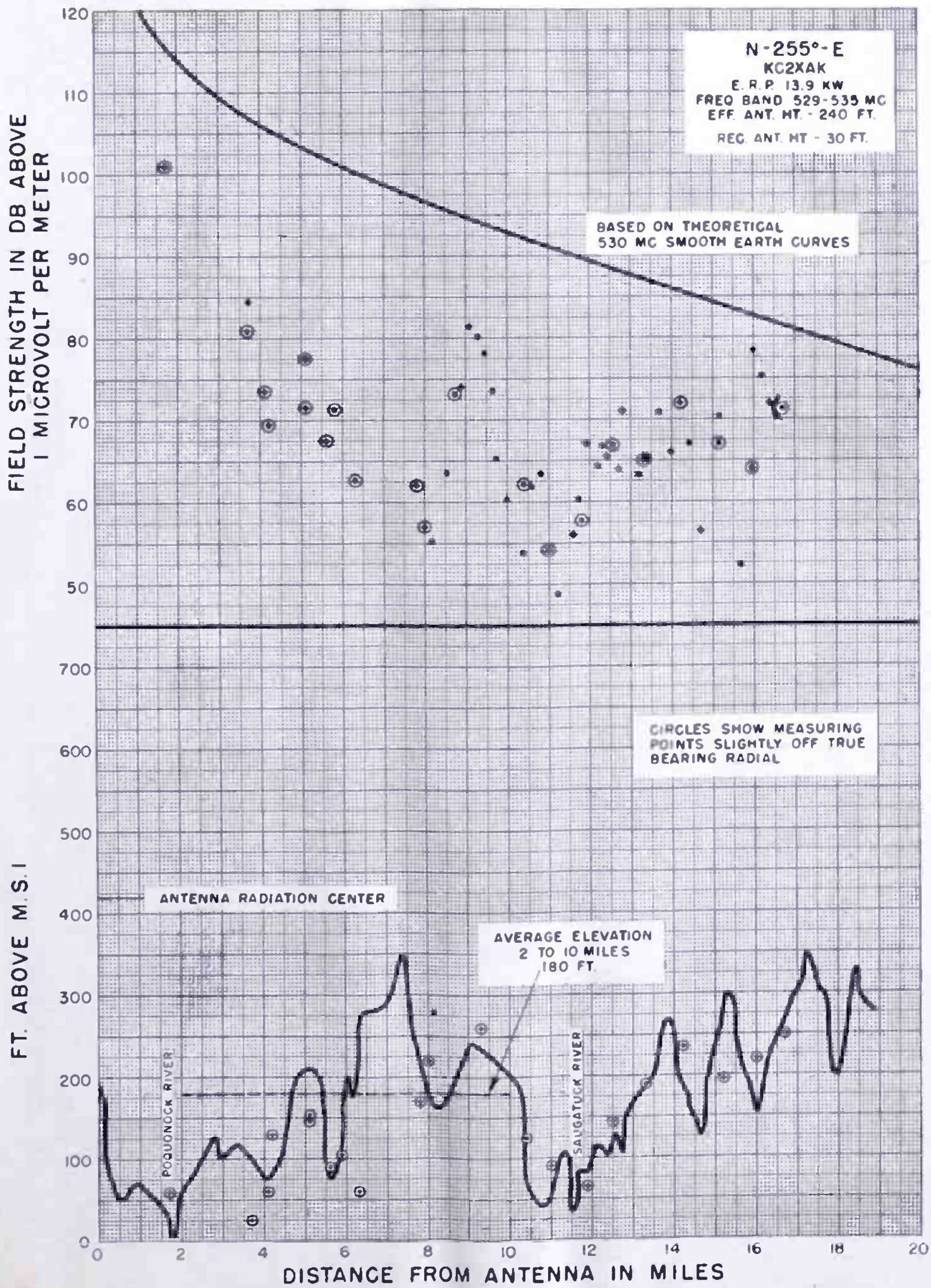


Fig. 13—Field intensity measurements, N-255°E radial.

stood and easily expressed. Since the voltage delivered to the receiver input terminals, from typical and practical antennas and transmission lines, is basically the factor which determines picture quality, it is important to know the average distribution of terminal voltage with distance.

Measurements and picture ratings have been made at home locations and also at spot locations throughout the area of interest. They have been analyzed in separate groups.

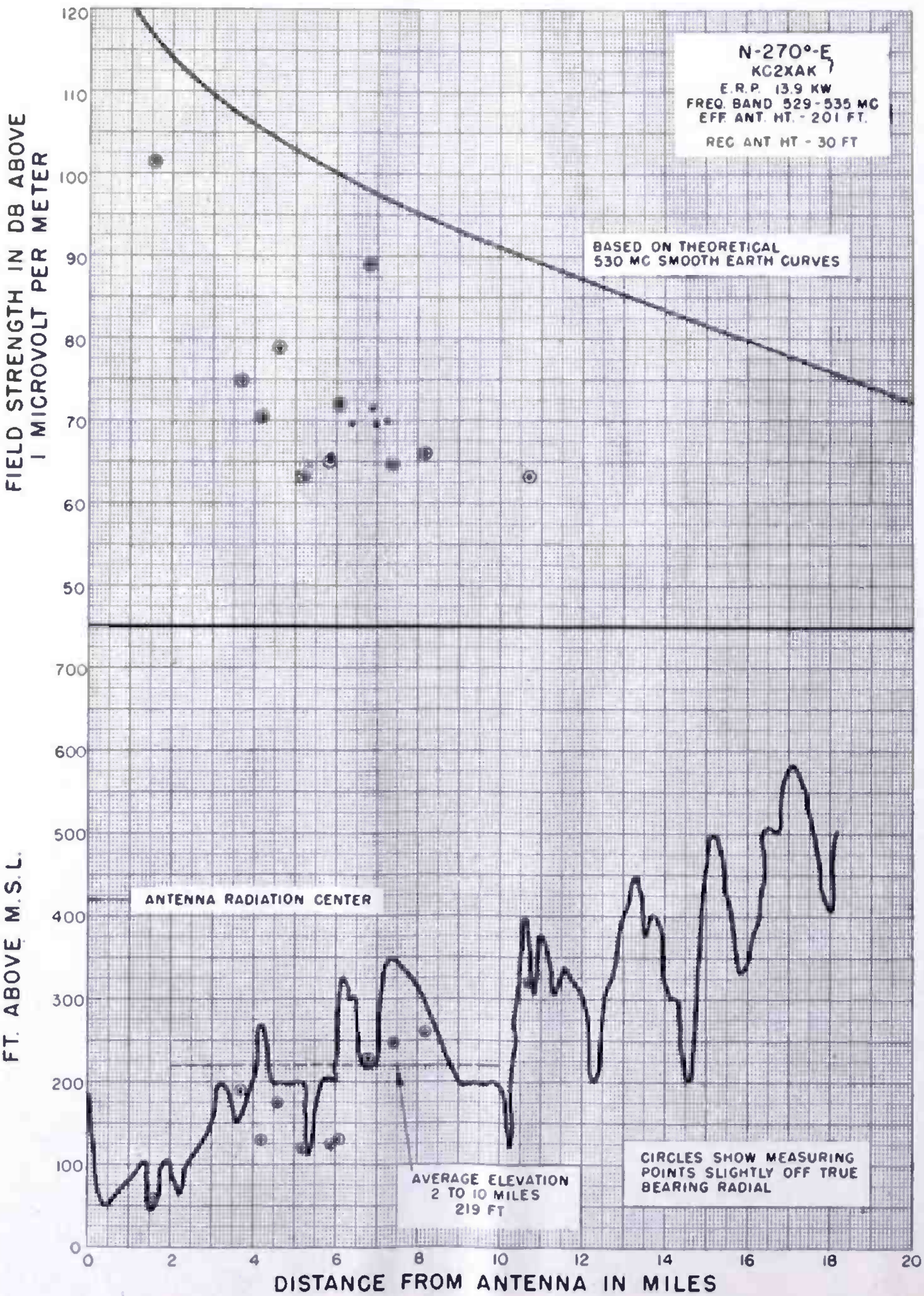


Fig. 14—Field intensity measurements, N-270°E radial.

Home receiver locations—For the home receiver measurements, several types of high gain directional antennas were used to pick up the signal. The simplest type which gave a satisfactory picture was installed at each location. The antennas were matched to the receivers, and the full antenna energy, less transmission line loss, was delivered to the receivers.

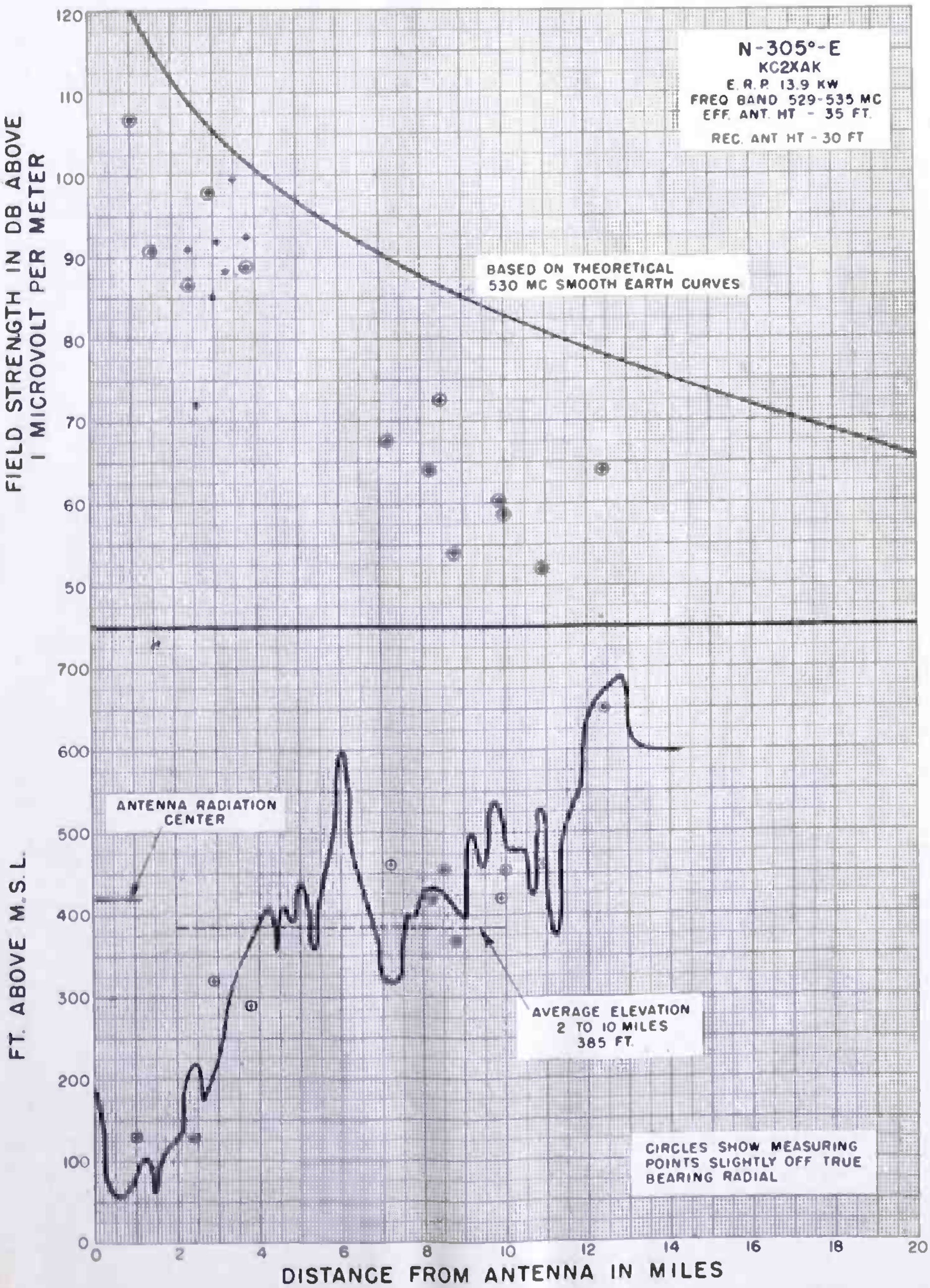


Fig. 15—Field intensity measurements, N-305°E radial.

Picture quality was observed and peak voltage at the receiver terminals was measured at homes within 15 miles of the transmitter. The data were analyzed to determine the distribution of both picture quality and receiver terminal voltage with distance. The results are shown in Tables I and II.

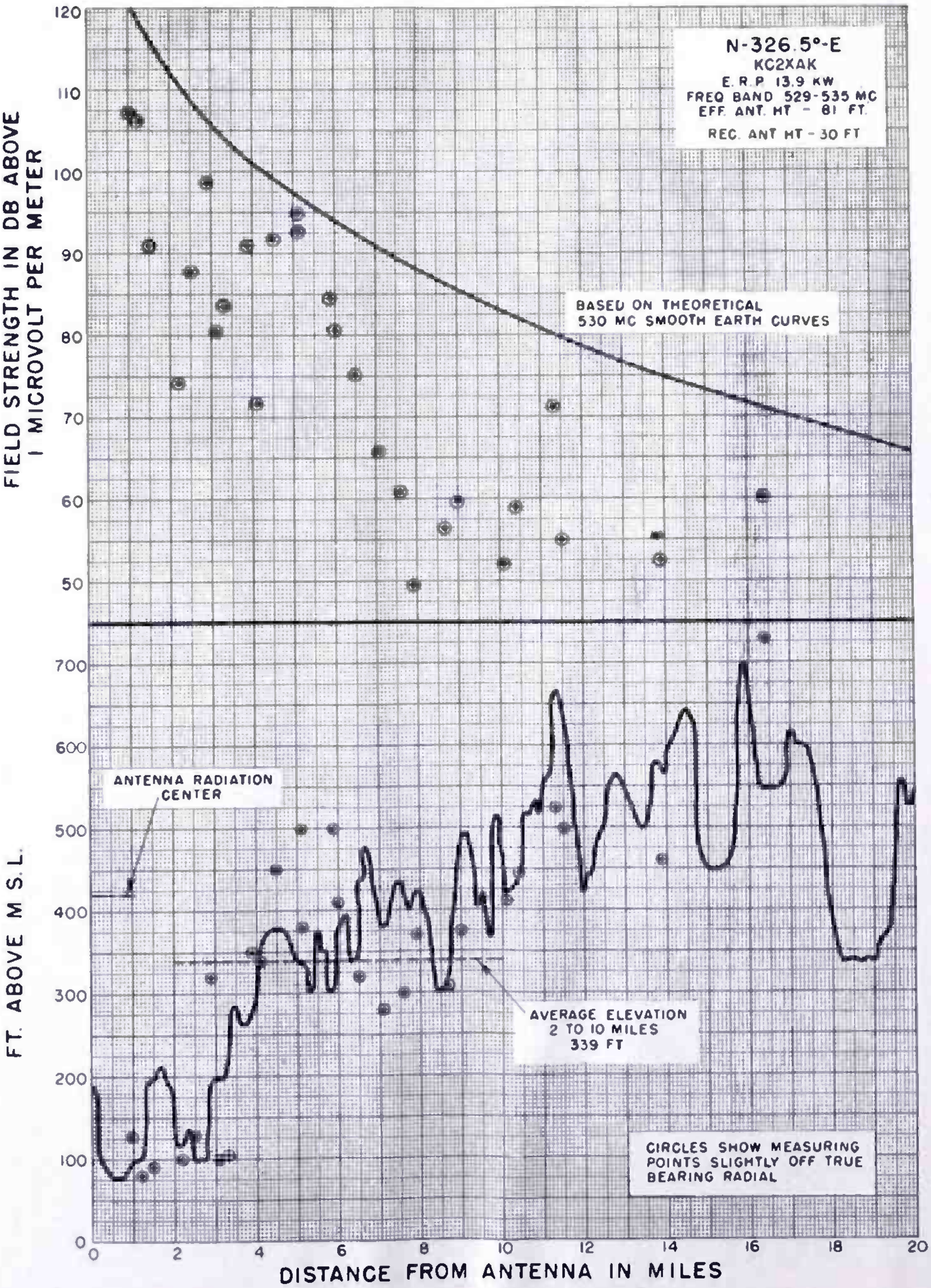


Fig. 16—Field intensity measurements, N-326,5°E radial.

Table I — Per Cent of Locations Which Received a Picture of the Indicated Quality or Better.

Distance (Miles)	Excellent	Good	Fair	Poor
0 to 5	59.1%	95.5%	100.0%	100.0%
5 to 10	26.3%	68.4%	94.7%	94.7%
10 to 15	5.0%	35.0%	55.0%	90.0%

Table II — Per Cent of Locations With Peak Receiver Terminal Voltage of the Indicated Value or Higher.

Distance (Miles)	400 μv	150 μv	60 μv	20 μv
0 to 5	81.2%	93.8%	100.0%	100.0%
5 to 10	40.0%	86.6%	100.0%	100.0%
10 to 15	7.1%	28.6%	64.3%	85.7%

The values of terminal voltage listed are those which were determined, in another section of this report, to be the approximate minimum values to give, respectively, "excellent", "good", "fair", and "poor" pictures.

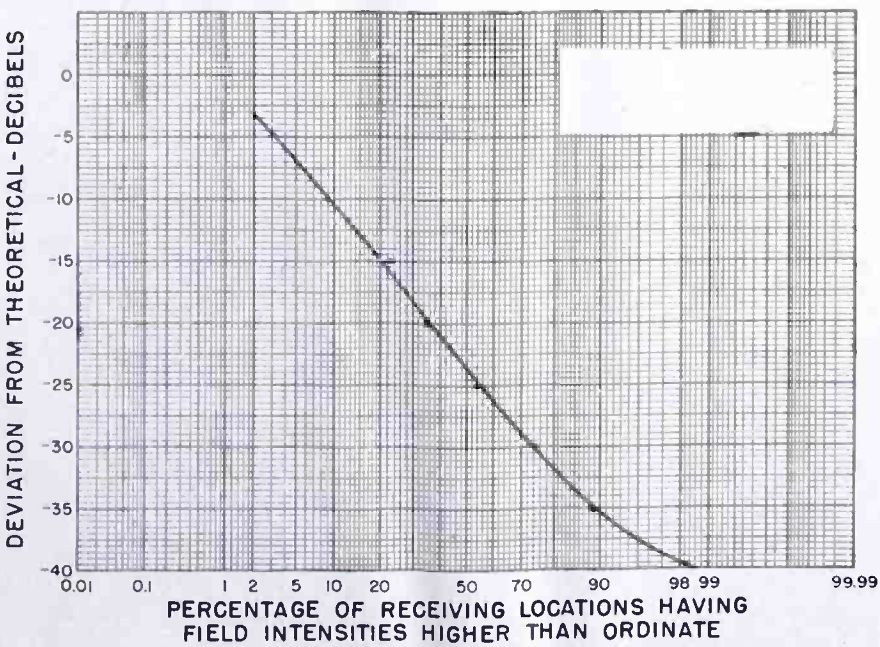


Fig. 17—Statistical distribution of measured field intensities as compared to theoretical values (534.75 megacycles).

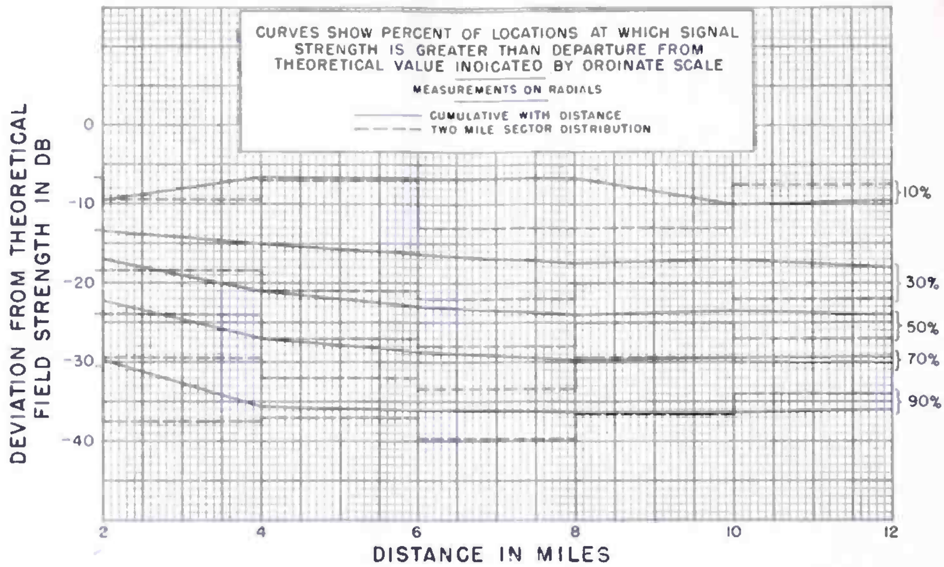


Fig. 18—Curves showing changes in the statistical distribution of measurements with distance.

Field measurements with mobile equipment—For the spot measurements, a non-directional superturnstile antenna with a gain of only 1.1 over a half-wave dipole was used to pick up the signal. Therefore the ratings and input voltages are lower than for home locations. This receiving antenna was elevated 30 feet above the ground. A 51.5-ohm coaxial cable and a matching pad were used to deliver the signal to the receiver.

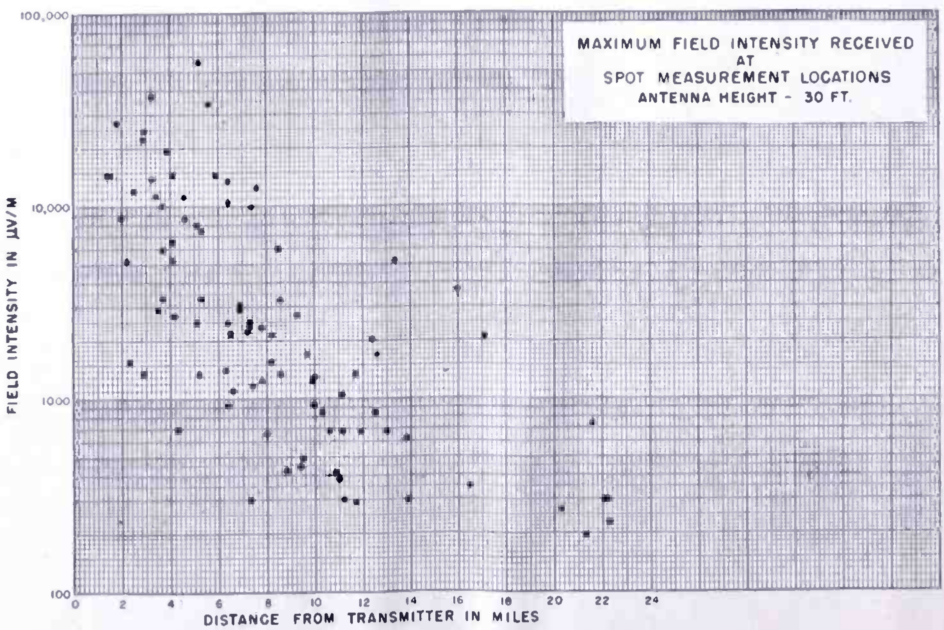


Fig. 19—Maximum field intensity received at spot measurement locations (antenna height 30 feet).

Picture quality was observed and peak receiver terminal voltage was measured at spot locations within 15 miles. At each location, the antenna was moved a horizontal distance of several feet to obtain a maximum and a minimum value of signal intensity. The maximum value observed was used to determine the distribution of receiver terminal voltage with distance. The results are shown in Table III. Table IV shows the distribution of picture quality with distance. It should be borne in mind that in the case of the spot measurements the ratings and input voltages are lower than for home locations because of the absence of antenna gain, and the higher transmission line losses.

Table III — Per Cent of Locations With Peak Receiver Terminal Voltage of the Indicated Value or Higher.

Distance (Miles)	400 μv	150 μv	60 μv	20 μv
0 to 5	36.0%	76.0%	88.0%	100.0%
5 to 10	10.0%	25.0%	57.5%	87.5%
10 to 15	0.0%	11.8%	17.6%	64.7%

Table IV — Per Cent of Locations Which Received a Picture of the Indicated Quality or Better

Distance (Miles)	Excellent	Good	Fair	Poor
0 to 5	32.0%	68.0%	76.0%	96.0%
5 to 10	14.3%	28.6%	52.4%	85.7%
10 to 15	0.0%	7.1%	10.7%	32.2%

As in the case of the home measurements, the correlation between these two tables is very good.

The analysis has been carried one step further by increasing the receiver terminal voltage for the spot measurements by 12 decibels, representing losses in the spot measuring apparatus arrangement, compared with typical home installation. The higher values could be obtained in practice by using a directional antenna with a gain over the superturnstile antenna, by matching the antenna to the receiver so that the matching pad may be eliminated, and by using transmission line of lower loss. When 12 decibels allowance is made, and the data is reanalyzed, the results are as shown in Table V.

The close correlation between Table I, for the home measurements, and Table V, for the upgraded spot measurements, indicates that the home receiver and spot measurements would agree quite well if the same measuring equipment were used.

Table V — Per Cent of Locations With Peak Receiver Terminal Voltage of the Indicated Value or Higher.

Distance (Miles)	400 μv	150 μv	60 μv	20 μv
0 to 5	76.0%	96.0%	100.0%	100.0%
5 to 10	25.0%	72.5%	90.0%	100.0%
10 to 15	11.8%	29.4%	70.5%	100.0%

c. Correlation of Picture Quality Ratings with Receiver Input Voltage

In making field intensity measurements it was found desirable to transmit test pattern and obtain picture quality ratings in terms of receiver input voltage. The receiver input noise is an important factor in determining the receiver input terminal voltage required to obtain pictures of various ratings. The selected receiver used in the caravan had a noise factor of 12 decibels. Those installed in homes varied from 12 to 22 decibels. Units which have noise factors of 12 decibels or slightly less appear to be approximately representative of good commercial UHF television receivers which could be built at present.

The rating of pictures in terms of "Excellent", "Good", "Fair", and "Poor" is subject to interpretation and opinion on the part of individuals. The commercial television standards impose no limitation on signal-to-noise ratio at the transmitter. Moderate but visible camera tube noise which now imposes a practical but not serious limit on the over-all system excellence may reasonably be expected to be overcome with future camera tube development, making possible television pictures free of visible noise. Thus, our present standard for "excellence" may in the future be improved to include true noise freedom provided only that the carrier input terminal voltage is adequate to overcome the thermal and receiver input noise and external interference.

Based upon observations of the author and his associates, thermal and receiver input noise become visible without causing degradation on the kinescope of a receiver having a noise factor of 12 decibels, when the carrier input voltage across the 72-ohm input circuits is reduced to about 1000 microvolts. Such observations of necessity were made under conditions where the receiver input noise was masked to some degree by the noise of the camera tube which created the test pattern at its source. Thus, 1000 microvolts under these conditions, may represent approximately the lowest input voltage required to provide noise-free pictures in the future. Under existing conditions where camera tube noise is always present and completely noise-free pictures are unattainable, a lower standard of excellence is accepted.

An analysis of picture ratings and receiver input terminal voltages in the KC2XAK service area, indicated that under present conditions 400 peak microvolts at the receiver input terminals was the lowest value that could reasonably be used for "excellent" ratings, using a receiver with a noise factor of 12 decibels. Because of variations in judgment from time to time by individuals, differences in judgment among individuals, and the need to make field observations under less than ideal conditions, there was a spread in the conclusions reached among the 125 locations at which ratings and input voltage measurements were made. The ratings shown below apply to receivers with a noise factor of 12 decibels.

Quality	Peak Voltage at 72-ohm Receiver Input in Microvolts
Excellent	400 and up
Good	150 — 400
Fair	60 — 150
Poor	20 — 60

The analysis of picture ratings and input voltages for homes having receivers with random noise factors, usually in excess of 12 decibels, was made independently. The correlation was good. The values given in the table above may be used for receivers having noise factors higher than 12 decibels by applying the noise factor differential. Correspondingly, if lower noise factors are attained the values in the above table may be reduced.

f. Comparison of Picture-Carrier Field Intensities and Sound-Carrier Field Intensities

Since both sound and picture carriers were being transmitted simultaneously, it was found desirable to obtain some statistical data as to the ratio of the field intensity of the two signals at any particular location.

Measurements were made at 91 locations using the 30-foot super-turnstile antenna. At each location the antenna was moved several feet backward and forward to obtain a maximum reading for each carrier.

The carrier field intensities were corrected for equality for 50 per cent of the measurements. The variation of the sound and picture intensities with respect to each other are summarized statistically in Figure 20. These were maximum values found by moving the receiving antenna over a distance of several feet. The exact locations of the maximum picture and sound intensities generally differed by distances of several inches to a few feet.

g. Variation in Signal with Antenna Position at Spot Locations

It is known that field intensities at UHF vary widely over a distance of a few feet at many locations. With this in mind, measurements were made at 91 spot locations to determine the degree of variation

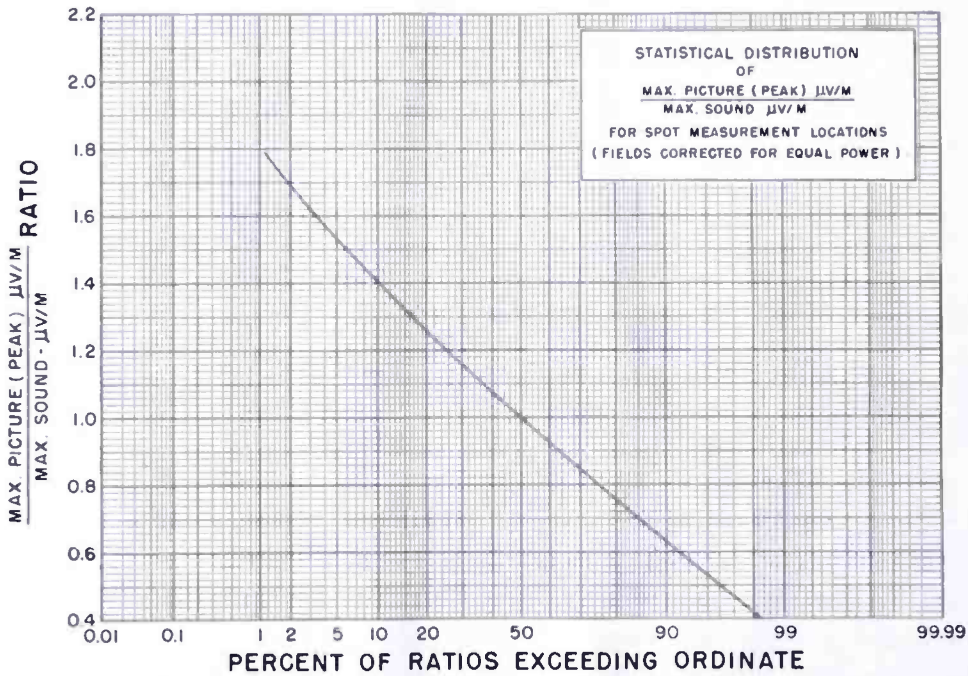


Fig. 20—Statistical distribution of the ratio of maximum picture signal (peak) to maximum sound signal for spot measurement locations.

of picture carrier intensity. The locations chosen were as clear as could be obtained for the type of terrain. A superturnstile elevated 30 feet was again used. At each location the antenna was moved a distance of five feet or more forward and backward. This minimum of five feet was chosen so as to get all variations within several wavelengths. The maximum and minimum values were recorded and analyzed statistically, and the results obtained appear in Figure 21.

The measured ratios varied from a maximum of 7.0 to a minimum of 1.0. As can be seen, fifty per cent of the measurements showed a maximum to minimum field intensity exceeding 1.49. It is conceivable that in congested built up areas wider variations could be expected. Hence, in a weak signal area it is necessary to search for optimum receiving antenna locations in a horizontal as well as a vertical direction.

h. Comparison of KC2XAK and WNBT Picture Ratings at Home Locations in the Bridgeport Area

At each of the sixty-seven home locations where comparison ratings were made, an effort was made to employ a combination UHF-VHF installation which would give as good pictures as possible under practical conditions. In most locations, a two-section VHF antenna with a 4-decibel gain at Channel 4 was used. The UHF antenna varied from a fan to a parabola, depending on which one was required. 300-ohm twin ribbon lead was used with the VHF antennas. The 300-ohm

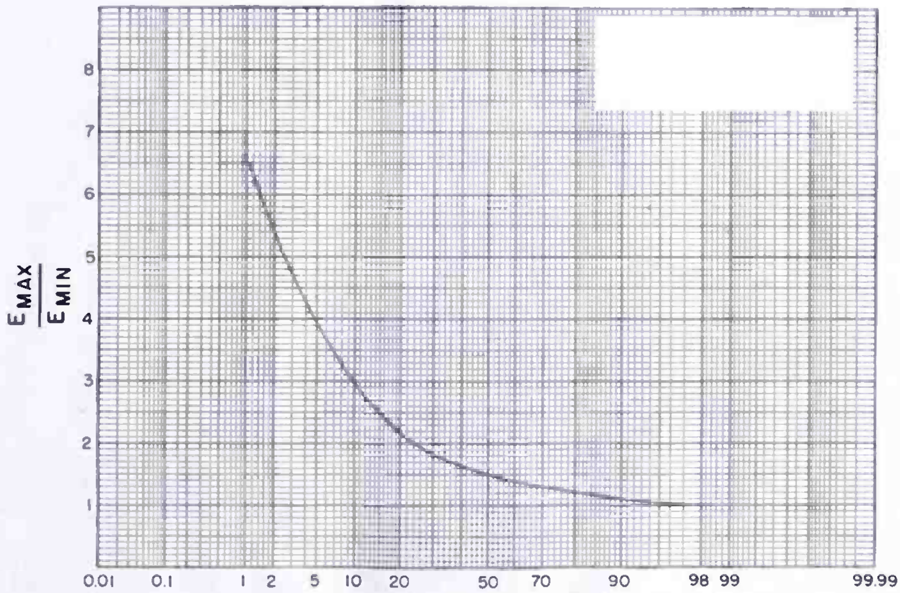


Fig. 21—Statistical distribution of ratio of maximum to minimum video carrier signals for spot measurement locations (530.25 megacycles).

tubular cable was used in weak signal areas for UHF, RG/59U being used in strong signal areas where the loss could be tolerated.

Picture quality ratings were assigned at each home location to the UHF and also the VHF Channel 4 signal from WNBT in New York City, 55 miles distant. The following is a tabulation of the ratings.

UHF PICTURE QUALITY	Locations	Per Cent of Total	VHF PICTURE QUALITY	Locations	Per Cent of Total
Excellent	18	26.9	Excellent	6	9.0
Good	23	34.3	Good	16	23.9
Fair	10	14.9	Fair	31	46.2
Poor	12	17.9	Poor	14	20.9
Unusable	4	6.0	Unusable	0	0.0

In making any comparison between the UHF and VHF reception in this case, the following data should be remembered.

	UHF Station	VHF Station
Effective Radiated Power	13.9 kw.	7 kw.
Antenna Height Above Average Terrain	330 ft.	1280 ft.
Average Distance to Receiving Set	9.6 Miles	54 Miles

The Bridgeport-Stratford area is outside the nominal half-millivolt area of WNBT, so it hardly would be expected that the few hundred microvolts per meter provided there would give "excellent" or even "good" pictures to many locations. But it is not unreasonable to expect that service might be "fair". The ratings for WNBT of "fair" or better total 79.1 per cent, compared with 76.1 per cent for the UHF station ratings.

i. Height — Gain Analysis of Receiving Antennas

The height-gain measurements are divided into two sections;

(a) Spot Measurements Along Radials,

(b) Continuous Measurements Along Radials.

Spot Measurements Along Radials—The radial spot measurements were made with the intention of obtaining statistical data as to antenna height versus field intensity. Locations were chosen approximately one mile apart on seven radials. In all, 107 locations were used to obtain data for the probability curves. Figure 22 shows a comparison of the 46-foot, 38-foot, and 30-foot height-gain curves. In Figure 23, the 50 per cent point on each of the height-gain curves has been selected and plotted against antenna height.

From the data obtained, it can be seen that there is a median variation in voltage gain over a 10-foot antenna which is almost a linear function between 30 and 46 feet. Individual cases showed wide discrepancies in a number of places. However, when a sufficient number of locations are analyzed, a definite pattern of behavior can be seen when the data is plotted in a statistical manner.

Continuous Height-Gain Measurements Along Merritt Parkway—Since a relatively clear section of road extending approximately east and west of the transmitter was available for continuous recordings at thirty feet, a run was made using the mobile caravan, with antennas at the 10- and 30-foot levels. In all, 96 sectors of one-quarter mile or less were obtained from recordings. The resultant distribution by

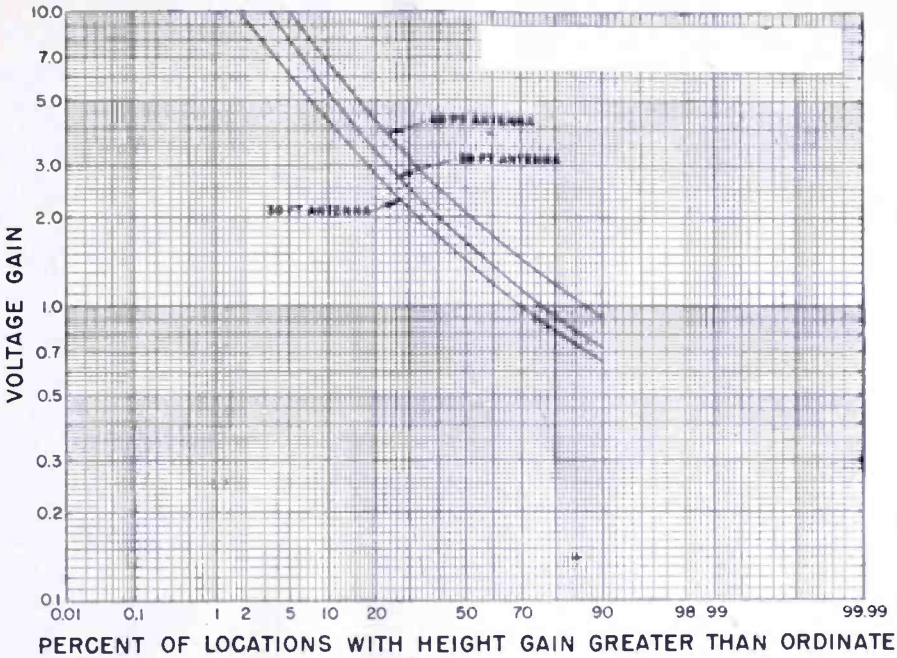


Fig. 22—Comparison of gain of 46-foot, 38-foot and 30-foot antennas over 10-foot antenna.

sectors of voltage gain of a 30-foot antenna over a 10-foot antenna is shown in Figure 24. The median voltage gain for 50 per cent of the sectors was 1.95 (5.8 decibels) or better. The relatively unobstructed Merritt Parkway terrain accounts for the difference in height-gain characteristics compared with the group first measured and analyzed.

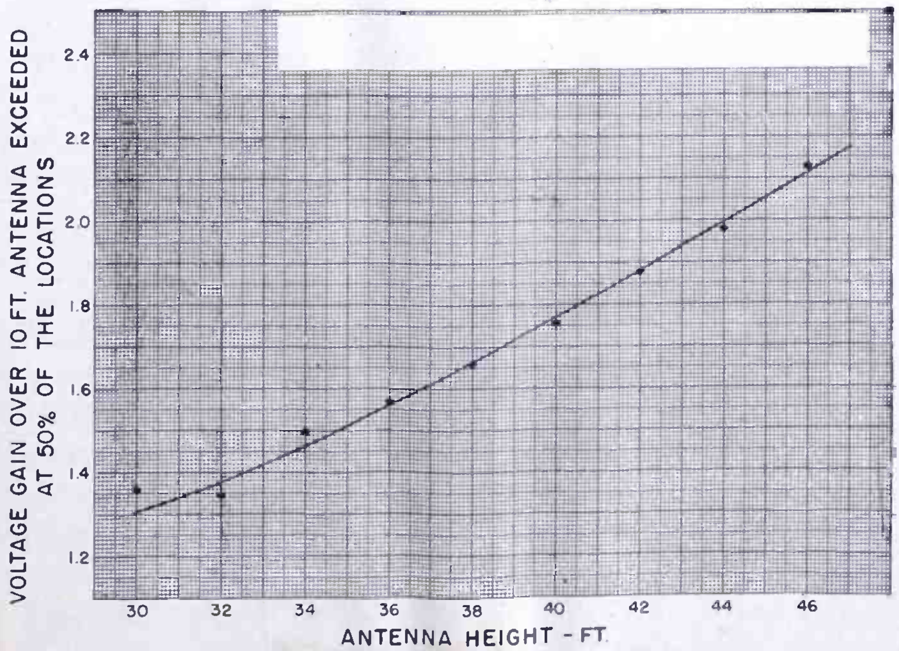


Fig. 23—Voltage gain over 10-foot antenna exceeded at 50 percent of the location versus antenna height in feet.

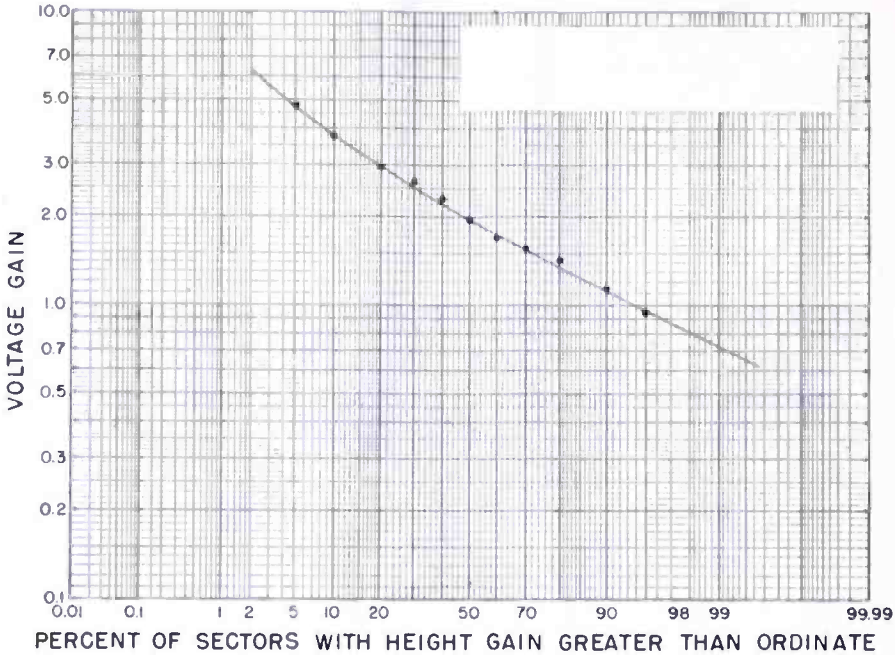


Fig. 24—Gain of 30-foot antenna over 10-foot antenna, Merritt Parkway east and west radials, total of 96 sectors (continuous radial measurements).

UHF-VHF Comparison

In addition to the comparison of KC2XAK and WNBT picture ratings at home locations in the Bridgeport area previously reported, a more direct comparison of UHF and VHF transmissions was made.

Fig. 25 — 67.25-megacycle transmitter used in UHF-VHF comparison survey.

In addition to the UHF transmitting equipment described, the visual portion of a standard RCA Type TT-500A transmitter was installed in the transmitter building, as shown in Figure 25. This transmitter was operated at a carrier frequency of 67.25 megacycles (Channel 4) and adjusted for 500 watts (peak) output as measured in a dummy load. The VHF and UHF transmitters were modulated with the same test pattern signal as received on the 2000-megacycle link from WNBT.

A single layer turnstile antenna, having a power gain of 0.5, was used for the VHF transmissions.

This antenna was mounted at the radiation center of the UHF antenna, shown in Figure 26, so that the effective height of both antennas would be the same. A diplexer was mounted on top of the supporting tower to provide the necessary phasing and impedance matching for the turnstile antenna. Two hundred and fifty feet of RG 17/U transmission line was used to connect the transmitter to the diplexer. The effective radiated power of the VHF transmitter was 217 watts (peak) visual, and the corresponding UHF power was 13,900 watts, a ratio of 1:64.

UHF-VHF Comparison Measurements

The mobile equipment used for the UHF-VHF comparative measurements was essentially the same as that described. The only change in the equipment was the antenna and transmission line. A standard 2-bay Channel Master antenna with a gain of 4 decibels over a dipole and 50 feet of K-111 transmission line were used for the VHF measurements and a stacked Vee antenna with a gain of 5.7 decibels and 50 feet of RG-11/U transmission line were used for the UHF measurements. Both antennas were mounted 30 feet above the ground, as shown in Figure 27 and were equipped with a rotator so that they could be oriented for maximum signal.

Fig. 26—67.25-megacycle turnstile antenna used in UHF-VHF comparison survey mounted at radiation center of UHF antenna.

At each location the VHF field intensity was measured with the WX-1A receiver and the UHF field intensity was measured with the AN/APR-4 receiver. Both signals were quality rated using a single UHF-VHF television receiver.

Results

In this comparison of UHF with VHF, spot measurements were made at a total of 120 points along 9 radials, as shown in Figure 28. The measuring points were taken at approximately 1-mile intervals along the radials and as close to the radials as the terrain and available roads would permit.

Tables VI, VII, and VIII below show three different comparisons of UHF and VHF reception with regard to picture quality rating. Table VI indicates the percentage of total locations within specific mileage brackets which receive the indicated quality. Table VII presents the same data in cumulative form.

Table VIII shows the relative quality of UHF versus VHF reception for each individual radial. For example, on Radial A, picture quality was assessed at 14 separate locations. At 10 of these, the UHF picture

Fig. 27 — Mobile equipment used in UHF-VHF comparison survey.

quality equalled that obtained on VHF. At 4 of these, the VHF reception was superior by one grade to the UHF, etc.

Computations indicate that if the UHF power were increased from 64 times to 460 times the VHF power, the average of the picture ratings would be about equally good for UHF and VHF within 10 miles.

In this survey the general observation was made that the UHF signal was higher than during the summer when previous measurements were made. This is probably due to the absence of heavy foliage encountered previously.

Table VI — Per Cent of Locations which Received a Picture of the Indicated Quality

Distance (Miles)	Excellent UHF	Excellent VHF	Good UHF	Good VHF	Fair UHF	Fair VHF	Poor UHF	Poor VHF	Unusable UHF	Unusable VHF
0 to 5	84.5	97.0	12.5	3.0	3.0	0	0	0	0	0
5 to 10	53.0	67.7	14.7	11.8	8.8	17.7	11.8	2.8	11.7	0
10 to 15	6.9	10.4	6.9	10.4	10.4	41.4	17.2	27.6	58.6	10.2
Beyond 15	0	0	8.0	8.0	8.0	36.0	32.0	20.0	52.0	36.0

Table VII — Per Cent of Locations which Received a Picture of the Indicated Quality or Better

Distance (Miles)	Excellent UHF	Excellent VHF	Good or Better UHF	Good or Better VHF	Fair or Better UHF	Fair or Better VHF	Poor or Better UHF	Poor or Better VHF	No. of Locations
0 to 5	84.5	97.0	97.0	100	100	100	100	100	32
5 to 10	53.0	67.7	67.7	79.5	76.5	97.2	88.3	100	34
10 to 15	6.9	10.4	13.8	20.8	24.2	62.2	41.4	89.8	29
Beyond 15	0	0	8.0	8.0	16.0	44.0	48.0	64.0	25
								Total	120

Table VIII — Comparative Quality Rating

Radial	No. of Locations	UHF 2 Grade Better	UHF 1 Grade Better	Equal	VHF 1 Grade Better	VHF 2 Grade Better
A	14			10	4	
B	15			8	4	3
C	16		1	6	5	4
D	8			7	1	
E	6			6		
F	21	2		12	4	3
G	15		1	8	3	3
H	10	1		4	4	1
J	15			4	7	4
Total	120	3	2	65	32	18

Summary of Several UHF Projects

In addition to the ultra-high-frequency television field tests which have been referred to herein, other companies have undertaken projects.

For example, the Philco Company made an independent investigation of UHF propagation utilizing the transmissions of the RCA-NBC station at the Wardman Park Hotel. Independently, the Westinghouse Company made propagation tests at Pittsburgh. And also independently, the Columbia Broadcasting System made tests in the New York area on 480 megacycles. The results of all of these projects have been reported.[6-8]

In each of these projects reported, the results have been statistically analyzed in a manner corresponding generally to the NBC analysis

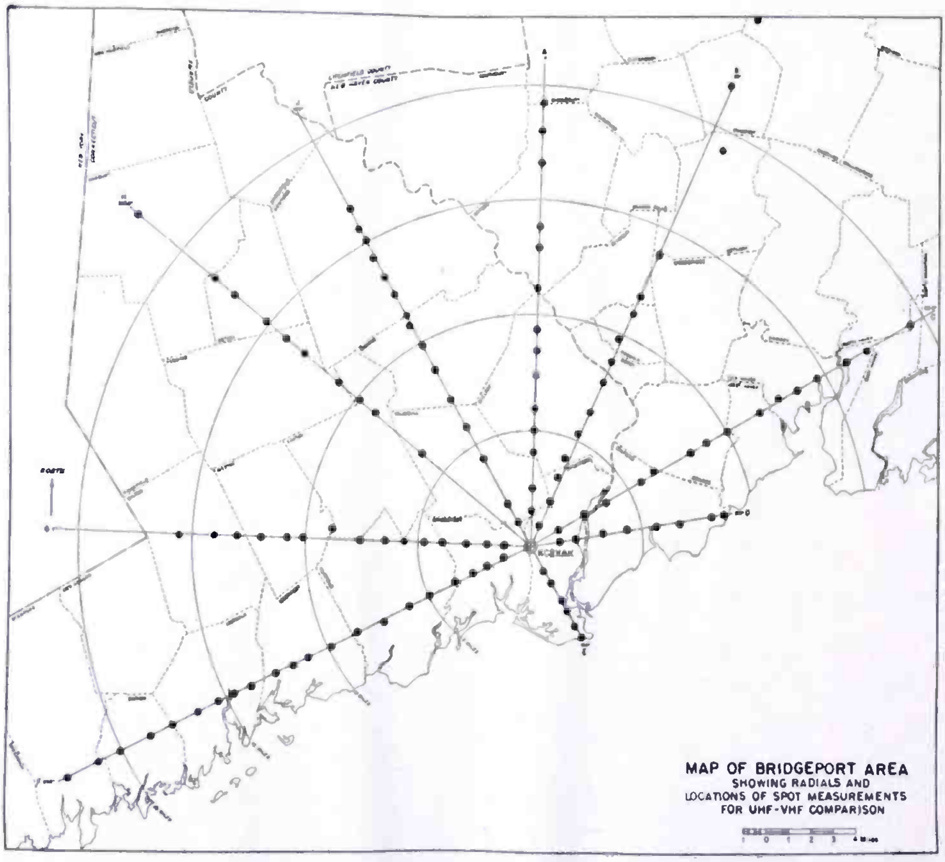


Fig. 28—Map showing spot measurement locations used in UHF-VHF comparison survey.

of the Bridgeport project. In each, the field intensity obtained at points of measurement was compared to that predicted for the location, using the theory developed by K. A. Norton. The disparity was expressed as the number of decibels below the theoretical value.

[6] W. B. Lodge for the Columbia Broadcasting System, FCC Docket #7896.

[7] J. Fisher, "Field Test of UHF Television", *Electronics*, Vol. 22, No. 9, September, 1949.

[8] R. N. Harmon, "UHF Coverage in Pittsburgh", *FM-TV*, May, 1950.

The data on all of these projects have been assembled in Figure 29, which follows. In Figure 29 an average curve, H, has been drawn in to represent the composite result of all of the projects indicated. It may be seen that although there are differences in frequency, geographical location and terrain, there is remarkable agreement in results, the greatest deviations from the average being those reported by CBS. If the latter were corrected to use the same assumptions, the disparity would be reduced by 6 decibels and the average curve H, would be lowered.

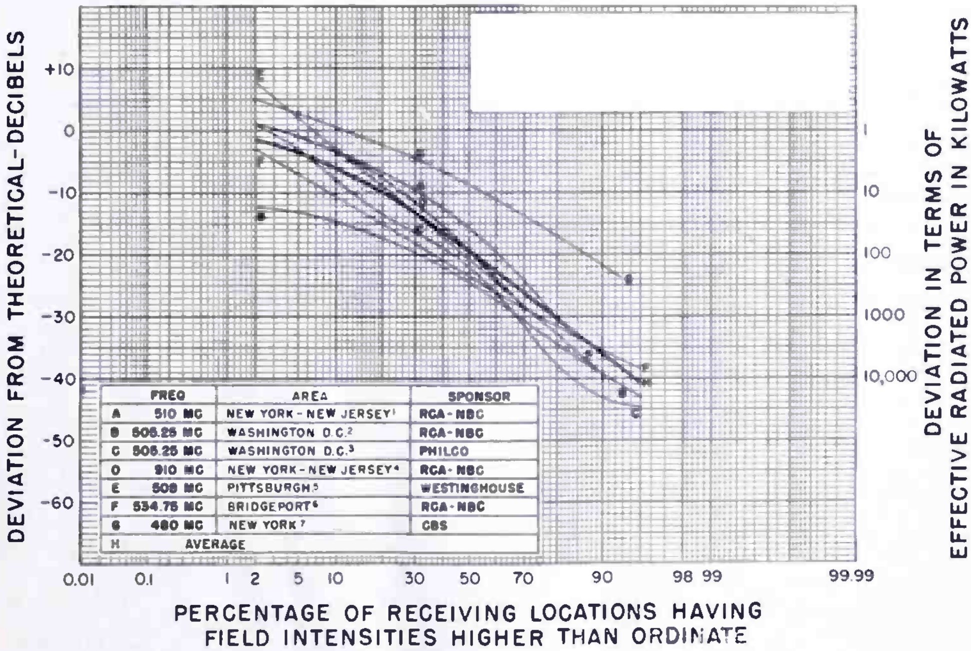


Fig. 29—Experience curves. Measured UHF field intensities compared to those predicted by the smooth earth theory.

The following is a comparison of actual versus smooth earth theoretical results.

(a) At 10 per cent of the receiving locations the actual measured field intensity for the average of all of these projects is below theoretical values by 36 decibels or more.

(b) At 30 per cent of the receiving locations the actual measured field intensity for the average of all of these projects is below theoretical values by 26 decibels or more.

(c) At 50 per cent of the receiving locations the actual measured field intensity for the average of all of these projects is below theoretical values by 20 decibels or more.

Because of space limitations it was not feasible to include valuable data on height-gain investigations made by H. O. Peterson in the general area of Riverhead, L. I., nor was it feasible to include analysis of other measurements such as those made of tropospheric transmission recordings at a number of recording stations. It is anticipated that this information will be published later.

ACKNOWLEDGMENTS

The RCA-NBC UHF research program in the Bridgeport, Connecticut area was initiated and is being coordinated by C. B. Jolliffe, Executive Vice President in charge of RCA Laboratories Division, with the cooperation and assistance of D. F. Schmit, Vice President and Chief Engineer of the RCA Victor Division, which developed and built the transmitting and receiving apparatus, and O. B. Hanson, Vice President and Chief Engineer of the National Broadcasting Co., which built the station and is conducting the field investigations.

It is not feasible to outline the contributions of the many individuals connected with the over-all project, but the author would be remiss in not acknowledging the assistance of V. E. Trouant, O. Fiet, T. A. Gluyas and M. Sinnett of the RCA Victor Division, L. A. Looney, J. Seibert, F. W. Smith and S. E. Piller of the National Broadcasting Co. and H. Laessle of the RCA Service Co.

LOW-REFLECTION FILMS PRODUCED ON GLASS IN A LIQUID FLUOSILICIC ACID BATH*

By

S. Milton Thomsen

Research Department, RCA Laboratories Division,
Princeton, N. J.

Summary—The low-reflection film known as "Magicote C" is produced on glass surfaces by the chemical attack of a liquid in which the glass is immersed. The liquid is a modified aqueous fluosilicic acid solution of controlled composition.

This is a low-cost process, capable of handling large sheets of glass. The reflection of white light is reduced to 5 per cent of that of the original unprocessed glass. Best suited for filming by this method are the crown glasses, including window glass and polished plate glass.

Introduction

A METHOD of producing low-reflection films on glass surfaces by the chemical attack of fluosilicic acid vapor has been described by Nicoll.[1,2] The film consists of a porous silica skeleton, remaining after the removal of part of the glass from a surface layer about 1000 Å in thickness.

Subsequently Nicoll and Williams[3] observed that such films could also be produced by liquid fluosilicic acid baths in which some glass had been dissolved. From this observation has been developed the method to be reported here.[4] The chief advantages of this method over the earlier vapor method are that both sides of a sheet (or all surfaces of an object) are processed simultaneously, less handling and less floor space are required, and the actual processing time is about one-twentieth as long.

The name "Magicote C" is applied to the low-reflection film produced by either the vapor or liquid bath method.

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* Decimal Classification: 540.

[1] F. H. Nicoll, "Low Reflection Films on Glass by an Improved Chemical Method," *RCA Review*, Vol. X, No. 3, pp. 440-447, September, 1949.

[2] F. H. Nicoll and F. E. Williams, U. S. Patent 2445238, July 13, 1948.

[3] F. H. Nicoll and F. E. Williams, U. S. Patent 2486431, November 1, 1949.

[4] S. M. Thomsen, U. S. Patent 2490662, December 6, 1949.

Equipment

The container for the filming bath is a tank of such size and shape that the glass sheets to be treated may be fully immersed. The tank is inert to attack by warm fluosilicic acid, and preferably transparent. Pyrex glass and methyl methacrylate plastics are suitable materials. Provision is made to agitate the liquid, and to control its temperature within one degree centigrade. To retard evaporation from the warm liquid, the tank is provided with a close-fitting cover. Adequate ventilation is maintained in the region, and the equipment is surrounded by a hood, because the vapor is toxic and corrosive.

An auxiliary tank, containing 0.5 per cent hydrofluoric acid at room temperature, permits sheets of glass to be given a short cleaning dip before being filmed.

Processing Bath

(a) *Preparation*

The bath is 1.25 molar fluosilicic acid, saturated with silica. This concentration is approximated by diluting the commercial acid with an equal volume of water. Titration with sodium hydroxide at 80°C to the phenolphthalein end-point is used to determine the concentration.

The saturation with silica is accomplished at room temperature by adding hydrated (precipitated) silica to the diluted acid and agitating the mixture. With frequent or continuous agitation, the saturation is complete in several hours. When the addition of a pinch of fresh hydrated silica (having some of the silica present as "dust") produces a permanent turbidity, the saturation process is complete.

About 15 grams of hydrated silica per liter are dissolved in the process. The saturated acid is filtered through paper, put into the tank, and brought to the operating temperature (about 45°C).

A small adjustment of the composition is required, to match the performance of the bath to the requirements of the glass. This is accomplished for 45°C operation by adding from 1 to 6 milliliters of 4 per cent boric acid per liter of treating liquid, the amount being determined by trial, or by a method to be described later.

(b) *Control of Composition*

Commercial fluosilicic acid contains about 30 per cent H_2SiF_6 in water solution, which makes it about 2.5 molar. After this acid has been diluted with water and saturated with silica, the composition of the solute approximates that of fluodisilicic acid,[5] $H_2SiF_6 \cdot SiF_4$. The

[5] S. M. Thomsen, "High-Silica Fluosilicic Acids," *Jour. Amer. Chem. Soc.* Vol. 72, No. 6, p. 2798, June 1950.

silica-saturated acid, like the original acid, dissolves glass, but much more slowly.

The term "potency" is applied to that property of a fluosilicic acid solution which determines the vigor of its attack on silica and, similarly, on glass. As a reference value, the silica-saturated acid is assigned zero value of potency. Addition of fluoride to the solution increases potency, while the addition of silica diminishes it. Unit increase in potency is defined as that produced by the addition of one millimole of HF per liter of solution. In practice, 3-molar potassium fluoride solution is used because it is easier to handle. Addition of one milliliter per liter of solution increases the potency 2 units.

To obtain the selective attack on glass required to produce a low reflection film of skeletonized silica, the bath must be used in the region of small negative potency values, in which case the solution is supersaturated with silica. These negative potency values can be achieved with a number of reagents, but not with silica itself. A 4 per cent boric acid solution is used, of which one milliliter added per liter of solution diminishes the potency 2 units.

Two circumstances require occasional adjustment of the potency. First, the supersaturated solution slowly deposits silica. To compensate for the resulting slow progressive increase in potency, amounting to one or two units per day, boric acid is added at intervals. Second, when a different glass is to be filmed, adjustment of the potency, up or down, may be required.

(c) *Useful Life*

Continued use of the bath results in two changes. Slow precipitation of silica occurs, producing turbidity. In the solution, both fluoboric acid and salts accumulate. The salts are principally the fluosilicates of the sodium and calcium from the glass being filmed. In spite of these changes, a bath gives many months of service without noticeable change in performance. Filtration about once a month is desirable to keep the bath nearly clear, which, in turn, keeps the potency drift rate low. The rate of silica precipitation increases as the amount of surface available for its deposition increases.

(d) *Concentration*

The concentration of acid, 1.25 molar, was selected as a compromise between the slower attack of more dilute solutions, and the greater volatility of the more concentrated solutions. The time required for filming varies inversely with the concentration. From dilute solutions, only water is appreciably lost by evaporation. As concentration increases, progressively greater loss of silicon fluoride occurs. The rapid loss of silicon fluoride impairs the control over the composition of the bath, and increases the health hazard.

Operation of the Process

(a) *Typical Procedure*

A sheet of glass, or other object such as a lens, is first washed with water and a gentle abrasive cleanser to remove surface contamination, particularly the grease which prevents uniform wetting. The clean glass is given a half-minute dip in an 0.5 per cent hydrofluoric acid solution to remove glass altered by atmospheric action. The rinsed and drained sheet of glass is immersed in the filming bath for the required time. The time varies from 30 to 90 minutes at 45°C, as can be seen from Figure 1, on which several brands of glass are designated by numbers, identified in Table I. The glass sheet is taken from the bath, rinsed with water, allowed to drain, and is usually wiped dry with cheesecloth.

(b) *Hydrofluoric Acid Dip*

Glass surfaces undergo chemical alteration known as "weathering" from exposure to air. These weathered surfaces behave erratically in the filming process, resulting in non-uniform film, or unfilmed, or fogged areas. The hydrofluoric acid dip dissolves a layer of glass, including the weathered portion.

Table I — Sheet Glasses Studied

Manufacturer "A"

1. Picture Glass
2. Window Glass
3. Polished Plate
4. "Clear" Polished Plate

Manufacturer "B"

5. Picture Glass
6. Window Glass
7. Polished Plate

(c) *Temperature*

Each 10°C rise in temperature doubles the filming rate, as shown in Figure 1. To make the filming time as short as possible, the temperature is chosen as high as other factors permit. As temperature increases, losses from the solution increase, both by evaporation and by silica precipitation. Both losses change the solution composition, and thereby impair the processing control.

(d) *Types of Glass*

This process is practically limited to glasses having compositions which class them as "crown" glasses. The common window and polished plate glasses are in this classification. Pyrex glass, for example, is not ordinarily attacked by the processing bath, and is, therefore, a suitable container material.

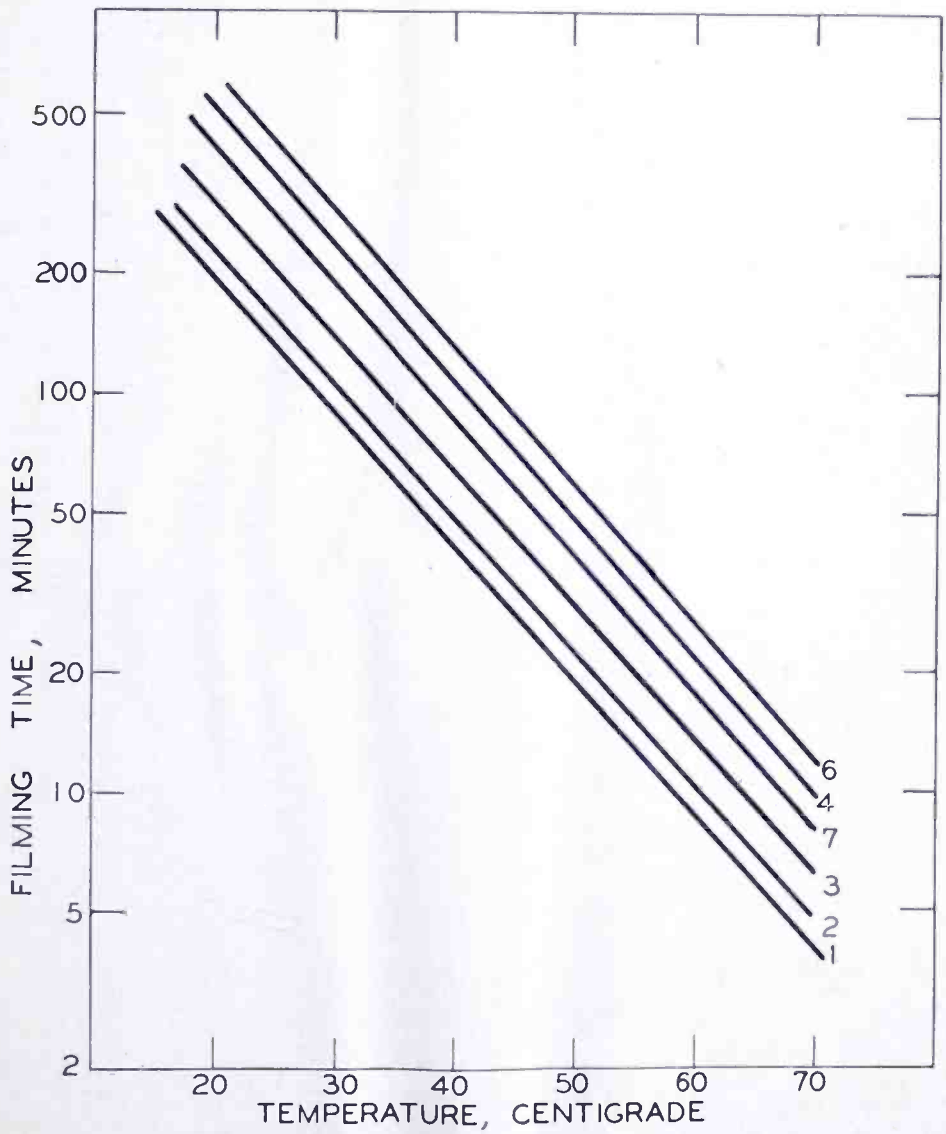

Fig. 1—Effect of temperature on filming time for six typical sheet (plate and window) glasses.

The nature of the glass, particularly its composition and perhaps its thermal history, determines (1) the processing time, after concentration and temperature have been fixed; and (2) the potency value at which the bath will function.

Most of the experimental work was done on sheet glass, because of the convenience in handling test samples. The process functions equally well on objects of other shapes, such as lenses or prisms.

(e) *Potency Adjustment to Match Glass*

The failure of the bath to produce a film on a particular glass indicates that the potency requires adjustment (or that the glass is one not capable of being filmed by this process). To determine in what direction the potency should be changed, and by how much, test tube trials are made. Some of the solution is put into a number of test tubes, and boric acid or potassium fluoride is added to each, to provide a range of potencies, two units apart. These test tubes are brought to the operating temperature, and strips (or pieces) of the glass are introduced. The glass strips are observed for film formation. If film appears on the glass in one or more tubes, the lowest of the potency values in such tubes is taken as the value to which the bath is to be adjusted. The approximate filming time is also observed from this test. Further minor adjustment in potency is generally required, based on actual processing in the nearly-adjusted bath.

(f) *Control of Film Properties*

The filming bath produces an observable film over a range of potency values (3 to 5 units). The properties of the film, however, change with the potency of the bath. As may be seen from Figure 2, the reflection reaches a minimum value at the middle of the potency range.

Films produced at this optimum potency reflect less than 1 per cent (original glass reflection = 100 per cent) of filtered green light, and about 5 per cent of white light. The index of refraction is nearly equal to that required by theory,[6] the square root of that of the glass. Weighings have shown that about 50 per cent of the glass (by weight) in the film volume dissolves during film formation. There has been some silica lost and presumably all the soda and lime (crown glasses are approximately 70 per cent silica).

As the potency diminishes from the optimum, less material is removed from the glass, the refractive index increases, and the film becomes more resistant to abrasion. In the other direction, as potency increases, more material is removed from the glass, the refractive index diminishes, and the film becomes more fragile.

Film properties may be chosen to suit particular applications. Minimum reflection is obtained for any wavelength of light at any viewing angle, by changing the filming time (up to about 10 per cent) in either direction. Longer time moves the extinction wavelength

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[6] K. B. Blodgett, "Use of Interference to Extinguish Reflection of Light from Glass", *Phys. Rev.*, Vol. 55, No. 4, pp. 391-404, February 15, 1939.

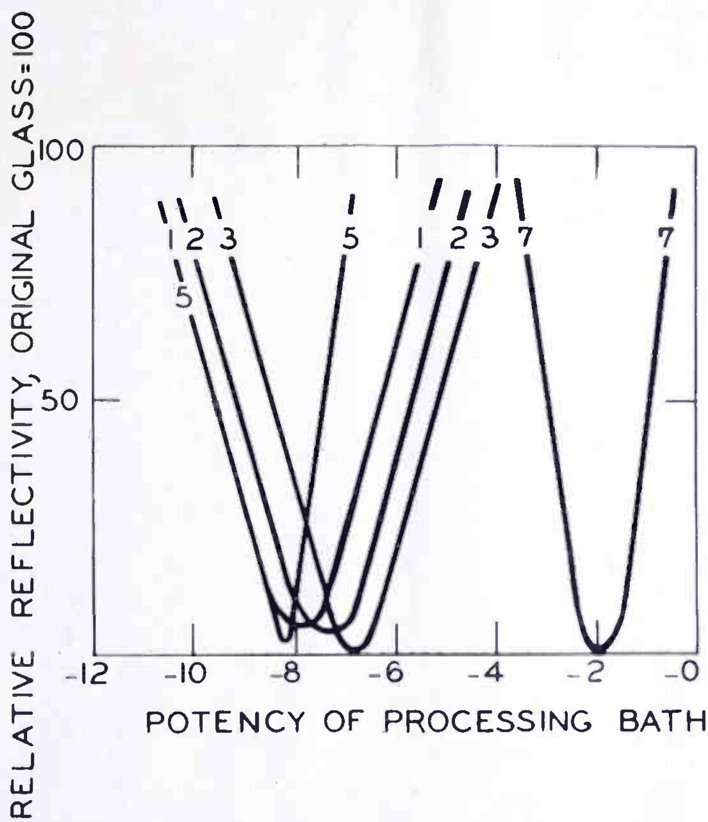

Fig. 2—Effect of potency of the filming bath on film reflectivity for five typical sheet (plate and window) glasses. Light filtered through green filter (wratten 58).

toward the red, and for a given wavelength, makes the viewing angle for extinction greater than 0° (normal incidence). Films with greater resistance to abrasion and contamination are made at the expense of increased reflectivity, either by filming at lower potency values than optimum, or by filming for about half the usual time.[7]

Care of the Film

Freedom from chemical deterioration is assured in most applications of the film, because of the chemical inertness of silica. Abrasion resistance of the film is ample to prevent damage by ordinary handling and washing.

Contamination of the film increases the reflection from it. Even small increases are noticeable, because of the low initial reflection. Therefore, like other glass surfaces used for optical transmission, the filmed glass surfaces must be kept clean. Sheltered films retain their low reflection; those exposed to circulating air need occasional cleaning with a non-alkaline, non-abrasive detergent.

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[7] F. H. Nicoll, U. S. Patent 2415703, February 11, 1947.

RCA TECHNICAL PAPERS†

Fourth Quarter, 1950

Any request for copies of papers listed herein should be addressed to the publication to which credited.

"An Achromatic Microwave Antenna", N. I. Korman and J. R. Ford, *Proc. I.R.E.* (December) 1950

"Amplifiers", L. E. Barton, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950

"Antennas", E. A. Laport, Section of RADIO ENGINEERING HANDBOOK, (Henney) McGraw-Hill Book Co., New York, N. Y., 4th Edition 1950

"Application of Organic Films to Cathode-Ray Tube Screens by Flotation Methods", S. M. Thomsen, *RCA Licensee Bulletin, LB-807* (November 20) 1950

"Automatic Audio Gain Controls", J. L. Hathaway, *Audio Eng.* (October) 1950

"An automatic Nonlinear Distortion Analyzer", H. F. Olson, *RCA Licensee Bulletin LB-812* (December 10) 1950

"Base-Metal Effects in Thoria-Coated Filaments", H. Nelson, *Jour. Appl. Phys.* (November) 1950

"A Cathode Controlled Multivibrator for the Production of Short Waves", G. W. Gray, *RCA Licensee Bulletin LB-814* (December 15) 1950

"Cathode-Ray Tubes", L. E. Swedlund, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th edition 1950

"A Center-Tuning Circuit for FM Detectors", E. O. Keizer, *RCA Licensee Bulletin LB-815* (December 28) 1950

"Characteristics and Operation of an RCA Developmental Three-Gun Tri-Color Kinescope", *RCA Licensee Bulletin LB-808* (December 5) 1950

"Crystal Diodes in TV Studio Equipment", R. Kuehn, *Electronics* (December) 1950

"The Cultural Properties and Pathogenicity of Certain Microorganisms Isolated from Various Proliferative and Neoplastic Diseases", V. Wuerthele-Caspe, E. Alexander-Jackson, J. A. Anderson, J. Hillier, R. M. Allen and L. W. Smith, *American Journal of Medical Sciences* (December) 1950

"Deflection of Cathode-Ray Tubes in Sequence", G. W. Gray and A. S. Jensen, *RCA Review* (December) 1950

"Design of High-Pass, Low-Pass and Band-Pass Filters Using R-C Networks and Direct-Current Amplifiers with Feedback", C. C. Shumard, *RCA Review* (December) 1950

"Designing the Bridgeport UHF Antenna", R. M. Scudder, *Electronics* (November) 1950

"Detectors", V. D. Landon, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York N. Y., 4th Edition 1950

"Developmental Tri-Color Kinescope Receivers for the RCA Color Television System", *RCA Licensee Bulletin LB-811* (December 5) 1950

"Distortion in Multichannel Frequency-Modulation Relay Systems", L. E. Thompson, *RCA Review* (December) 1950

"Effect of Coating Composition of Oxide-Coated Cathodes on Electron Emission", E. G. Widell, and R. A. Heller, *Jour. Appl. Phys.* (November) 1950

† Report all corrections or additions to RCA Review, Radio Corporation of America, RCA Laboratories Division, Princeton, N. J.

"Effect of Infra-Red on Emission and Trapping in ZnS:Cu Phosphors", R. H. Bube, *Phys. Rev.* (November 15) (Letter to the Editor) 1950
"Effects of Contact Pressure on Transistor Gain", J. I. Pantchechnikoff, *Licensee Bulletin LB-805* (October 25) 1950
"Electroacoustic Equipment", R. J. Kowalski, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"The Electron Microscope in Industrial Research", J. Hillier, Section of SECOND SYMPOSIUM OF VARNISH AND PAINT INDUSTRY, New York University 1950
"Electron Microscopy", J. Hillier, Section of BIOPHYSICAL RESEARCH METHODS, F. M. Uber, Interscience Publishers, Inc., New York, N. Y. 1950
"Electron Microscopy of Microorganisms and Viruses", J. Hillier, Section of ANNUAL REVIEW OF MICROBIOLOGY, Volume IV .. 1950
"Electron Optics", D. W. Epstein, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Electronic Reading Aid for the Blind", V. K. Zworykin, L. E. Flory and W. S. Pike, Section of MEDICAL PHYSICS, Volume II, O. Glaser, Yearbook Publishers, Inc., Chicago, Ill. 1950
"Electronics for Chemistry", G. A. Morton and M. L. Schlutz, ENCYCLOPEDIA OF CHEMICAL TECHNOLOGY, Interscience Publishing Corp., New York, N. Y. 1950
"Facsimile Transmission and Reception", Maurice Artzt, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition. 1950
"Ferrite-Core Yoke for Wide Deflection Angle Kinescopes", M. J. Obert and W. A. Needs, *Tele-Tech* (October) 1950
"A Floating Double Probe Method for Measurements in Gas Discharges", E. O. Johnson and L. Malter, *Phys. Rev.* (October) 1950
"Frequency Modulation", J. E. Young, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Gaseous Discharge Noise Sources for SHF", K. R. De Remer and H. Johnson, *RCA Licensee Bulletin LB-803* (October 12) 1950
"Geometrical Considerations of an RCA Tri-Color Kinescope", *RCA Licensee Bulletin LB-809* (December 5) 1950
"Geometrical Optics", D. W. Epstein, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Getter Materials for Electron Tubes", M. Knoll, W. Espe and M. P. Wilder, *Electronics* (October) 1950
"A High Performance Transistor with Wide Spacing Between Contacts", B. N. Slade, *RCA Licensee Bulletin LB-804* (October 16) .. 1950
"A High-Performance Transistor with Wide Spacing Between Contacts", B. N. Slade, *RCA Review* (December) 1950
"A High-Quality Sound System for the Home", H. F. Olson and A. R. Morgan, *Radio and Television News* (November) 1950
"Improved Sync Separation in Television Receivers in the Presence of Impulse Noise", E. S. White and J. Avins, *RCA Licensee Bulletin LB-813* (December 15) 1950
"Insulating Materials", W. R. Dohan, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Lighting Methods for Television Studios", H. M. Gurin, *Jour. S.M.P.T.E.* (December) 1950
"Loudness Balance Methods for Earphone Response Measurements", D. W. Martin and M. L. Touger, *Jour. Acous. Soc. Amer.* (November) 1950

"Luminescent Solids (Phosphors)", H. W. Leverenz, Section of COLLOID CHEMISTRY, THEORETICAL AND APPLIED, J. Alexander, Reinhold Publishing Corp., New York, N. Y. 1950
"Luminescent and Tenebrescent Materials", H. W. Leverenz, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McElwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Luminescence and Trapping in Zinc Sulfide Phosphors with and without Copper Activator", R. H. Bube, *Phys. Rev.* (November 15) .. 1950
"A Magnetic Record-Playing Head", M. Rettinger, *Jour. S.M.P.T.E.* (October) 1950
"Magneto-Optic Transducers", A. W. Friend, *RCA Review* (December) 1950
"Mass Production Test and Alignment", R. G. Peters, *TV Engineering* (December) 1950
"Methods of Calibrating Frequency Records", R. C. Moyer and D. R. Andrews, *Proc. I.R.E.* (November) 1950
"Microscopy: Electron", V. K. Zworykin and J. Hillier, Section of MEDICAL PHYSICS, Volume II, O. Glaser, Year Book Publishers, Inc., Chicago, Ill. 1950
"Minimizing Pulse Voltages in Television Vertical-Deflection Amplifiers", *RCA Application Note AN-146*, RCA Tube Department, Harrison, N. J. (December 1) 1950
"Modulators", J. E. Young, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Motion Pictures and Television", V. K. Zworykin, *Jour. S.M.P.T.E.* (December) 1950
"New Developments in Radar for Merchant Marine Service", C. E. Moore, *RCA Review* (December) 1950
"New Limiting Amplifier", G. A. Singer, *Audio Eng.* (November) 1950
"New Worlds for Study", J. Hillier, Section of A.A.A.S. CENTENNIAL VOLUME 1950
"Pencil-Type Triodes RCA-5675 and RCA-5876", *RCA Application Note AN-147*, RCA Tube Department, Harrison, N. J. (December 1) 1950
"Performance and Operation of a New Limiting Amplifier", G. A. Singer, *Audio Eng.* (November) 1950
"Power Supply", J. E. Young, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Projection Practices", J. D. Phyfe, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Public-Address Systems", R. J. Kowalski, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Radio Receivers", V. D. Landon, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Radio Telegraph Systems", J. L. Finch, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Radio Transmitters", J. E. Young, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Recording Practices", O. B. Gunby, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Sensitivity, Directivity and Linearity of Direct Radiator Loudspeakers", H. F. Olson, *Audio Eng.* (October) 1950
"Sheath Formation in Ion-Neutralized Electron Beams", E. G. Linder, *Phys. Rev.* (October) (Letter to the Editor) 1950

"Single-Mesh and Coupled Circuits", V. D. Landon and K. McIlwain, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y. 4th Edition 1950
"A Small Electron Microscope", J. H. Reisner and E. G. Dornfeld, *Jour. Appl. Phys.* (November) 1950
"Space-Charge Effects in Electron Beams and Their Reduction by Positive Ion Trapping", E. G. Linder and K. G. Heinqvist, *Jour. Appl. Phys.* (November) 1950
"Special-Purpose Amplifiers", E. L. Clark, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Television Broadcasting", T. J. Buzalski, A. L. Hammerschmidt and F. J. Somers, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Television Pick-Up Tubes", V. K. Zworykin and E. G. Ramberg, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Testing and Aligning Video Amplifiers", F. E. Cone and N. P. Kellaway, *Tele-Tech* (November) 1950
"Thermionic Vacuum Tubes", A. P. Kauzmann, Section of ELECTRICAL ENGINEERS' HANDBOOK, H. Pender and K. McIlwain, John Wiley & Sons, Inc., New York, N. Y., 4th Edition 1950
"Transmitter Diversity Applied to Machine Telegraph Radio Circuits", G. E. Hansell, *RCA Review* (December) 1950
"A Tristimulus Photometer", G. C. Sziklai, *Jour. Opt. Soc. Amer.* (November) 1950
"Video Relay Switching", C. R. Monro, *TV Engineering* (December) .. 1950
"Video Special-Effects System", E. M. Gore, *TV Engineering* (October) 1950
"V. K. Zworykin", E. W. Engstrom, Section of LES INVENTEURS CELEBRES, Lucien Mazenod 1950

NOTE—Omissions or errors in these listings will be corrected in the yearly index.

Correction:

In the paper "Design of High-Pass, Low-Pass and Band-Pass Filters using R-C Networks and Direct-Current Amplifiers with Feedback", by C. C. Shumard, which appeared on pages 534-564 of the December 1950 issue of *RCA Review*, Figure 3 on page 535 is incorrectly captioned. The caption should read "Band Pass" rather than "High Pass".

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