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PART 7

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Audio

Broadcast and Television Receivers

Broadcast Transmission Systems

at

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1957 IRE WESCON CONVENTION RECORD

PART 7 - AUDIO AND BROADCAST

TABLE OF CONTENTS

The Traveling-Wave VHF Television Transmitting Antenna*

MATTI S. SIUKOLA†

Summary—A new simple and rugged television antenna has been developed for vhf high channels. The antenna utilizes pairs of slot radiators cut longitudinally in a vertical pipe. In each pair the slots are fed in opposite phase, and the pairs are displaced one-quarter wavelength from each other along the pole. Every other pair is in one vertical plane and the remaining pairs in another perpendicular to this. The slots are fed with a traveling wave within the pole.

The excellent characteristics, most of them inherent in the principle derive from the unique combination of supporting structure, transmission line, and radiators. The horizontal pattern is formed on a turnstile principle from excellent patterns of single slot pairs even at large pole diameters. The smooth vertical pattern which helps to achieve uniform field strength within the service area is based on exponentially decaying traveling-wave feed. This is especially important in high gain applications. Proper electrical characteristics of simple slot radiators maintain constant wavelength instead of constant velocity in the transmission line for frequencies in the channel. This assures proper bandwidth. Low input vswr is due to the traveling-wave nature of the feed. The construction of the antenna is particularly suitable for handling high power. Simplicity of the antenna and low wind load result in economy of both the antenna and the supporting structure.

THE superturnstile antenna has been the standard antenna of the television broadcast industry for about a decade. However, already in 1952 when the present high-gain superturnstile antennas were still under development, it was foreseen that a newer design was needed both to improve on the superturnstile antenna and also to meet the higher standards required by improvement in the state of art.

Although superturnstile antennas perform satisfactorily and are very reliable, the trend toward higher power and higher gain has introduced increasing complexity. Also, with even moderately high-gain antennas it was realized that the uniform current distribution used resulted in field intensity patterns having deep nulls in parts of the service area. This effect has been overcome by adjusting the current distribution along the antenna, producing a contoured vertical pattern. The best current distribution obtainable, however, fell considerably short of ideal, because the variation was limited by the number of radiators along the length of the antenna. Furthermore, as the gain was increased the feed system grew more and more complicated. This increased the cost of the antenna, and at the same time the large diameter of the pole and feed system had an effect on the horizontal circularity of the pattern.

Since the maximum economy of operation of the television transmitter plant for a prescribed erp is obtained by taking advantage of high antenna gains, these factors in the vhf range became especially important at channels 7 through 13, where three times as much erp is permitted as on channels 2 through 6, and where the plant becomes more expensive to build and to operate. Therefore a research program was undertaken to develop a vhf high-band antenna which would permit simplicity in its mechanical construction and feed system, in addition to being capable of ideal radiation characteristics for low, medium, and high-gain usage at power levels required in the foreseeable future. This led to evolution of a new concept in television transmitting antenna design and to the development of the traveling-wave antenna, which should be immediately attractive in the trend toward high gain.

Description

The traveling-wave antenna utilizes slot radiators in the form of a tube with longitudinal slots cut around the periphery. In its simplest form the antenna consists of three parts: input and supporting section, main aperture, and top loading (see Fig. 1).

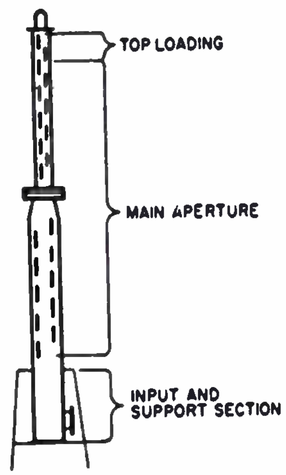

Fig. 1—Traveling-wave antenna.

The tube or the outer shell of the antenna forms the outer conductor of a large coaxial transmission line. Power is fed into the buried section of the antenna near the bottom end by a tee section to facilitate good rigid mechanical support for the inner conductor of the antenna.

In the aperture area of the antenna there are four rows of slots used in pairs, one opposite another, and capacitively coupled to the inner conductor.

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The separation between the pairs along the pole is approximately a quarter wavelength. Every other pair is in one vertical plane, and the remaining pairs in another vertical plane perpendicular to this. All slots in the main aperture are alike and all coupling capacitors or probes the same size.

At the top of the antenna the inner conductor is short-circuited to the outer conductor. Provision is made to allow for differential expansion between the pole and the inner conductor due to temperature variations. The last two pairs of slots differ in characteristics and also in dimensions from the other slots and similarly the coupling networks are different. These two slot pairs including coupling networks and the short are called the top loading.

Development

The principle of operation of the traveling-wave antenna is clearly demonstrated by a short description of the evolution of the antenna. The primary objective was to develop a vhf high-channel antenna for omnidirectional service, which would have excellent performance at low cost especially at medium and high-gain values. This requires a simple construction of the antenna.

In the traveling-wave antenna an extremely simple method of feed is employed. The supporting pole forms the outer conductor of a coaxial transmission line, and power is tapped off from the line at each radiator thus eliminating the need of any feed lines.

Fig. 2 shows the principle of the feed system using short rod radiators to illustrate the theory. A number of radiators per wavelength uniformly spaced are loosely coupled to a coaxial line. Because of the number of radiators and the relatively slight reflection due to each, the effect is essentially that of a uniform loading. The result is a uniformly attenuated traveling wave in the line. Since a traveling wave has a linear phase characteristic, the excitation of each successive radiator will be lagging from the previous one by an amount which depends upon the spacing between the radiators and the velocity of propagation in the line. If the radiators are alike, their currents will have the same phase relationship as the excitation. Thus the radiating currents will be successively lagging, and repetition of phase occurs after every guide wavelength.

To obtain an omnidirectional pattern the radiators, instead of being in line, can be moved around the periphery to form a "spiral" as shown in Fig. 3. For a horizontal main beam the pitch of the spiral has to be equal to the guide wavelength in the transmission line. In this arrangement, all of the radiators in any one vertical plane on one side are in phase, and the phase difference between radiators in different planes equals the azimuth angle difference between the planes; that is, the phase rotates around the periphery. The rotating phase produces a rotating field which, because of the relatively small amount by which the magnitude of current changes from layer to layer, produces an omnidirectional pattern.

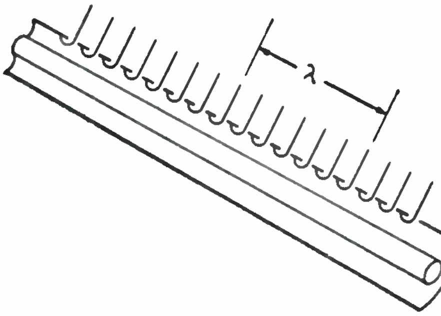

Fig. 2.

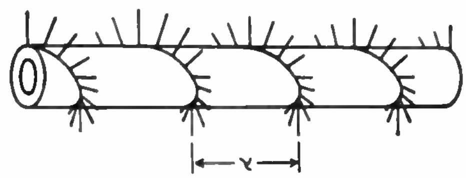

Fig. 3.

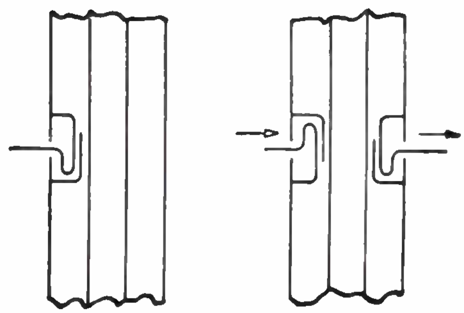
Fig. 4.

In practice, however, to insure proper operation two modifications have to be made. First, excitation has to be made symmetrical instead of asymmetrical in order to avoid undesired excitation of the pole itself (see Fig. 4). Second, mutual coupling between the radiating elements has to be minimized to maintain proper phase relationships.

Both results can be accomplished by limiting the number of radiator pairs to four per turn of spiral. This modification reduces the antenna to two sets of center-fed dipoles, the sets being confined to two perpendicular vertical planes fed in phase quandrature (see Fig. 5). The traveling-wave nature of the signal in the line remains, however, since the relatively small reflections produced by the radiator pairs for the most part still cancel each other, not because of uniform loading by radiator pairs but because of the approximately quarter wavelength spacing between them. We thus have, in essence, a traveling-wave-fed turnstile antenna.

By this evolution a simple feed system capable of high-power handling capacity has been obtained. However, horizontal radiation pattern formed on turnstile principle from dipole patterns tends to higher noncircularity at higher gain values. Therefore, to further improve the antenna, the dipoles can be replaced by pairs of slots which are fed in opposite phase to produce a dipole type of radiating current (see Fig. 6). This improves the

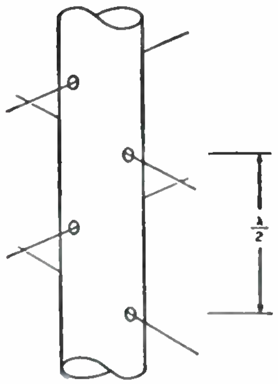

Fig. 5.

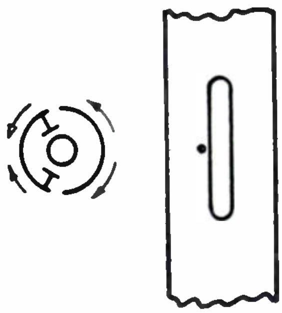

Fig. 6.

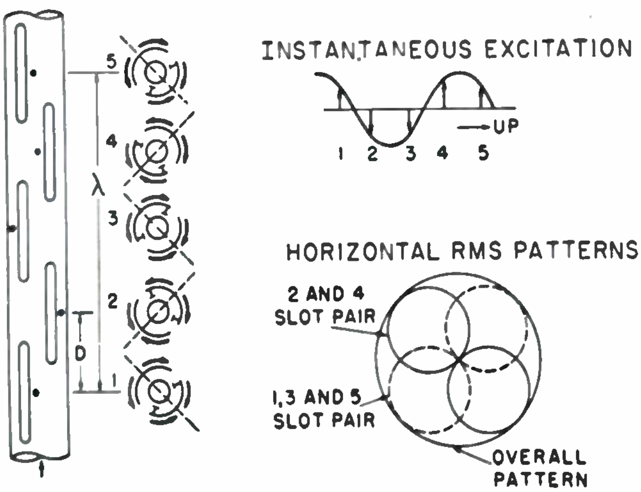

Fig. 7—Excitation and horizontal pattern formation of traveling-wave antenna.

horizontal pattern by proper radiating-current distribution. Additionally, the wind loading is reduced by eliminating protruding radiators, and the coupling circuit simplified by allowing capacitive coupling. Thus, the final basic form of a section of the traveling-wave antenna has been derived (see Fig. 7). Except for the top loading the antenna is a traveling-wave-fed turnstile antenna which utilizes pairs of slot radiators.

Operation and Electrical Characteristics

Among the several important characteristics accomplished in the traveling-wave antenna are the excellent electrical features listed below.

Improved Horizontal Pattern Circularity

In the transmission line within the antenna the signal propagates upwards. Along the main aperture each slot pair successively extracts a portion from the traveling wave. Since the slots in each pair are fed in opposite phase, the radiating current on the outside of the pole simulates that of a dipole radiator. Therefore, an approximately figure eight horizontal pattern is produced (see Fig. 7). The slot pairs in one vertical plane, that is, for instance, pairs 1, 3, and 5 each form a figure eight pattern at the same azimuthal orientation. Since these pairs are spaced at every half wavelength along the pole, and the coupling capacitors alternate to counteract the reversal of the excitation voltage within the line, the radiating currents as well as the fields are in phase. Thus, all the fields from slot pairs in one vertical plane form a combined figure eight horizontal pattern. Similarly, the slot pairs in the vertical plane 90° from it radiate with a figure eight horizontal pattern but 90° displaced from the previous one. Since the spacing (D) between successive slot pairs is a quarter wavelength, the radiated fields are in phase quadrature. According to turnstile principle, these two figure eight patterns added in phase quadrature form a circular horizontal pattern.

The horizontal pattern of the traveling-wave antenna is, of course, only ideally a perfect circle. The physical size of the pole and the finite width of the slots result in a slightly scalloped shape. Representative measurements made at channel 10 by using an 18-inch diameter, 30-foot-long antenna section, which corresponds to the lower sections of an antenna for a gain of approximately 18, gave a circularity of ± 0.7 db (see Fig. 8).

Thus by using pairs of slot radiators, with successive pairs in space and phase quadrature, a radiating-current distribution is obtained so that even for large pole diameters a circular pattern is inherently produced. In addition, there are no interfering external elements in the field. The antenna combines radiating elements, feed system, and the supporting structure in one unit, giving excellent horizontal circularity.

Low-Voltage Standing-Wave Ratio

As has been mentioned previously, each slot pair in the main aperture successively extracts a small portion of the energy from the traveling wave. Therefore, the signal in the transmission attenuates progressively. Since all the slots in the main aperture are alike and all the coupling capacitors of the same size, the loading at each slot pair is the same, effecting an equal attenuation of the signal along the line. This equal loading of the slots results in exponentially tapered signal within the transmission line as well as exponentially tapered illumination of the aperture (see Fig. 9).

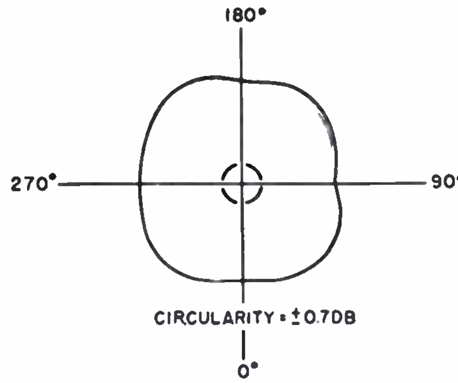

Fig. 8—Measured horizontal pattern of bottom quarter of gain of 18 antenna. 18-inch diameter, 30-foot length, channel-10.

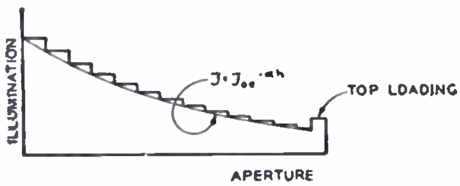

Fig. 9.

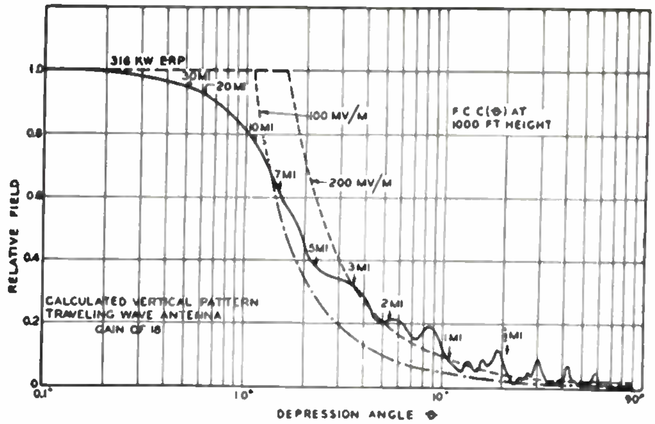

Fig. 10.

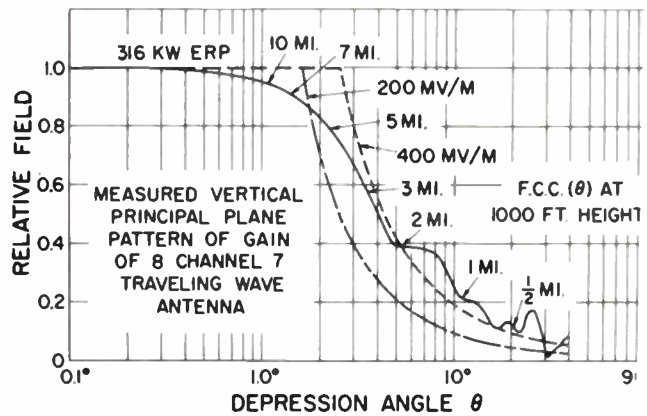

Fig. 11.

When the signal reaches the top of the antenna, all but a small percentage of the energy has been extracted and radiated. The remaining energy is radiated by the top loading. The top loading forms a relatively broad-banded, low-power, low-gain portion of the antenna which radiates omnidirectionally and in phase with the main aperture. Besides contributing to the radiation, the top loading also provides a proper termination for the main aperture. This termination in conjunction with the quarter-wave spacing between the slots and the light coupling of the slot pairs into the line keeps the reflections small and thus maintains the traveling-wave nature in the transmission line.

The traveling-wave nature of the feed already reveals that a low vswr exists all along the antenna except for the top loading. This characteristic inherently gives the antenna a good input vswr without any compensating or matching devices. Further improvement is obtained by adjusting the input slot pair to half the loading of the others. The input tee has been broad banded to provide a smooth transition from the transmission line to the antenna. Measurements made on a representative section of a gain of 20 channel-10 antenna showed the vswr to be equal to or less than 1.036 to 1 across the channel. At lower gains it is slightly higher.

Improved Vertical Pattern Without Substantial Loss of Gain or Simplicity of the Antenna

Most broadcasters prefer a vertical pattern which provides a uniformly high field strength over the service area. In this respect the traveling-wave antenna is outstanding.

In the simplest form of the antenna, when all the slots are alike and coupling capacitors are of the same size, the attenuated traveling wave in the transmission line and the illumination of the aperture decrease exponentially, thus providing an automatic power division (see Fig. 9). To take full advantage of.the excellent pattern characteristics of this principle, the antenna would have to be tall enough for the energy in the line to diminish to a very low value, and the antenna should not have any discontinuities in the illumination of the aperture. However, the antenna, for the same value of gain, can be shortened up to one third, without any appreciable deterioration of the performance. This is accomplished by the top loading, which collects and radiates all of the remaining power, producing an increase of illumination at the top end of the antenna. If no beam tilt is employed, the relative phase along the whole aperture is zero. Because an exponential distribution of illumination with linear phase produces an extremely smooth null-less pattern, then even by cutting the infinite antenna to a finite length, by using the top loading, and leaving the slots off at proper places to accommodate the flange joints, a vertical pattern is obtained which still is almost ideal. It provides the service area at most locations with a uniformly high field strength. Figs. 10 and 11 show vertical patterns of two traveling-wave antennas of different gains. The measured patterns have very closely approximated the theoretical calculated patterns.

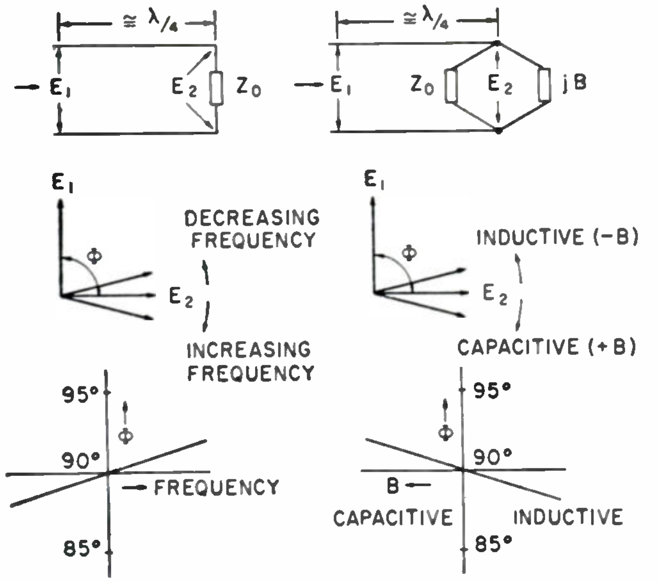

Fig. 12—Phase delay as a function of freq. 8 shunt loading in λ/4 transmission line.

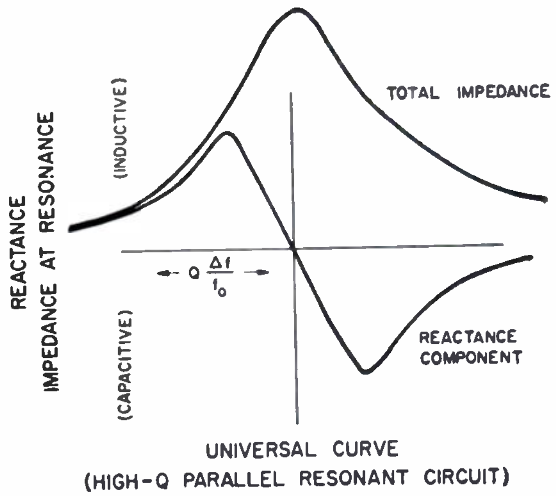

Fig. 13.

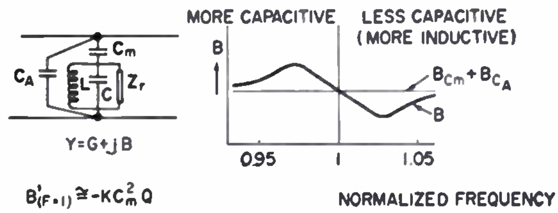

Fig. 14.

As is shown, the traveling-wave antenna possesses an inherently excellent vertical pattern, without utilizing any complicated phase or artificial power distribution along the aperture as is often needed in other types of antennas to achieve such a null-less pattern and still maintain a high gain per wavelength of aperture.

Excellent Pattern Bandwidth

For an antenna to provide distortion-free television service, proper phase relationships between the radiators have to be maintained across the channel. For high-gain antennas it is customary to excite separate radiators, or at least small groups of radiators through feed lines of equal length to maintain radiating current phases equal enough to have essentially the same radiation pattern across the channel. This means that a number of feed lines are needed, and many of the lines are longer than would be mechanically necessary. This results in a complex construction.

In the traveling-wave antenna the phase relationships have been maintained by a very unique design of the antenna. The principle used can be demonstrated by referring to two properties of the transmission lines (see Fig. 12). If a quarter-wave-long transmission line is terminated to its characteristic impedance, then the phase difference (ϕ) between the input and output voltages varies linearly with frequency (Fig. 12, left). On the other hand, if the same transmission line, besides being terminated to its characteristic impedance, is also shunted with a varying susceptance (B) and at a fixed frequency the value of susceptance is varied, the phase difference between the input and output voltages again varies (Fig. 12, right). Within a certain range the variation of the phase with respect to the variation of the susceptance is approximately linear.

If these two properties are combined so that the phase changes caused by varying frequency and varying susceptance cancel each other, a quarter wavelength of transmission line with a constant wavelength has been obtained. A shunt loading with a negative susceptance slope of proper magnitude vs frequency is required. This can be obtained by use of a parallel resonant circuit (see Fig. 13). Close to the resonance frequency the reactance of a parallel resonant circuit possesses a relatively linear negative slope. If the parallel resonant circuit is capacitively coupled to a transmission line, this negative reactance slope results in a negative slope of the susceptance imposed on the line (see Fig. 14). The magnitude of the susceptance slope is determined by the size of the coupling capacitor and the Q of the resonance circuit, as indicated by the equation. The coupling capacitance is very closely determined by the gain of the antenna; therefore, the only independent variable is the Q of the circuit.

This type of capacitively coupled parallel resonant circuit is obtained by utilizing capacitively coupled slot radiators as previously discussed. Fig. 15 shows the equivalent circuits of slot pairs displaced an equal distance (D) from each other. This spacing is slightly less than a quarter wavelength. The curve indicates the type of phase compensation obtained. The high slope linear phase delay of the slightly less than a quarter-wavelength-long transmission line is first modified by an approximately constant phase delay due to the coupling capacitors (C_m) and any additional shunt capacitors (C_A). Additionally, the varying susceptance of the parallel circuit changes

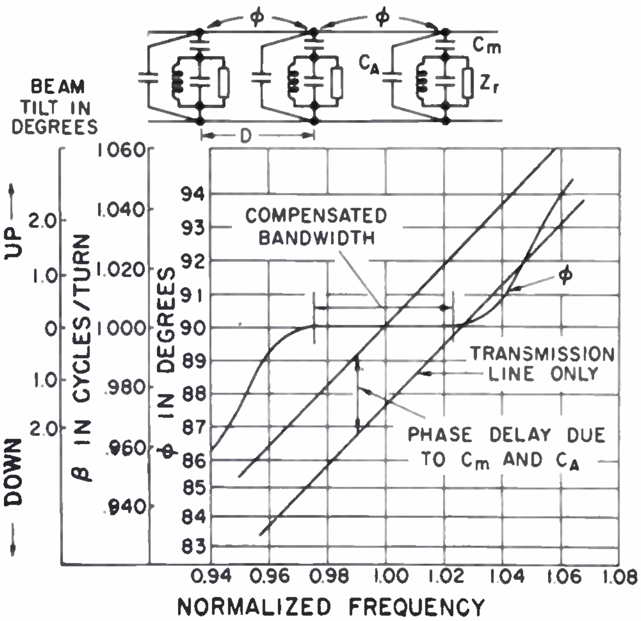

Fig. 15.

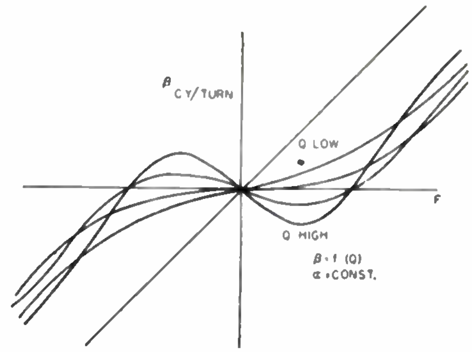

Fig. 16.

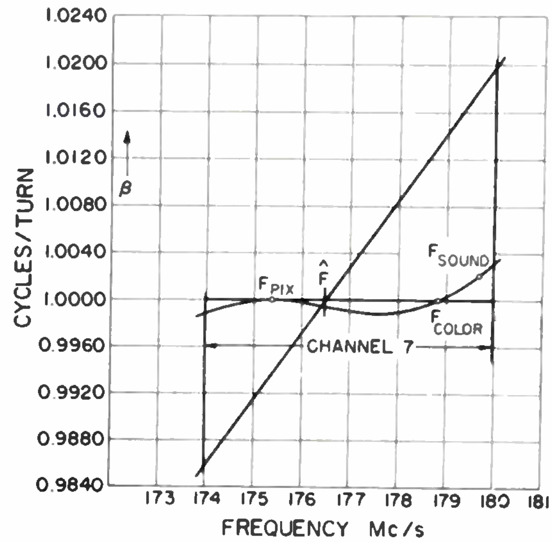

Fig. 17.

the phase delay. By selecting a proper value of Q for the slot pairs a constant phase delay about the desired frequency can be maintained. Outside of the compensation region the phase delay approaches asymptotically that of the transmission line modified by capacitors only. Thus a compensated bandwidth has been obtained, wherein the phase difference between the input and output voltages, that is, from one slot pair to the next, remains constant, normally 90°.

The phase progression is often measured in other units such as cycles per turn or pairs of slots and then called phase propagation function. Also, the phase difference can be changed to beam tilt produced by it. A 90° phase delay per slot spacing results in a horizontal main beam. With increasing phase delay, the beam tilts upwards, and decreasing delay produces downward tilt.

Since the slope of the phase delay curve can be changed by varying the Q, different degrees of compensation can be obtained (see Fig. 16). A low Q results in a positive slope of the curve, and a high Q, in a negative slope. For a specific design a Q of the slot radiators can be selected which produces minimum variation of the phase propagation function, that is, a stable beam. This, however, may not be the correct value for best operation of the antenna. This is illustrated by curves of a gain of 9 channel-7 traveling-wave antenna. Fig. 17 shows the best phase propagation function (β) for the antenna design in question. The high sloping straight line again indicates the phase function of the transmission line including only the effect of the capacitors. As can be seen, the maximum compensation has not been utilized, but phase delay at the upper edge of channel has been allowed to rise slightly. Also, the design frequency ($\hat{F}$) has not been placed at the middle of the channel but slightly towards the picture carrier. Both of these have been selected because, at the same time as the phase propagation function varies, also the attenuation (α) within the line and thus the illumination of the aperture vary somewhat as shown by Fig. 18. The maximum attenuation occurs at the design frequency, and the attenuation decreases gradually at each side of it. To select a proper Q for the radiators, the phase propagation function and the attenuation function have to be transformed to vertical patterns of the antenna at various frequencies. From the vertical patterns the video frequency response of the antenna can be calculated as detected by an ideal receiver. If the analysis of the video frequency response is performed for several values of Q, the optimum value of Q can be determined. In the example of gain of 9 channel-7 antenna, the video response at the major service area varies less than $\pm 1/2$ db.

Thus, although in the transmission line the velocity of propagation is normally constant and the wavelength in the line varies with the frequency, by imposing loadings with proper characteristics on the line, the wavelength in the line can be made constant within a certain bandwidth. This has been done in the traveling-wave antenna by making characteristics of the slots such as to maintain a

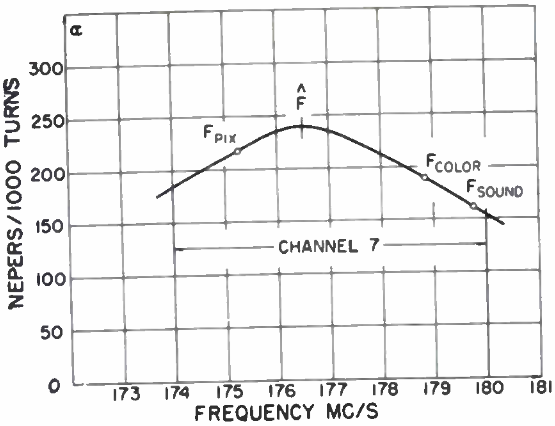

Fig. 18.

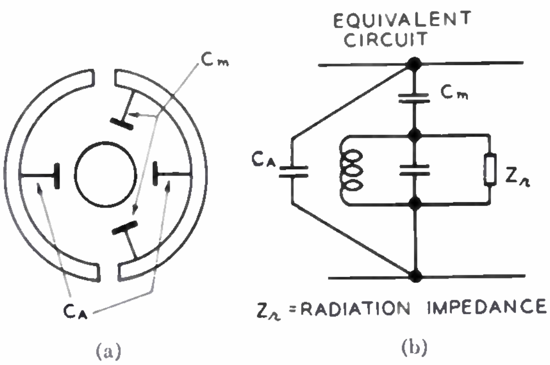

Fig. 19.

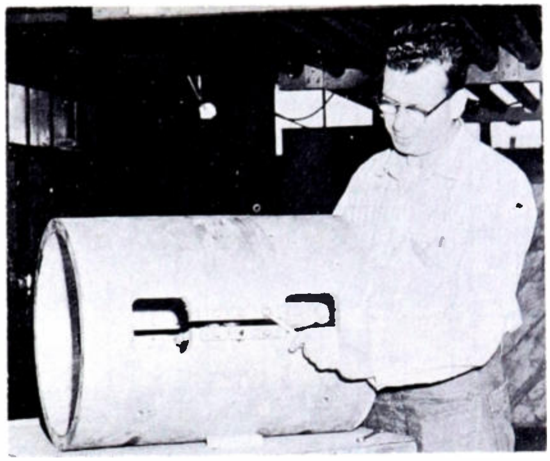

Fig. 20.

Fig. 21.

constant wavelength in the line across the channel in question, thus insuring proper phase of the slots. In vhf television service this type of compensation is necessary if the feed path lengths to different radiators vary appreciably.

In connection with Fig. 15, it was mentioned previously that, by placing a capacitor parallel to the slot loading, an approximately constant change of phase can be obtained. This has been utilized in the traveling-wave antenna to provide for beam tilting and fine tuning of the phase propagation function during development. In practice, this capacitance is in the form of a pair of probes or phase compensators (C_A) in the neutral plane of each layer of the main aperture (see Fig. 19).

The required value of the coupling capacitors (C_m) is essentially determined by the gain of the antenna.

The desired value of Q of the slot pairs can be obtained by determining proper shape and length, in particular, for the slots. For range of gains from 8 to 20 the required slot length varies from approximately half to about a quarter wavelength.

Simultaneously with the proper Q, a desired resonance frequency also has to be obtained. This is controlled by the shape of the slots. By changing the width of the slots either at the center or at the ends the relationship between the resonance frequency and the Q can be varied (see Fig. 20).

To provide for some fine tuning of frequency, either inductive change at the ends of the slots or capacitive loading in the middle of the slots can be used (see Figs. 20 and 21).

High-Power Handling Capacity

With the trend towards increases in allowable and desired erp, an antenna should have not only adequate power handling capacity, but preferably more than needed to allow for future increases of erp. For mechanical reasons, the transmission line in the traveling-wave antenna has a very high power handling capacity. Since there are no feed lines, the maximum power is limited by voltage gradients in the components, the capacitors, the slots, and the input section. All of these have very little effect on the other characteristics of the antenna and therefore have been designed for any practical power handling capacity needed in the near future.

Summary

These outstanding electrical characteristics, excellent vertical and horizontal patterns, and pattern bandwidths, combined with high power handling capacity and low vswr, are provided in the clean simple design of the traveling-wave antenna.

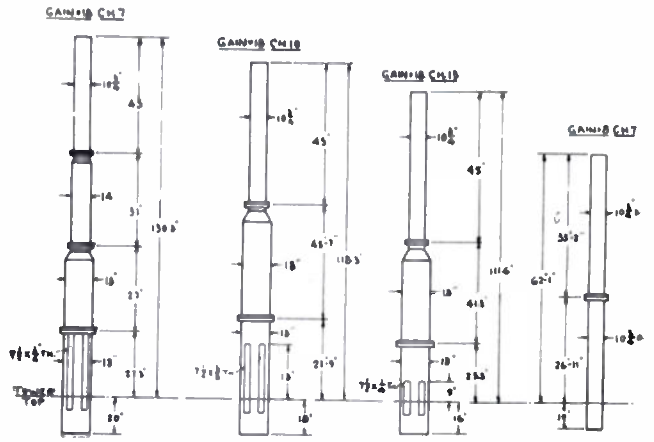

Fig. 22—Traveling wave antenna.

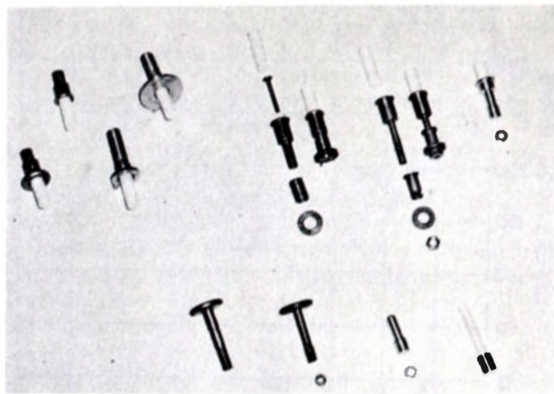

Fig. 24.

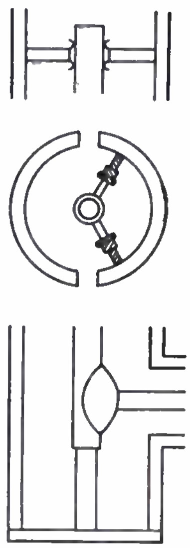

Fig. 23.

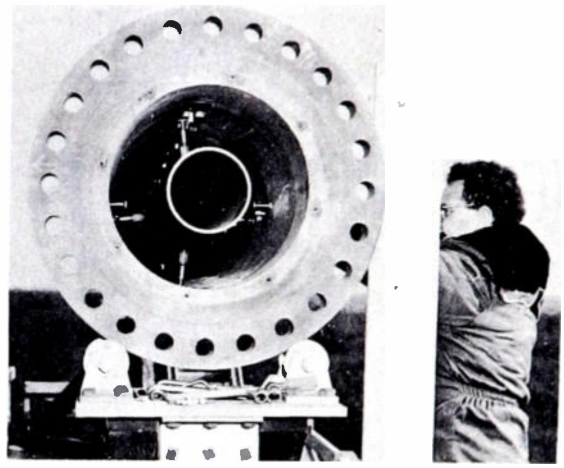

Fig. 25.

Mechanical Features

Wind load is an important consideration because it greatly affects the cost of the supporting structure. The smooth cylindrical shape of the antenna is ideal in this respect.

The steel outer shell of the antenna is made of rolled steel pipe with weld-neck flanges welded to each end for coupling to its mate. Three different sizes of steel pipe are needed to cover all gains at vhf high band (see Fig. 22). The antennas are designed to withstand wind pressures of 50 pounds per square foot on flat or 33 psf on cylindrical surfaces. They are rated for a wind velocity of 110 miles per hour.

The steel outer conductor, after fabrication, is hot dip galvanized for both better conductivity and weather protection.

The antenna has been designed for tower mounting. A guide flange at the tower top guides the pole into the tower during erection and prevents the pole from swaying. The pole socket carries the weight of the antenna.

The inner conductor of the antenna is rigidly supported at the bottom end without having to rely on any insulator type of support to carry the dead weight (see Fig. 23). The inner conductor can be removed from the bottom if required. The coupling capacitor probes, besides being used for electrical coupling, also provide centering support for the inner conductor (see Figs. 23, 24, and 25). This is accomplished by insulator pins, which are built into the probes and form a portion of the dielectric of the coupling capacitors. The short at the top of the antenna has a sliding contact to allow for any thermal differential expansion between the inner conductor and the pole (see Figs. 21 and 26).

The power is fed into the the antenna through the tee at the bottom end of the antenna. To provide a broad-band low-reflection input, the bottom end of the pole forms a quarter-wavelength shorted stub and the inner conductor on each side is equipped with compensating transformers (see Fig. 27). In addition, great care has been taken to maintain proper characteristic impedance in the tee joint itself ("ball" in Fig. 27). A gas stop is mounted right alongside the pole to facilitate pressurization of the transmission line (see Figs. 27 and 28). The antenna itself is not pressurized.

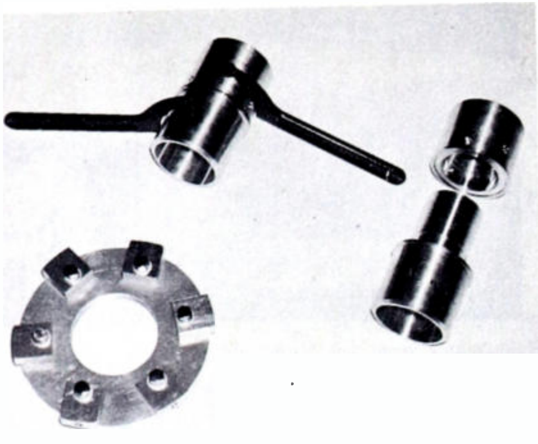

Fig. 26.

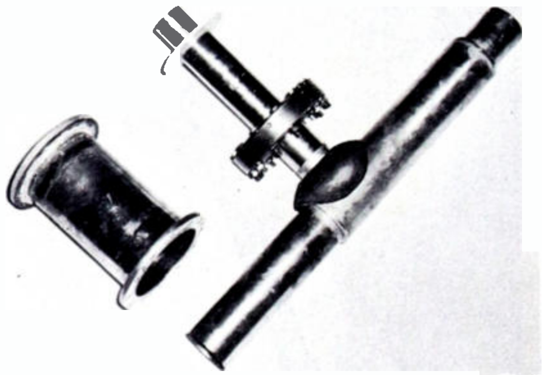

Fig. 27.

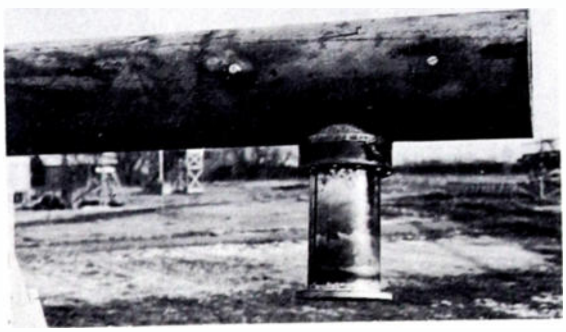

Fig. 28.

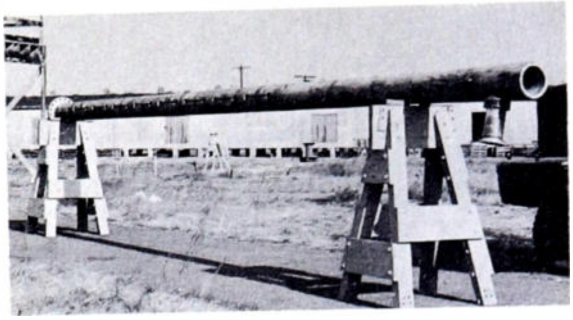

Fig. 29.

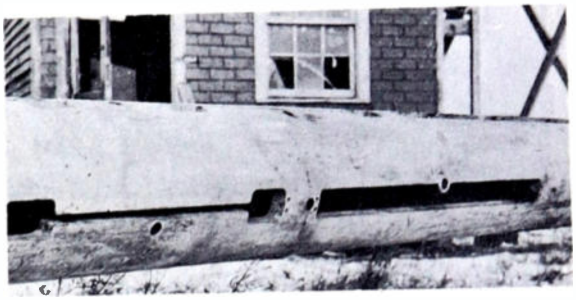

Fig. 30.

As an example Fig. 29 shows the bottom half of the pole for a gain of 8 channel-7 antenna. For this low-gain, long low-Q slots are needed (see Fig. 21). The plates at the ends of the slots are used for fine frequency tuning during development. To achieve proper operation of the top loading still lower Q slots are used (see Fig. 30 right). Additionally, the coupling networks in the top loading are different from those in the main aperture.

Many television transmitting antennas, such as the superturnstile antennas, can be measured and developed at a relatively low elevation, since the radiating systems can be separated so that reflections from the ground do not cause excessive errors. However, this is not possible with the traveling-wave antenna. Thus, the development of the antennas has been performed at about 25 feet elevation (see Fig. 31). The important measurements in the development of a traveling-wave antenna are the phase propagation and attenuation function measurements instead of the input impedance measurements as is normal.

The patterns of the traveling-wave antennas are measured at the Gibbsboro, N. J., test site (see Fig. 32). The antennas are rotated on turntables to measure the vertical patterns and on the rollers at the top of the turntables to obtain the horizontal patterns.

Like all broadcast antennas exposed to the elements, great care must be exercised to design for wind, snow, ice, lightning, electrolysis, effects of the sun, and sudden changes in temperature.

As mentioned previously, the steel outer conductors are hot dip galvanized. Slot covers are fastened to the pole over each slot. They are made of special polethylene that is not affected by ultraviolet rays from the sun nor embrittled the cold. To take care of condensation inside the antenna, provisions are made for drainage.

The materials used for component items are selected to insure no corrosion effects from electrolysis due to dissimilar metals or chemical action due to industrial gases. All aluminum components are anodized.

Since slot radiators are used, it is highly improbable that lightning will damage the antenna. However, to protect the 300-mm code beacon at the top of the antenna, provision is made for lightning by installation of a lightning protector which also supports the beacon. The grounding of the inner conductor at each end of the antenna provides further protection for the transmission line and transmitter.

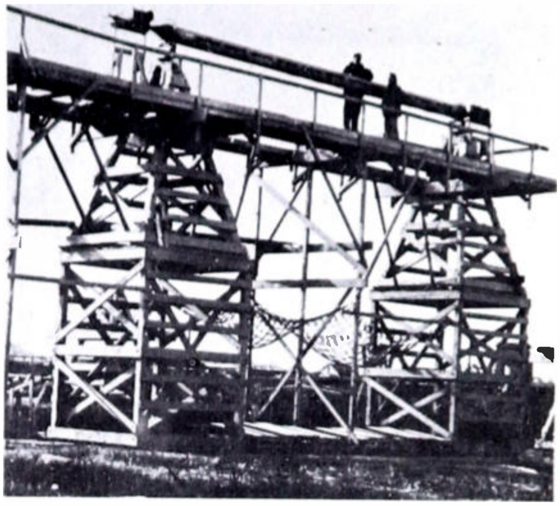

Fig. 31.

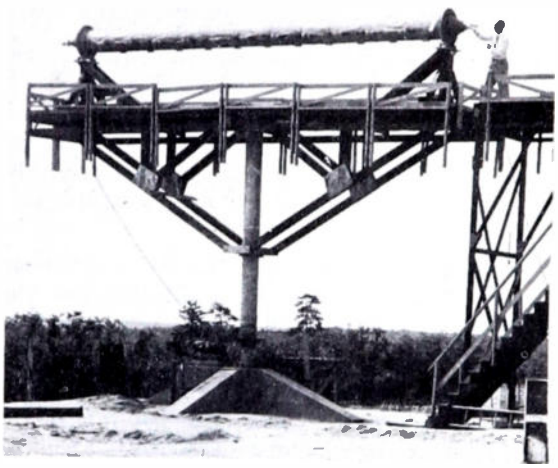

Fig. 32.

If ice is collected on the antenna it lowers the resonant frequency of the slots, thus affecting the performance of the antenna, mainly at aural carrier. Therefore, in areas of heavy icing, sleet melters (optimal item) are recommended to avoid excessive accumulation of ihe. Fig. 33 shows icing tests.

Conclusion

The traveling-wave antenna[1] provides an answer for the need of a vhf high-band antenna which combines outstanding electrical characteristics with mechanical simplicity and economy (Fig. 34 shows a gain of 8 channel-7 antenna.)

[1] This antenna is based on fundamental theory developed at the Ohio State Univ. Res. Foundation under contract between the RCA Victor Div. and the Foundation. See W. Masters and C. J. Rauch, "A new television transmitting antenna," 1955 IRE Convention Record, part 7, pp. 28-29.

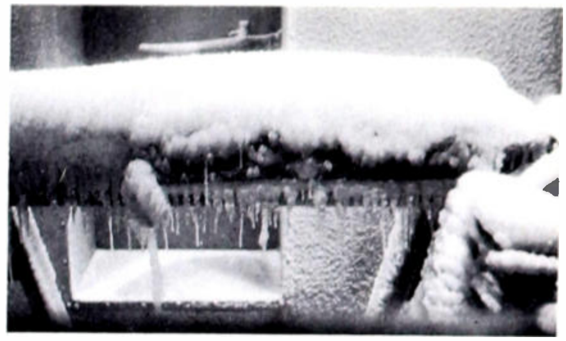

Fig. 33.

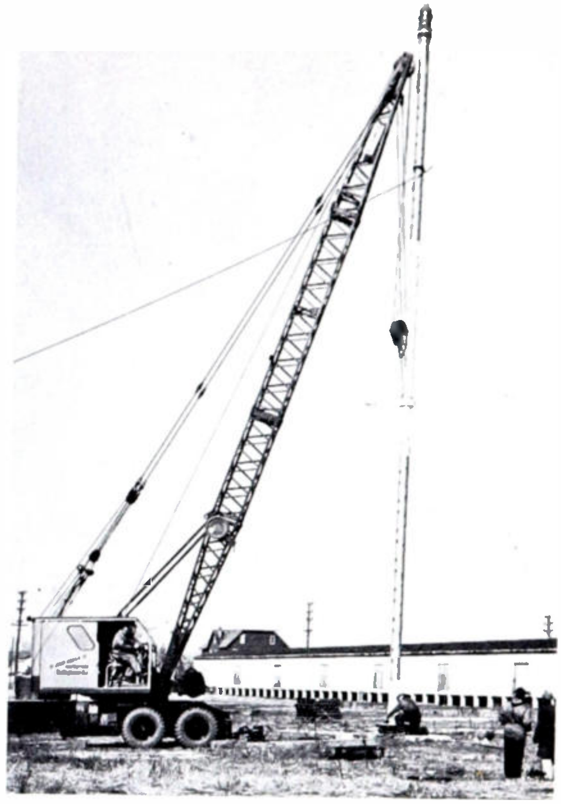

Fig. 34.

The improved characteristics, most of them inherent to the principle, derive from the unique combination of supporting structure, transmission line, and radiators. The horizontal pattern is formed on the turnstile principle from the excellent patterns of single slot pairs even at large pole diameters. The smooth vertical pattern, which helps to achieve an even field strength within the service area, is based on the exponentially decaying traveling-wave feed. The proper electrical characteristics of the simple slot radiators maintain constant wavelength instead of velocity in the transmission line for frequencies in the channel. This assures proper bandwidth. The low input vswr is due to the traveling-wave nature of feed in the antenna. The over-all construction of the traveling-wave antenna is clean, simple, and rugged. Simplicity of the antenna and low wind load result in economy of both the antenna and the supporting structure.

VIDEO TAPE RECORDER

SYMPOSIUM

The first prototype models of the AMPEX video tape recorder have been used for several months in delaying television programs for later release to network stations on the West Coast. The operational experiences and the solutions to the many problems inherent with putting new equipment into field use were discussed by technical representatives from networks using the recorders and AMPEX.

Speakers: ROSS SYNDER, Ampex Corporation, Redwood City, Calif.

CHARLES GINSBURG, Video Engineering Dept., Ampex Corporation, Reedwood City, Calif.

HELMER ANDERSEN, Columbia Broadcasting System, Los Angeles, Calif.

OSCAR WICK, National Broadcasting Company, Hollywood, Calif.

ROBERT VON BEHREN, Minnesota Mining Manufacturing Company, St. Paul, Minn.

UNDERSTANDING ART PROBLEMS IN COLOR TELECASTING *

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This text explains some of the major problems faced by the artist when he works with color television. He uses subtractive color and should understand how the engineer produces color additively. Besides using different terms, the artist and engineer have other problems, such as chromatic induction and gray scale equivalents. They must coordinate their efforts in order to produce good color, as well as a good black and white picture, and stay within the twenty-to-one light reflection limitations.

To solve these new problems brought on by color TV and to better understand color in general, a series of clinics was conducted at KRON-TV. This led to the development of Chroma-CHron, a palette which has been a great aid in assisting scenic and graphic artists, has led to a better understanding between the artist and engineer, and has resulted in the understanding of the gray scale equivalent, which in turn assures compatible color art.

The Cathode Ray Oscilloscope

When it comes to understanding color, the one thing perhaps that causes more confusion between artist and engineer is the Cathode Ray Oscilloscope (CRO).

It is difficult for the artist to understand why the engineer is hypnotized by the CRO during color telecasting. The engineer states that the CRO is a device for portraying an instantaneous graphical presentation of electrical signals which represent picture information and is used for the evaluation of a color signal.

This explanation does not help the artist. But if he were to take an hour off and look over the shoulder of the technician and also watch the CRO, the artist would note that the CRO contained three graphs - each one representing one of the primaries of the TV system. Besides some information which is of little interest to the artist, each graph displays the amount of each of the primary colors in the scene.

When only one graph displays picture information, it means that the color before the camera is a fully saturated primary. When two of the graphs display picture information, it means that the color before the camera is a mixture of two primaries or a binary.

Any fully saturated primary - or mixture of two primaries - is what the engineer considers "hue." If the CRO registers a color as a monochromatic, or as a binary, he has a fully saturated color. However, the addition of two colors, as displayed by the graphs, does not correspond to the result that the artist would have if he were to mix these colors with pigment.

The CRO registers yellow, chartreuse and orange as a mixture of red and green. The artist could not get yellow, chartreuse or orange in this manner; a mixture of green and red pigment would result in "cool mud," "warm mud," or "neutral mud."

This is one of the reasons why it is difficult for the artist to understand CRO. The information it gives is that of color as mixed additively. The artist mixes color subtractively - or the exact opposite as shown by the CRO.

Additive processes for color reproduction are based on the use of three primaries - red, blue, and green. When these three colors overlap in light, white appears. When red and green overlap, magenta appears. When blue and green overlap, cyan appears. Any two together make a compliment of the third.

This last sentence also holds true for subtractive primaries (yellow, magenta, and cyan) and this is a rare instance where the additive and the subtractive have the same results. Also, the primaries of one system are the binaries of the other.

In pigment, when the three additive primaries are intermixed, the result is mud. When red and green are intermixed, the result is mud. When red and blue are intermixed, the result is a muddy violet. When blue and green are intermixed, the result is a toned blue-green.

With these three additive colors as primaries, the artist could never produce white, yellow, magenta, or cyan. He could not paint in a realistic manner. And most of the colors which he could produce with these three (red, blue, green) as his primaries would be very "toned" colors. Actually, he does not have to know additive color - although he would find it interesting and helpful.

By understanding additive color, he could cope with the engineer who often complains that the artist does not use saturated colors. The engineer shows the CRO to the artist as proof that the colors used in scenery and/or graphic art are not fully saturated.

The engineer must be aware of the fact that it is next to impossible to consider any color fully saturated due to the impurities of pigment. The CRO will also show contamination depending upon the angle of the light on the color, and/or the amount of light on the color. If a color is over-illuminated, then it is over-exposed and the additional light dilutes the color. Diluted color activates all graphs in the CRO. Even if a pigment could be found that is a fully saturated primary, the mode of lighting could make it appear desaturated or contaminated by the two remaining primaries.

Nevertheless, the engineer wants nice, clean, saturated color because the system is capable of producing such color. He wants to have the system reproduce the color as well as possible.

If a color is very saturated, the engineer is happy. But saturated colors in a scene clash among other saturated colors. They fight for importance and over-power other colors which may be toned. Actually, an artist paints in harmonies, which often require subtle color contrast, and thus he avoids raw saturated color.

This thinking, of course, is the exact opposite to the engineer who wants to produce pure color - without realizing that he may be destroying the harmony or mood which the artist is trying to achieve. But in most instances, the engineer is not after pure color. Instead, he is after luminosity - which is often the opposite to saturation. Since his language is not the same as that of the artist, there is a lack of understanding which also leads to confusion.

A fully saturated blue is composed of only eleven percent of the light that is necessary to produce white. This small percentage of light automatically makes it a low luminous color. In order to bring up the luminosity of the blue, it must be mixed with the other two primaries; or, in reality, it must be contaminated by the two remaining primaries. This results in a luminous blue. It is not saturated; it is desaturated. But if the blue does not have "sparkle," the engineer is apt to say that it lacks saturation, not that it lacks dilution.

The engineer achieves luminosity by using all three signals; the artist merely dilutes his color with white pigment. By diluting his colors, the artist is assisting in illuminating the scene, since dark and/or vivid colors absorb a greater percent of wave lengths of light.

By using the white pigment to dilute his color pigments, the artist is causing all three graphs in the CRO to activate. This does not produce saturated color, but it does produce luminous color. Now all graphs are jumping and the engineer is as happy as the artist. To further understand each other, the following terms can be used to advantage by the engineer and the artist:

Hue is pure color, although it will not register pure in the CRO. Hue is the purest color obtainable in pigment or the basic, vivid colors which the artist chooses for his palette.

Saturation is the degree of hue in mixed color. De-saturation is described in the following terms:

- Tint - Hue diluted with white (in pigment).
- Tone - Hue mixed with gray (black & white).
- Shade - Hue darkened with black.

Brightness refers to the luminosity of hue, tint, tone, or shade. Brightness or luminosity of a color or color mixtures corresponds to its value or to its position in the gray scale.

In common terms, tint is pale color, tone is grayed color, hue is vivid color, and shade is deep color. The brightness of these mixtures is their gray scale equivalent. (Knowing the brightness of a color is the most important thing for the artist to learn.)

Chromatic Induction

Once the artist and the engineer understand each other, the next problem is to understand "What is Color?". Basically speaking, color is the property of light. Pigment, that which gives color to objects, does not contain color, since color is produced by light. There is no color in darkness.

Because of this, color cannot be seen, cannot be felt, nor can it be identified without light. Pigment has the characteristic of absorbing certain wave lengths of light and reflecting others. The reflected wave lengths are responsible for the experience of color.

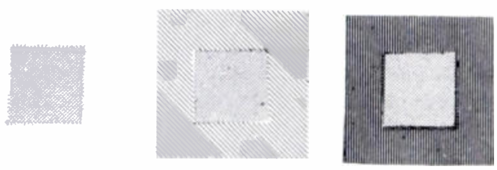
The artist, as well as everyone else, just sees color on surfaces without much care or thought as to how this is experienced. But just seeing color can have its complications, for a color is only that particular color depending upon its environment. The light sources illuminating the color, the angle of light on the surface, the area surrounding the color, and the length of time which the human eye views the color have an effect on the color.

For instance, a certain orange is only that particular orange depending upon the colors which surround it. If the colors surrounding it are darker than the orange, then the orange is lighter. If the surround is lighter, the orange is deeper. If the surround is warm, the orange is toned. The orange as viewed under incandescent light, flourescent light, or daylight is different. And if the person looks at that orange color too long, the eyes become weary and cause it to lose its purity.

The surround can induce brightness and/or grayness or chroma; or it can induce darkness and/or chroma. That is, depending on the surround, a color or gray appears to change in value and/or to have a foreign color mixture.

Not knowing what a color actually is (because of its environment) can be confusing to the artist. By simple tests he can prove to himself that any color is darker when surrounded by light and lighter when surrounded by dark. If he knew what the gray equivalent to that color was, its gray would also appear light or dark, depending upon the surround, for a gray, just like a color, is only that gray depending upon the surrounding colors, or other surrounding grays.

If a medium gray is surrounded by darker values, then it appears lighter; if the same medium gray is surrounded by lighter values, then it appears darker. This is noticeable in the three squares on this page. Only one value of gray is being surrounded. However, the values of the surrounded gray appear lighter or darker depending on the surround.



The value, as experienced by the eye, of any gray is dependent on its environment. If surrounded by color, the gray appears to contain the complimentary hue of the surround. The term "chromatic induction" is used to explain these color experiences.

Chromatic induction can make a color appear lighter or darker, which in turn makes its gray equivalent do the same. Because of this the artist can never be certain just what is the gray equivalent of a color when it can appear to be different under various circumstances. Although he may not be completely aware of it, his main concern is understanding "What is a gray?".

There is a consolation in knowing that the gray equivalent of a color is physically the same regardless of surround. Two adjacent colors will have both color and gray scale contrast if their gray equivalents are apart in the gray scale. The only way that these two grays can appear the same is by isolating them and giving them separate surrounds.

Chromatic induction, as well as "value surround," can be used to advantage by the artist.

ChromaCHron

Clinics were conducted by KRON-TV in order to introduce color to advertising agencies and to artists. These resulted in the adoption of the color terms used earlier in this text. These clinics and experiments also led to the development of ChromaCHron, which is a palette designed to assist scenic and graphic artists working in color TV.

Although color lithography and photography can be achieved by use of the three subtractive primaries, the artist needs a more versatile palette. Even three primaries or four chromatic "primitive sensations" plus black and white are not enough.

A palette consists of the necessary basic colors with which the artist can produce his art work. The ChromaCHron palette allows the artist maximum freedom in regard to color moods and harmonies. Like all palettes, ChromaCHron limits the number of basic paints needed and reduces the number of variations of these colors for the sake of simplification. These limitations add to its effectiveness and do not restrict the artist in color selections or in creativeness. These are beneficial limitations in the respect that they make ChromaCHron easy to use and to understand. The number of paints and formulas for intermixing these paints are few, making ChromaCHron a practical and economical approach to painting for television, whether it be for color or for black and white.

Although most artists are capable, few have had any training in art for television, let alone color television, and are at a loss when it comes to understanding what the twenty-to-one light ratio limits are. Very few, if any, are equipped with the tools by which to measure light reflection in order to determine the difference between the most reflective and the most light-absorbing colors. There is no human way of predicting the gray equivalent of any color other than by the results of past experience.

A piece of art containing five grays can appear to have enough contrast, depending on how the grays are arranged in the picture. It can appear to be within the system's limitations, or appear to exceed the system's limitations. Only a camera preview can give him the answer. The preview is possible only after the art work has been produced. If the art work is wrong, then it must be re-done.

Since surround can induce color and value changes, it adds to the complication of knowing just what is the gray equivalent of any color. Its TV gray equivalent would also be different - and even more difficult to understand. Because of this, most artists at some time produce color art with adequate color contrast but lacking in brightness contrast. It televises well in color, but results in an unusable black and white picture. Correcting this type of unpredictable error can be costly and embarrassing.

Since art departments work on limited budgets, errors must be avoided, for re-doing any art is costly in man hours and materials. The paints selected are versatile and inexpensive. Paints are subject to waste, and a large selection of paints adds to the waste. The palette colors must intermix in order to produce an acceptable "color wheel" of adequate hues.

The ChromaCHron "wheel" is the palette, and each of these wheel colors has tint, tone, and shade variations. All of the color mixtures in the ChromaCHron master palette have a gray scale equivalent. By knowing the physical gray scale of each color, the artist is assured of contrast. The light reflection also is determined in order that, besides having a proper gray scale contrast, the picture is not exceeding any limitations and is staying within brackets of safety. This gives the artist great assurance that his black and white picture will be as successful as his color picture.

ChromaCHron does not teach the artist how to paint or tell the artist what colors to use in his painting. Color choice is a personal thing which must be left entirely to the artist. ChromaCHron cannot tell him if his color selection is in good or poor taste. It will only tell him that the colors will telecast as they were selected and if their gray scale equivalents have proper contrast. When the artist begins his mental process of selecting colors, it is possible that the result will be completely acceptable for color as well as the black and white picture. But by checking the ChromaCHron charts, it is verified that his original color choice will work.

If he had selected one or more colors which have color contrast, but are of equal brightness, the black and white picture then would be incompatible. This would not mean that he must completely change his selection of color, but it would mean that he must make minor changes in the saturation of one or more of the colors. The color mood or harmony would not necessarily be affected by the corrections, and these changes would meet the requirements of the engineering departments.

ChromaCHron differs from other palettes, systems, and theories in the respect that it has been designed specifically for television. Colors were chosen because they are readily available and have been manufactured for a number of years. These are inexpensive paints which may be used for both scenery and graphic art. If special colors were to be manufactured for ChromaCHron, the cost would make it impractical. A compromise was made on the selection of paints. Due to the limited number of parent or basic colors, there is very little waste in this paint. Both color and black and white originating with color and/or black and white equipment are aided by ChromaCHron. Stations not telecasting in color may use it to improve the gray scale in their present art work. It also gives them a library of color art which will have proper color selections when color equipment is installed.

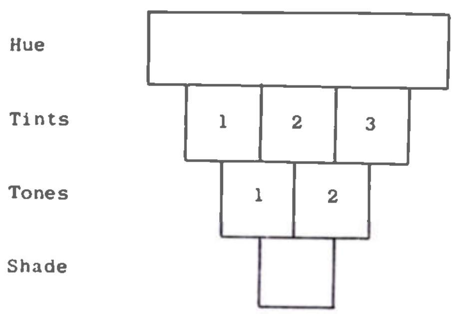
ChromaCHron requires seven water soluble paints; five are chromatic and two are achromatic. These paints are suitable for both scenic and graphic art. The chromatics are neither primary colors nor primitive sensations. They are five basic colors which intermix to produce an adequate palette of thirteen common hues.

There are thirteen hues in the ChromaCHron palette. In all diagrams they are placed so that they form a circle with yellow on top. Moving clockwise they are called yellow, light orange, deep orange, bright red, deep red, wine red, purple, blue, blue green, forest green, green, light green, and chartreuse.

The thirteen hues are achieved in the following manner. Mixtures of yellow and bright red produce light orange and deep orange. Mixtures of deep red and blue produce wine red and purple. Mixtures of blue and green produce blue-green and forest green. Mixtures of yellow and green produce light green and chartreuse.
Formulas:

8 yellow plus 1 bright red = light orange
8 yellow plus 7 bright red = deep orange
3 deep red plus 1 blue = wine red
1 deep red plus 1 blue = purple
5 blue plus 2 green = bluegreen
5 blue plus 6 green = forest green
6 yellow plus 4 green = light green
16 yellow plus 1 green = chartreuse

In ChromaCHron each hue has three tints, two tones, and one shade. This gives each color seven variations, or there is a total of ninety-one color mixtures in the palette. Each color appears in the master palette as shown in the diagram:



Formulas:

Shade 15 hue plus 1 black
Tone 1 20 hue plus 15 white plus 1 black
Tone 2 10 hue plus 5 white plus 1 black
Tint 1 1 hue plus 5 white
Tint 2 1 hue plus 1 white
Tint 3 5 hue plus 1 white

The most important thing in ChromaCHron is knowing the gray equivalent of a color. The hues, tints, tones, and shades as arranged in the palette have little, if any, meaning in regard to painting for television. Along with the ChromaCHron palette, a gray scale is essential.

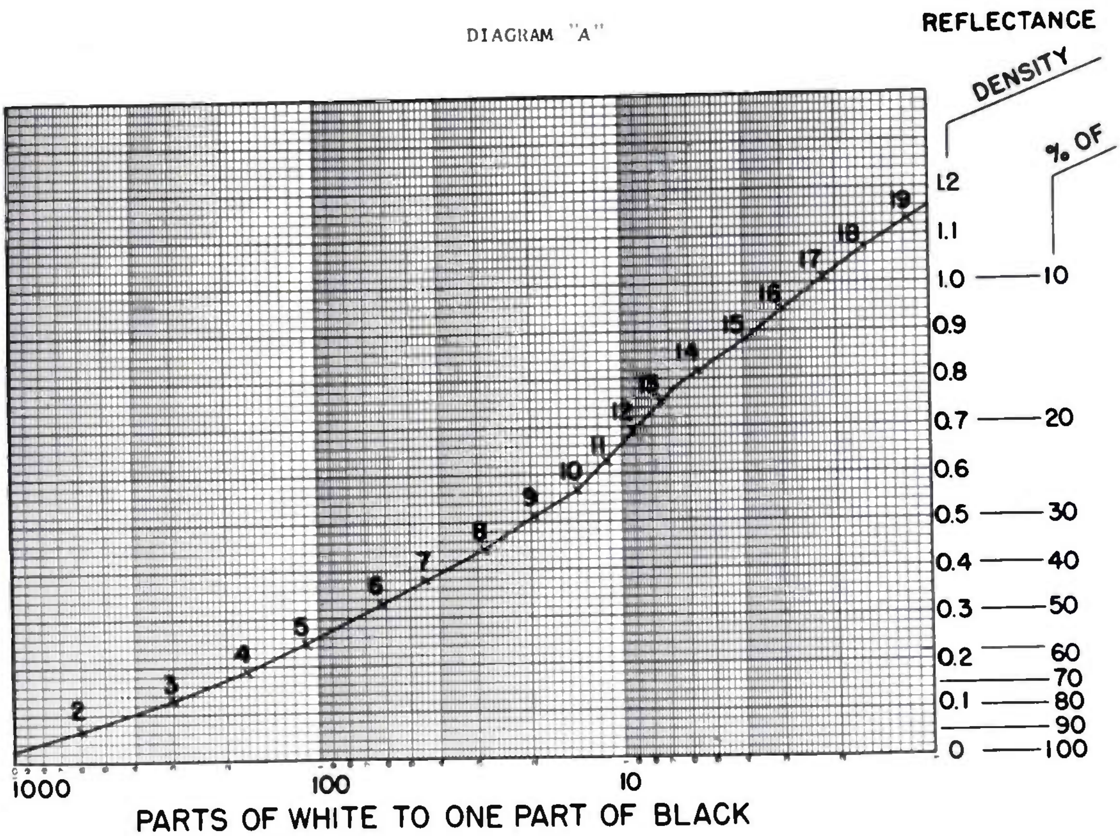
ChromaCHron's selection of gray scale is based upon the measurement of reflectance of ChromaCHron's white and black and the division of this range into nineteen equal parts in terms of reflectance density, thus achieving twenty steps including white and black. White is step number one and black is step number twenty.

The ChromaChron paints are in a paste form and must be diluted with water. In order to assure uniformity in the amount of pigment, a test is made and checked with previous results. In the case of the grays, if the test gray was darker than previous tests it meant that the black was not properly diluted and more water was added to the black pigment. Once the paints are balanced, the artist can proceed in making his color mixtures with the assurance that his tints, tones, shades, and grays will not vary to any great degree. There also will be some mixtures which will be slightly lighter or darker than the anticipated value; however, the shift in value will be insignificant.

Due to the importance of the gray scale, special precautions are taken to assure its accuracy. Color mixtures can be accurately matched by eye with the ChromaCHron palette as a guide. Diagram "A" lists the reflectance density of each gray in the scale and the formulas by which to achieve each of the twenty steps.

After the ChromaCHron gray scale was established, all colors were illuminated with 300 foot candles, 3200 Kelvin, at a 45 degree angle. They were viewed through a color camera on a black and white monitor and also on a Tektronix oscilloscope. The color camera was balanced against a known value of gray in order to obtain a minimum chrominance modulation out of the color plexer. Amplitude of colors of approximate

DIAGRAM "A"



value was compared to the value of gray used in the setup. First, the lightest or those colors with the most brightness were compared to ChromaCHron white (gray #1). There was a rebalancing after each step of gray. This produced minimum modulation for each gray value.

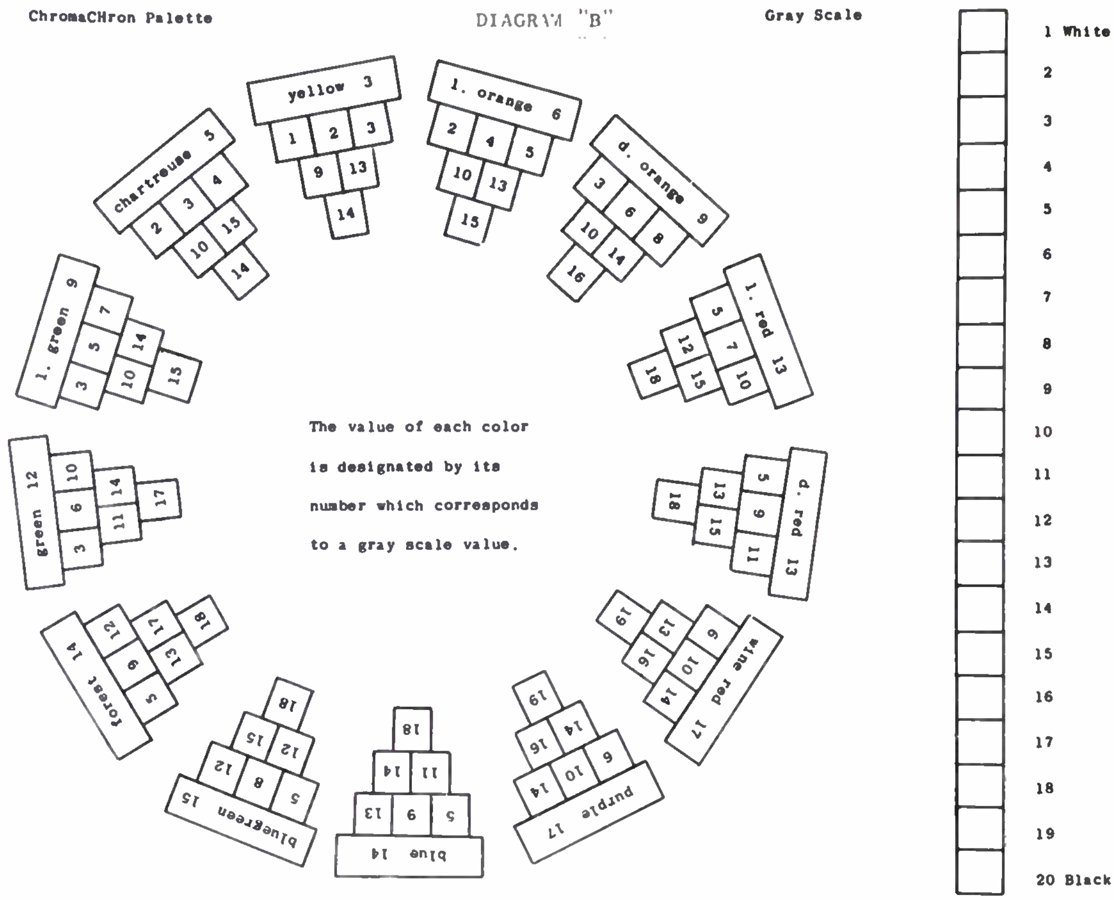
Every color in the ChromaCHron palette was tested and its gray equivalent listed. Diagram "B" contains the complete ChromaCHron palette, along with its gray scale. The number within each color chip in the palette corresponds to the number of the value in the gray scale. Merely by looking at this palette in full color the artist becomes aware of the approximate gray scale equivalent of each color. He can use the ChromaCHron palette as a guide for determining the grays of colors which may not be incorporated in this palette.

All graphic ChromaCHron testing was done on illustration board. Although all paints were tested on the same type of card, some cards with slight difference in surface reflectance were intermixed for the sake of comparison. The opacity of the paint and its texture completely overruled the reflectance surface of the card. In the case where the paints were tested on scenery canvas, the slight texture of the material was not enough to effect or alter the gray equivalent. However, this does not mean that these paints will always render the same gray regardless of the surface they cover. Had a coarser canvas been used the gray scale would have shifted evenly in any one direction. ChromaCHron is more successful when used on surfaces of similar reflection. That is, if card is going to be incorporated with scenery, the reflectance of canvas and of the card must be similar.

Under flat lighting conditions it is impossible to exceed the twenty-to-one light ratio limitations with the use of ChromaCHron paints. The brightest and darkest pigments or any mixtures of these pigments are well within the brackets of safety. It would not be possible to produce art work using any color combination on standard illustration board or on canvas and exceed the system's limitations. ChromaCHron under uniform lighting and painted on uniform textured surfaces will not exceed the twenty-to-one ratio.

This does not mean that the twenty-to-one ratio cannot be exceeded with the use of ChromaCHron. The mode of lighting or variety of lighting in one scene can alter the ChromaCHron

DIAGRAM "B"



gray scale. When the angle of light varies, the luminosity of the color varies. If ChromaCHron paints are used on surfaces such as combed plywood whose characteristics of light reflection cannot be altered by a coat of paint, then the ChromaCHron gray rendition is not accurate.

Incorporating other things with ChromaCHron also can effect the twenty-to-one ratio. A specular light source such as the reflection from jewelry, glass, chrome, foil, etc., can throw ChromaCHron off and make it exceed the system's limitations.

In general, to maintain a twenty-to-one relationship one should determine the lighting contrast on the set and divide this contrast figure into twenty to determine how many steps of gray value can be used safely. When persons of average complexion are incorporated with ChromaCHron, it is advisable to stay in the middle range (5 to 15) of the gray scale.

ChromaCHron must be used wisely. It is effective when used properly. ChromaCHron is easy to use.

This concludes the pigment phase of ChromaCHron but by no means does it stop here. Other phases will be incorporated, such as the testing of the various colored cards and papers presently used by graphic artists. ChromaCHron will catalog the gray scale of those card materials which are the most practical for color television graphic art work. At the present a great deal of graphic art is converted into color transparencies. The ChromaCHron palette will be converted into color transparencies to determine the gray scale after it has been through the photographic process. The most popular color films available will be used in this experiment and processing will be done under ideal conditions. KRON-TV is ready to co-operate with manufacturers of paints, draperies, paper manufacturers, and others who produce graphic and scenic materials for the purpose of giving the ChromaCHron gray scale equivalent of their product.

A COMPATIBLE SINGLE-SIDEBAND SYSTEM DESIGNED FOR USE IN THE BROADCAST SERVICE

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Introduction

A new technique has been developed to convert standard AM broadcast transmitters to single-sideband operation. Conversion is achieved by use of a special adapter which can be installed without any modifications of the transmitter and can be received on all existing AM receivers.

The system, known as Compatible Single-Sideband, or CSSB, is quite similar to standard AM transmission, except that the spectrum energy is concentrated on only one side of the carrier. Because of this important and basic difference, the broadcast industry is now able to obtain the following significant advantages over normal AM, which result in considerable improvement in day and night coverage of primary and secondary areas:

1. Provides an effective two-to-one power gain for a given signal fidelity.

2. Reduction of adjacent and co-channel interference through two-to-one spectrum economy.

3. Minimized selective fading distortion.

4. Improved audio fidelity.

CSSB has now been in continuous use for approximately one year on the Voice of America's megawatt station in Munich, Germany. The first domestic tests were recently completed on WMGM's 50 kw transmitter in New York, and arrangements have just been made for full-time tests over WABC, the American Broadcasting Company's New York outlet, and over WSM's clear channel station in Nashville.

Compatible Single-Sideband

We would like to propose a new system that does not have any inherent distortion at a 100 per cent modulation. This system, Compatible Single-Sideband (CSSB), allows the use of conventional AM receivers and is actually slightly less sensitive to incorrect tuning than AM transmission. It may be best to examine a simplified block diagram of a Compatible Single-Sideband transmitter at this point. Figure 1.

In the Compatible Single-Sideband system, a full carrier single-sideband wave is produced in what may be a conventional single-sideband generator. This wave is then passed through a limiter wherein the phase modulation component of the wave is isolated. This phase modulated wave is than amplified by Class-C amplifiers, or for that matter, any other class of amplifiers and finally fed to a modulated stage.

The full carrier single-sideband wave, at the output of the SSB generator, is also fed to a product demodulator wherein it is electronically multiplied by the carrier. The resulting output wave from the product demodulator is free of distortion. In other words, the spectrum components at the output of this demodulator are identical with those at the input of the single-sideband generator. This audio wave is then amplified and finally used to modulate the phase modulation component. If this modulation process is linear, the envelope wave at the output of the modulated amplifier is free of all harmonic distortion. This wave may then be demodulated in the conventional diode detector with theoretical zero distortion at 100 per cent modulation.

Analysis and measurements show that the undesired sideband of this wave is approximately 30 db below the desired sideband. Even though the Compatible Single-Sideband wave looks exactly like an AM wave on an oscilloscope, the wave is single-sidebanded.

Full Carrier SSB vs CSSB

Application of single-sideband techniques to AM broadcasting has been, heretofore, considered impractical, since its adoption would involve modification of home receivers. If conventional home receivers are used to receive a single-sideband full carrier signal, appreciable distortion results. This may be seen by referring to Figure 2, which shows a full carrier SSB wave modulated by a single tone under peak modulation conditions. It is seen that the envelope waveshape is far from sinusoidal and actually a little over 23 per cent harmonic distortion is present.

Also, it should be noted that the fundamental term has a peak value of 67 per cent relative to the d-c term. Therefore, the maximum effective modulation of a full carrier single-sideband wave is only 67 per cent. It has been suggested that in order to reduce the large amount of distortion of this envelope, the sideband level should be reduced. It may be shown that though the distortion is reduced, the effective modulation is also greatly attenuated. When the distortion is reduced to 10 per cent, which is normally considered to be barely tolerable for most equipment, though not quite acceptable for broadcast service, the effective modulation is 38 per cent. Such a low level modulation value reduces the effective power of the transmitter by a factor of over 5.3.

Two proposals[4,5] have been made for reducing this distortion effect at the lower frequencies by transmitting double-sideband for low audio frequencies and single-sideband for higher frequencies. However, these systems require more spectrum and are more critical to selective fading than pure single-sideband systems. These systems also require the replacement of most existing transmitters.

Advantages

The advantages of the Compatible Single-Sideband system are as follows:

1. Reduction of Adjacent and Co-Channel Interference

Since Compatible Single-Sideband systems concentrate energy in one side-band, CSSB transmission reduces adjacent and co-channel interference.

a. Adjacent Channel Interference. In the case of adjacent channel interference, the use of Compatible Single-Sideband can increase the spacing of adjacent sidebands by two times the highest audio frequency transmitted. This increased effective sideband spacing greatly reduces adjacent channel interference.

Adjacent channel interference comprises three main types; sideband monkey chatter, undesired crosstalk, and carrier heterodynes. In the United States, the 10 kc separation between carriers tends to reduce this form of interference because of IF and audio fidelity, as well as the listener's aural limitations. The crosstalk effect is greatly reduced by masking, and except for extremely large undesired signal levels, is relatively unimportant. The main source of adjacent channel interference is sideband monkey chatter wherein the desired carrier beats with the undesired sideband components. If CSSB techniques are correctly applied, the frequency of the monkey chatter would be sufficiently removed from the desired carrier to be above audibility. Thus, CSSB offers a means for greatly reducing adjacent channel interference.

b. Co-Channel Interference. If co-channel stations are equipped for CSSB operation, the listener, by tuning to the desired station's side of the carrier, can effectively reduce interference effects. The optimum tuning point appears to be 1 1/2 to 2 kc on desired sideband side of the carrier, and conventional home receivers offer a signal-to-interference gain of from 5 to 8 db. If a listener, in a particularly poor region, purchased a special high selectivity receiver, he could obtain approximately 30 db signal-to-co-channel interference gain.

In addition to the reduction of this type of interference, there is also a reduction in beatnote distortion, which in AM systems is caused by the "phase beating" of interfering carriers. This reduction in beatnote distortion is discussed in the next section on "Selective Fading Distortion Reduction".

2. Selective Fading Distortion Reduction

The second advantage of Compatible Single-Sideband operation is the reduction of selective fading distortion, offering an increase in night-time coverage. It may be shown that the main cause of fading distortion is incorrect relative phasing of the carrier and the sidebands. This condition may be demonstrated by eliminating the carrier from an amplitude modulated wave and then reinserting the carrier at different phase relationships. It is then seen that when the carrier differs by 90° from its correct phase, the signal, demodulated in an AM detector, is completely distorted. This distortion is independent of the percentage of modulation.

Compatible Single-Sideband operation, by suppressing one of the sidebands, is much less sensitive to selective fading distortion. The relative phase of the components of the Compatible Single-Sideband wave is much less critical, and tests have established the fact that Compatible Single-Sideband waves are relatively free of fading distortion.

It should be pointed out that this insensitivity to phase deviation is another reason for the reduction in co-channel interference. When the undesired signal has a carrier frequency approximately equal to that of the desired carrier frequency, the combined carriers will be phase modulated at a low frequency beat note rate. This wave will then go in and out of proper phase, and there will be a form of beating distortion. Because Compatible Single-Sideband is less sensitive to phase discrepancies, CSSB is relatively free of this distortion.

3. Improved Fidelity

Because the Compatible Single-Sideband wave occupies one-half of the normal AM spectrum, the bandwidth of the IF and RF amplifiers of the receiver may be halved. Since the fidelity of most receivers is restricted by the IF bandwidth, Compatible Single-Sideband offers a means for appreciably improving the effective fidelity of existing receivers.

If desired, the future home receivers may be built for restricted fidelity service, and in this case, it will be possible to obtain a signal-to-noise improvement by the use of CSSB. For a given fidelity, the bandwidth may be halved, thus providing a 3 db signal-to-noise improvement. Such receivers could be made somewhat cheaper because more gain can be obtained in the narrower IF amplifiers than in wider units. Also, it will be easier to obtain improved selectivity with attendant reduction in co-channel and adjacent interference effects.

Discussion of Compatible Single-Sideband Tests

The first installation of Compatible Single-Sideband took place in Munich, Germany with the Voice of America megawatt transmitter wherein a 4 megawatt peak envelope power CSSB wave was produced. Since installation, this equipment has been used continuously by the Voice of America.

During the WMGM Compatible Single-Sideband tests in New York, numerous measurements and listening tests were made on various types of receivers. These included communications models such as the Hammarlund SP-600, the Collins 51-J, a Signal Corps R390A/URR, and the Hallicrafter 538-D. Home-type receivers such as a Grundig Majestic Model 80U/USA, a Westinghouse transistor set, a Craftsman tuner, and many other types were used. No difficulty was experienced in tuning these receivers, and reception was completely free of distortion. When listening to CSSB signals on a communications-type receiver, it was possible to reduce the receiver's bandwidth to 2 kc and still maintain usable quality.

Tests were also made at the WMGM transmitting site at Rutherford, New Jersey, and on automobile radios at considerable distances from New York. It was noted that even though at times the signal faded out completely, it remained free of distortion throughout the fading cycle.

Sideband rejection of CSSB transmissions under best conditions was slightly over 35 db, as read on a Panoramic adapter Model SB-86. This figure was confirmed by separate measurements on Collins, Crosby, and Kahn single-sideband receiving systems. However, average sideband rejection was 30 db. The amount of distortion added to the transmitter system by installing a CSSB adapter was only .4 per cent, and the total transmitted bandwidth was approximately 7 kc.

The CSSB adapter was run continuously without adjustment for two months, and the undesired sideband at the end of this period measured 30 db down from the desired sideband. Figure 3 shows an actual photograph of the CSSB wave viewed on a Panoramic Panalyzor after two months' operation without adjustment. Figure 4 shows the wave form immediately after switching back to conventional double-sideband AM transmission. When the equipment was adjusted at the beginning of the test, the spurious was down 32 db.

Conclusion

It has been shown in actual operation that CSSB systems offer the user significant reductions in interference and selective fading, as well as improvements in signal-to-noise and audio fidelity. The adapter is easy to install and maintain and does not require delicate or repeated adjustments. Because of its unique design and compatibility with existing AM receivers, the many advantages of single-sideband transmission may now be fully realized for the first time by the broadcast industry.

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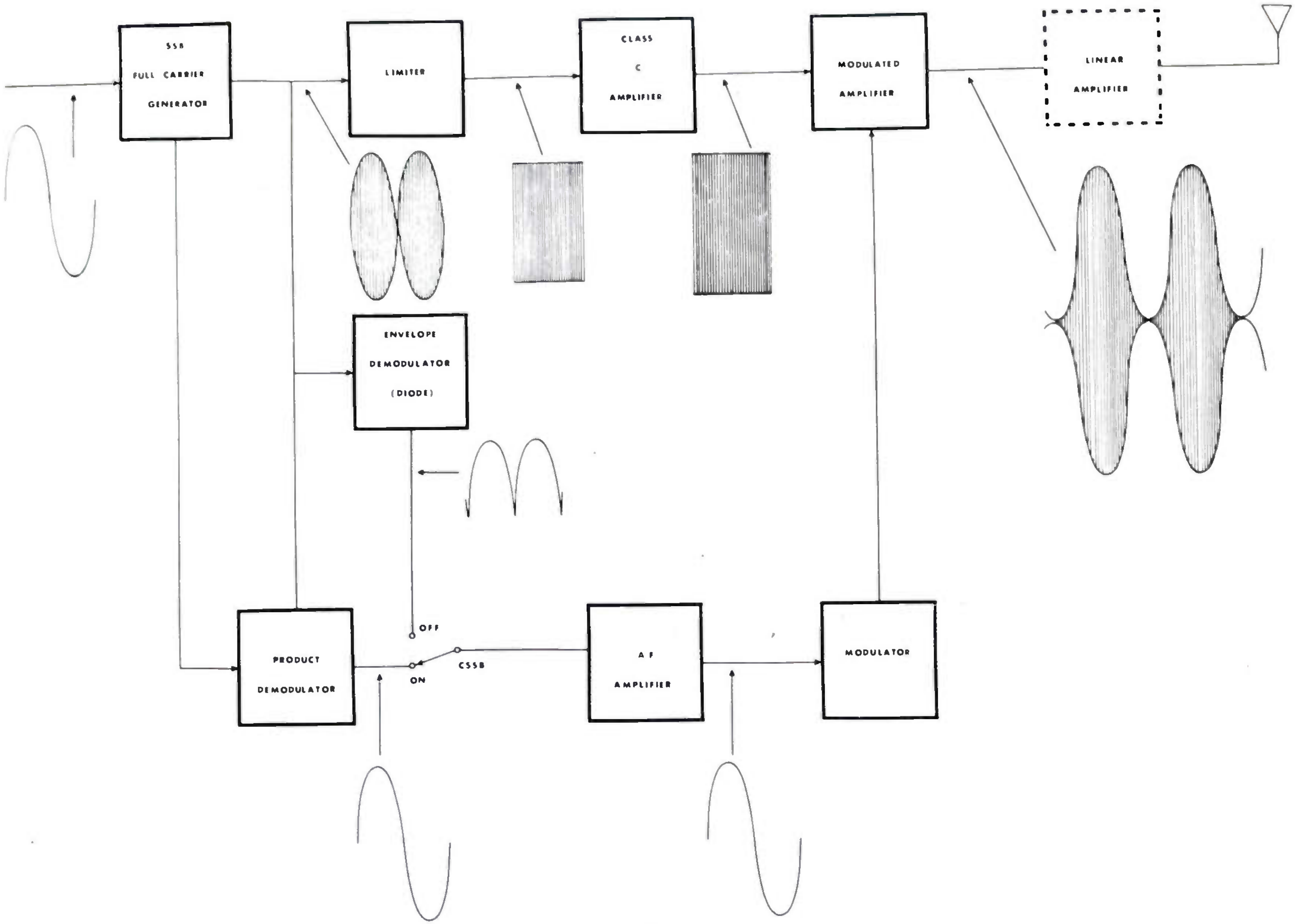
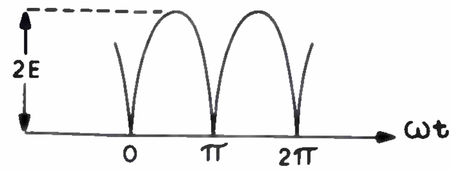


FIGURE 1

BLOCK DIAGRAM OF BASIC COMPATIBLE SSB SYSTEM



$$e = \frac{4}{\pi} E \left(1 + \frac{2}{3}\cos \omega t - \frac{2}{15}\cos 2\omega t + \frac{2}{35}\cos 3\omega t \ldots\right)$$

67% equivalent modulation approximately 23% distortion

For 10% distortion equivalent modulation 37.8%

FIGURE 2

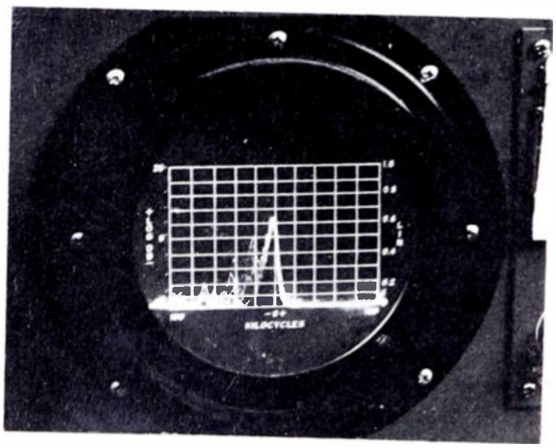


Fig. 3

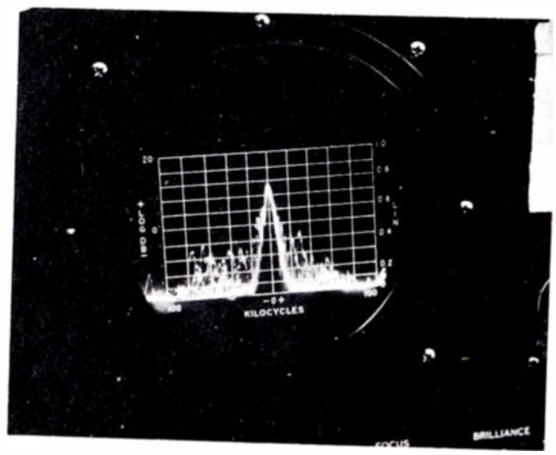


Fig. 4

A STABLE PRECISION TELEVISION DEMODULATOR

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Summary

Proposals are advanced that would enable the design of a more reliable and precise television demodulator. The present state of the art is reviewed with comments on short-comings that need to be corrected. Design features and experience gained from an experimental prototype using product detection and unusually low frequency IF amplification is presented. The possibilities of direct frequency conversion to video without the use of an IF amplifier, relying upon post detection filtering for band shaping is discussed. Proposals initiating industry wide standards for the performance of television demodulators is opened for discussion.

The Problem

In order to evaluate the operating characteristics of a television transmitter it is necessary to provide a true demodulated sample of the signal delivered to the antenna suitable for signal analysis with a precision waveform monitor or other TV testing device at a standard level and impedance. The term DEMODULATOR as referred to in this discussion is intended to mean the complete instrument used to detect and process the radio frequency TV signal in such a manner that a video replica of the original modulation is available for study.

So that the measurements being conducted on the transmission system are to be of any value, it is essential that the demodulation process be very precise and complementary to the established processes of modulation and sideband filtering so the output of the instrument will truly reflect the deviations of the transmitted signal from the norm. The demodulator must not include any observable defects that are not attributable to the transmitted signal itself.

Every TV broadcaster requires the use of such an instrument to check compliance with the FCC regulations, as well as to confirm the best possible signal is being obtained from the equipment on hand. It is customary for the station maintenance engineers to run amplitude response curves of the entire TV system, observing the results on the demodulator waveform monitor to compare with sideband analyzer measurements. The same is true for measurements for color systems, such as differential phase and amplitude response. Interpretation of the correct amount of sub-carrier to video (monochrome) ratio can only be determined through a demodulator because of the vagaries associated with such observations when checking samples obtained from diode probes at various points of the system output, due to the vestigial sideband nature of the radiated signal.

Hence, it is necessary that the demodulator closely adhere to the standards of the ideal receiver. It must have an accurate and stable amplitude response curve approaching the theoretical ideal, phase response that is known to be that expected of the ideal receiver as set forth in the NTSC specifications, and an overall transfer characteristic that is essentially flat.

Contemporary Design

The customary approach to demodulator design has been the employment of an elaboration of ordinary television receiver circuitry. Usually, no high gain RF preamplification is required since available signal voltages are usually on the order of several hundred millivolts and it is general practice to feed this directly to the 1st conversion mixer. Usual intermediate frequency amplification ranges are in the 20 to 30 megacycle region where typical stagger-tuned bandpass circuits have reasonable component values, considering the percentage bandwidth required to accommodate the full TV signal. Finally, a high level envelope detector is used to recover the video.

Of presently available equipment, the published tolerances for a demodulator sold by one major manufacturer is indicated to have deviations in the amplitude response curve no greater than plus 0.6 DB or minus 1.0 DB relative to that at 0.2 MC. This corresponds to a possible dip of 11% and a peak of 7.5%, or an overall excursion of some 18%.

The product of another major manufacturer lists a number of specifications regarding overshoot and ringing in regard to transient response and gives an amplitude tolerance of within 1 DB. Detailed specifications regarding the performance following color conversion of the unit were not available to the author at the time of this report, however, it appeared that the principal changes involved were the addition of a series of all pass phase compensating networks No major redesign of the IF amplifier was in evidence that would indicate improvement in the published specifications of amplitude except for the inclusion of a better sound trap.

It is certain that very careful design in every respect yields results that are in proportion to the attention paid to details. However, even the best units commercially available leave something to be desired. Foremost, the detector inherently has consideracle waveform distortion at high modulation percentages,[1, 2] principally due to quadrature distortion brought about by the partial suppression of the lower sideband in the vestigial sideband system. This defect manifests itself as increased rise time compared with that of the in phase component of the modulation envelope, and unequal disposition of the high frequency components around low frequency transitions. The rise times of signals of opposite polarity are not equal nor symetrical. The rise time is dependent upon the modulation percentage. The observed waveform therefore has defects that are not attributable to shortcomings in the transmission system, but due to the detector itself.

Another serious shortcoming in contemporary demodulator design is the lack of stability of the IF amplifier system. The major portion of the blame could be placed upon the questionable necessity of using a high frequency portion of the spectrum for the bandpass amplifier. Due to the limitations of the envelope detector, the practical limits have been reached in reducing the frequency range while still covering the entire video bandwidth without introducing complications in the design of the diode detector load.

New Approach

It is not the intent of this paper to unveil a panacea that will circumvent all the troubles resulting from previous attempts to build an ideal demodulator but rather to give a progress report, discussing a number of successful expedients that have been tried as well as to discuss the lines of future endeavour that appear most promising and delineate some needed standards of performance.

In a search for a partial solution for these various shortcomings, the author's attention focused upon the possibilities of the product detector.[1, 3, 4]

The basic components of a product type demodulator having the required features is outlined in block diagram form in Fig. 1. The well known process[3] involves the multiplication of the incoming modulated signal

$$e = F(t) \cos(w_K t + \phi_K)$$

by a carrier frequency of the appropriate phase to recover the original modulation component modified by the exclusion of information lost due to the suppression of a substantial portion of one sideband.

e = instantaneous voltage, $F(t)$ is a modulating component, $f_k = w_k/2$ = carrier frequency and ϕ_k = phase of carrier at time $t = t_o$.

When the above expression is multiplied by the instantaneous value of carrier, the product is

$$i = \gamma e \cos(w_K t + \phi_K) = \gamma F(t) \cos(w_K t + \phi_K)^2$$

$$= \frac{\gamma F(t)}{2} + \frac{\gamma \cos 2(w_K t + \phi_K)}{2}$$

where i = the output current and γ = a constant.

If integrated over a period of time exceeding one cycle, the second term reduces to zero, so

$$I_o = \frac{F(t)}{2}$$

yielding the original modulating component.

The output of the detector is filtered with a low pass configuration in order to remove the undesired modulation components lying about a frequency at the second harmonic of the carrier, as well as remove the fundamental carrier frequency and sidebands from the demodulated output.

Previous writers[1, 4, 5] have elaborated on the many features of the product detector that are suitable for TV work and it's place in the world has been amply demonstrated as the accepted method of separating the two components of the color signal. As applied to the use as a TV demodulator, several features make the use of this type of circuit a distinct possibility as an approach to a more ideal system. The product detector is immune to quadrature distortion caused by unequal amplitude distribution about the modulated carrier frequency.

It has been shown[1, 5, 6, 9] that provided the signal has a uniform phase curve symetrical about the carrier, the filtered demodulated output of the detector will be an accurate replica of the original modulation, vestigial sideband transmission not withstanding. This is of course due to the ability of the gated detector to remain blind to the quadrature component contributed by the single sideband portion of the spectrum. Since filtering of the signal does introduce phase nonlinearities, it is necessary to correct for them. This can be performed upon the detected output with typical all pass non minimum phase type filters.

The selectivity of the circuit can be completely specified by the response assigned to the post detection circuitry, if desired. This follows from the low pass analogy of band pass circuits when using product detection.[7] This means that should it be desired to suppress a portion of the input signal so it does not appear in the final detected output of the device, the selectivity may be employed after detection by video filtering rather than predetection circuits operating on the RF signal. This is not practical with envelope detectors where the intermodulation products of both desired and undesired signals would appear in the output when no predetection selectivity is employed.

This feature has much to offer, especially the ease with which sound carrier interference can be eliminated from the picture with high Q video trap circuits in lieu of IF traps.

Another feature offered by the product detector is the unique way in which it lends itself to demodulating a signal that is in a frequency range adjacent to the spectrum of the demodulated output of the detector. Due to time constant problems in the detector load impedance, conventional envelope detectors impose design headaches when an attempt is made to recover the modulation of a signal, the carrier frequency of which is nearly as low as the upper spectrum of the demodulated video. In mixer type synchronous demodulators, the signal circuit is not common with the output load impedance, hence were it not for other considerations of convenience, the two spectras could even overlap and using signal balanced mixers would still be practical to recover the modulation.

It was the author's thought that if by some means the IF frequency could be drastically reduced, it would be possible to shape the pass band curve more exactly and one would have some assurance the curve would remain as initially tuned.

An especially intriguing idea that was explored is the possibility of making the IF range coincide with the video spectrum. In other words, use direct synchronous detection at the incoming carrier frequency in lieu of converting first to an IF frequency band for filtering, thus eliminating the IF system entirely. Since the synchronous detector lends itself to post detection filtering, this seemed to be the most elegant way to create a precision demodulator. The output at video frequencies corresponding to double sideband output of the transmitted signal would be 6 db higher than the higher frequencies, so a step type filter would be interposed in the output video. (See Fig 2) A difficulty arises in a rather exasperating manner. Since the lower sideband attenuation curve of the transmitted signal is quite steep, signal energy in the spectrum approximately 3/4 megacycle from the carrier is subject to extensive ringing such as is associated with all sharp cutoff filters. The 6 db step in the proposed step response filter aggravates the situation, principally due to an extremely awkward phase delay curve in the vicinity of one megacycle. (See Fig 3) The correction of the video envelope delay imposed by this curve would demand the use of a very elaborate phase correcting network due to the large correction needed at both ends of the spectrum to match the relative delay at midband.

However, work is continuing in the effort to bridge these problems in a straightforward manner so as to open the possibility of the use of direct synchronous detection having precision characteristics. Experimental work completed thus far has yielded promising results along these lines.

By using an IF frequency that is immediately adjacent to the video spectrum, a much lower practical frequency for IF amplification is achieved. This allows the filtering operations to take place in the near video range of approximately 5 to 12 megacycles. (See Fig 4) The undesirable IF components can then simply be filtered out with a low pass filter in the detector output. This filter could be a part of the sound trap circuitry since this function can also be done in the video system following detection. The IF response curve needed is the typical skewed slope with the carrier placed at the minus 6 db point. For given component tolerances, the stability and precision with which the ideal curves can be obtained at the lower IF frequency is improved by a magnitude of approximately three to ten.

Upon first thought, it might seem that the building of a bandpass amplifier having a bandwidth nearly equal to the center frequency would pose a very difficult design problem. The use of conventional stagger tuned or over coupled tuned stages would be impractical. The approach used in this case was simply to use a wideband video amplifier, essentially flat from a few cycles up to approximately 12 megacycles, and to cascade low pass and high pass filters to obtain the desired intermediate frequency passband. One advantage gained is that each stage of the amplifier is responsible for flat amplification of the entire passband, thus the plate loads are more constant over the entire range and changes in tube characteristics are not as prone to change the shape of the passband.

Another convenience of the low IF frequency is the ability to observe the waveform of the modulation envelope itself on a wideband oscilloscope before detection. Falling well within the usable bandwidth of most good scope amplifiers, it is quite simple to compare the modulation envelope waveforms at various points in the IF amplifier both before and following the skew symetrical filters and to compare these waveforms with those obtained after detection. Thus detector deficiencies can be pinpointed. Shown in Fig. 5 is the observed envelope response to a particular TV transmitter, prior to entering the shaping filter. Fig. 6 shows the response after passing through the filter. Fig. 7 shows the detected inphase component as it appears following the product detector, prior to delay correction. Fig. 7b shows a sync pulse detected and phase corrected for IF distortion. IF and sound components removed by filtering.

The spike on the edge of sync is clearly caused by the response of the shaping filter and is typical of the delay error introduced by the suppression of the portion of the signal near carrier. The video output of the phase correction filter is compensated to correct for this.

There are two possible methods of conversion that will produce an IF band in essentially the same spectrum but each requiring slightly different treatment as far as filtering. The first method involves the choosing of a local oscillator frequency approximately 10 MC above the visual carrier frequency. For an exact difference of 10.25 MC, the IF passband will be from 5.75 to 11.0 MC with a filter requirement that is of the low pass variety to hold the curve to close tolerances between 9.5 and 11.0 MC with the carrier at the 50% point on the curve as shown in Fig. 4a. The band of frequencies is transposed relative to the carrier as transmitted.

An alternative method of mixing to obtain the low IF frequency is illustrated in Fig. 4b. A local carrier approximately 6.0 MC lower in frequency is beaten against the incoming signal to translate the modualtion into the region of 4.5 to 10.5 MC. Since the sideband spectrum is not transposed, the filter required would be of the high pass variety, with accurate control from 5.25 MC to 6.75 MC and constant gain from that point up to the region of the sound carrier just short of 10.5 MC.

Except for possible reasons of filter design, there is no clear reason to prefer one method over the other. It might be possible to obtain sharper edges at the transition points in the gain curve for given component losses by using the nontransposed conversion since the critical shaping is done at half the frequency of the counterpart shown in Fig. 4a.

The present development of the demodulator utilizes the first method described in the preceeding paragraphs; the conversion oscillator frequency is higher than the incoming visual carrier and the IF spectrum is transposed as shown in Fig. 4a.

A block diagram is given in Fig. 8. Typical of contemporary counterparts, the RF signal is fed into a mixer to convert to the IF frequency. In the specific case, satisfactory behavior of the converter is obtained with a type 6BN6 gated beam tube, used as an outer grid mixer. The output is quite linear as long as the input signal does not exceed about 1/2 volt. Approximately 5 volts of oscillator voltage is presented to the quadrature grid from an oscillator operating 10.25 MC above the visual carrier frequency.

The frequency of this oscillator is controlled by a reactance tube included in the AFC loop. The reference used is the oscillator which provides the demodulating carrier in the product detector. Higher overall stability is achieved by controlling the first oscillator than to move the final oscillator, since the lower frequency oscillator may be accurately positioned at the center of the slope of the IF curve. The error signal is obtained by comparing the fixed frequency of the final demodulating oscillator with the incoming carrier frequency after the conversion to 10.25 MC. The amount it lacks of being precisely on 10.25 MC and in phase with the reference, is observed in a phase detector and this error voltage in turn controls the reactance tube.

The IF amplifier consists of a succession of pentodecathode follower combinations with filter sections between pairs. Each pair is adjusted by means of shunt plate peaking coils to remain flat over the entire range to beyond 12 MC. The filter elements control only the shape of the suppressed portions of each edge of the passband, thus each tube combination amplifies the whole passband equally. However wasteful of gain due to the high gain-bandwidth product required, this is a stable and easily adjusted method to achieve the wide bandwidth at such a low frequency.

The only critical shaping operation performed in the IF stages is that of accurately setting the response so as to be a straight line slope, skew symetrical about the carrier. This is done with a bridged T filter and a series resonant shunt circuit between two stage pairs and an antiresonant interstage coupling circuit between two other pairs. The low frequency portion is broadly suppressed with one high pass interstage coupling. No attempt was made at this point to introduce a sound trap.

The output of the IF amplifier feeds the product detector, which again utilizes another gated beam tube, operating similarly to the first mixer except that in this case the local oscillator frequency matches the IF video carrier frequency. The output of the detector is filtered to remove the greater portion of the 10.25 MC carrier and the remaining video is amplified in a conventional video amplifier. The sound trap and phase correcting filters operate in the output of this amplifier at a 75 ohm level.

Since the color corrected transmitter is assumed to fully correct for the receiver sound trap phase discontinuity when minimum phase shift circuits are used, no correction for receiver cutoff is provided. Only the correction for the slope of the skew symetrical filtering is accounted for.

This filtering causes the low frequency video to lag the high frequency portions by approximately 0.14 microseconds. A bridge T phase equalizer circuit was designed to effect this correction. The nature of the correction desired is illustrated in the measured square wave response of the network pictured in Fig. 11.

The sound trap is intentionally left out of the IF amplifier for the reasons given earlier, as well as to allow the use of a direct 4.5 MC sound takeoff in the output video amplifier. The purpose of this innovation was to experiment with the feasibility of utilizing this signal as a carrier alarm actuating circuit and sound monitor.

Alignment of the bandpass amplifier is quite simple. The output of a standard sweep generator with variable marker capable of sweepingfrom 0 - 12 MC is fed into the RF input of the first mixer with the hetrodyne oscillator disabled. The signal is sampled at the grid of the second mixer. The demodulating hetrodyne oscillator is allowed to run, which through slight coupling in the detector stage will put a marker pip at 10.25 MC. The shaping filter elements between the stages of the amplifier are adjusted until the curve of Fig. 9 is obtained. By this method, the first mixer passes the IF sweep through its normal plate load. The CW output of the marker, used alone may be used to obtain phase delay measurements. Phase comparisons between input and output were made using a Hewlett Packard model 150-A switched pre-amp scope which allows simultaneous viewing of both input and output waveforms superimposed so as to measure the time difference.

The video amplifiers following the detector are swept in the usual way and at this time the sound trap and low pass video filter elements are set for a cutoff above 4.2 MC with a deep notch at 4.5 MC. It is desirable to adjust the filter for minimum ripple in the passband, with the least objectionable effects on the phase plot below cutoff.

Any of the conventional means of chopping the signal by mechanical or electrical methods may be used to obtain the DC reference axis.

Elimination of IF Amplifier

It might be interesting to review the method used to test the system outlined in a previous paragraph where the IF amplifier was dispensed with and direct conversion from RF to video was performed.

The incoming RF signal with visual carrier at 61.25 MC was sampled with a directional coupler at the output of the transmitter vestigial sideband filter. The reference carrier was obtained by sampling the unmodulated output of the driver amplifier to the modulated stage of the video transmitter with a pickup loop inserted near the driver plate tank. The phase was matched to that of the carrier of the modulated sample by means of an adjustable lumped circuit line stretcher. The in phase condition was simply checked by adjusting for maximum low frequency video output, which could be confirmed by metering the plate current of the mixer tube.

This carrier source was satisfactory enough to enable testing of the basic system but would prove impractical for general use. The principal drawback is the interaction of transmitter tuning and demodulator characteristics due to changing phase of the radiated carrier with respect to the sampled carrier in the unmodulated stage, depending upon the tuning of the bandpass circuits. Even though the response curve also specifies the phase of the transmitted components due to the minimum phase shift law[7, 8] there is no such constraint upon absolute delay, hence the relative delay(therefore phase) of the modulated sample could not be depended upon to remain in fixed phase reference to the unmodulated sample.

Later work was done with a phase locked locally generated carrier, controlled by a conventional AFC system. The demodulator output was then passed through a step filter of the ladder type, the design of which was simulated by the corner plot technique. The amplitude and delay response is shown in Fig. 2 and 3. Delay is plotted as group delay and was obtained by point phase measurements at closely speced increments throughout the band of interest and determinations of the incremental slope of each segment. The curve includes all errors inherent in the video amplifier as well, which in this case were of little consequence throughout the band of interest.

The type of response obtained from vestigial sideband transmission through the step filter is shown in Fib. 10 which clearly shows the ring in the 1 mc region that proved so exasperating.

It might be noted by the reader that the waveform overshoot from the output of the direct conversion demodulator is similar in type but opposite in sense to the inherent spike produced by IF amplifier distortion shown in Fig. 7. Since the two systems cannot be used in combination, there is no possibility that one might correct for the other.

Other Considerations of Direct Conversion Method

Aside from the ringing problems imposed by the step transitions in the response curves of the transmitter and receiver post-detection filters as described earlier, another consideration must be taken into account.

It will be noticed that due to lack of any predetection selectivity giving additional sideband suppression, the direct conversion detector views a greater total band of frequencies than does the conventional detector following a shaped IF passband. The receiver IF off center slope is required to be skew symetrical and to cause the response below minus 0.75 to be negligible. The specification of the remainder of the vestagial sideband below minus 0.75 MC is not detailed in the regulations covering the transmitted signal except that the suppression must be virtually complete at the lower band edge. Since the conventional receiving system is blind to energy radiated below minus 0.75 from carrier, the exact value of components below minus 0.75 MC is ordinarily of no great concern.

In the case of a detector employing post detection filtering only, this portion of the spectrum does become of some concern, since it contributes to the demodulated video output. Lacking detailed specifications of this response, let us consider the effects of possible deviations from an arbitrarily assumed straight line curve from full amplitude at minus 0.75 MC to complete suppression at minus 1.0 MC. Energy from each sideband is addative and to this sum is added the loss in DB of the complimentary filter. The result is that the total error of the complete combination is directly proportional to the deviation in DB from the straight line O-A in Fig. 12. Practical necessity requires that both lines become rounded at the corners.

Since the regulations do cover the situation at two specific frequencies, one half MC apart, this is tantamount to implying a restrictive curve in interval between.

From a practical standpoint there is little else that could happen due to the limited response possibilities that can be achieved with practical sideband filters in such a short frequency interval. This is true regardless of the method of obtaining the required response, whether by the use of high level passive filtering at the transmitter output, or with interstage couplings in low level modulated systems. It can be safely assumed that the response in this interval will be virtually according to the curve of Fig. 12 in all normal installations.

Standards Proposals

It is perhaps time that some uniform standard be set for the performance of demodulators, whatever method is employed. This statement is apt to raise many proposals on just what these specifications should entail. The approach should be that of an industry wide committee to work out an appropriate standard. In any event, the author would like to mention several desirable inclusions that should be a part of the detailed performance tolerances. To readily qualify as a precision instrument, it shouldn't be too much to expect that it should retain an amplitude tolerance of no greater than plus or minus 2 1/2 percent, as illustrated in Fig. 13. This could give rise to a curve of the type illustrated that just falls within the tolerances. A deviation any greater than that shown could hardly qualify as a useful measuring device.

This should not be too difficult a tolerance for a manufacturer to maintain in a production model. The requirements for phase delay might well be open for extended discussion, however it seems clear that if the entire system measurements are to mean anything, the demodulator should not be expected to contribute more than possibly 0.02 microsecond of error to the total. This deviation could represent 20% of the total of a measurement on a typical system just meeting FCC tolerances. Although posing a formidable production quality control problem to the manufacturer, it is certain to be a necessary requirement imposed by the television broadcast engineer as a criteria of the accuracy of his measuring equipment.

It is also proposed to identify the tolerances of the overall video response curve with certain requirements of the IF slope near the region of carrier. This would amount to a tightening up of the deviations near the 0.2 MC reference so that excursions near the maximum of 2 1/2% could not occur immediately adjacent to the 200 kc region. Reduced to practice it would mean a decreasing tolerance to zero in the vicinity of carrier and would also imply that an increase on one side of carrier must be nearly balanced by a like decrease on the other side of carrier, thus maintaining symmetry of the slope.

A tolerance referred to 200 kc must not allow any deviation at that frequency, otherwise it no longer has a reference and merely allows adroit maneuvering of the curve to fit almost any adjusted filter combination so that it appears to be within specifications. So the tolerance limits should reduce to zero near the carrier frequency in the manner shown in Fig. 13.

Some agreement should also be reached regarding permissible factors affecting color measurements. Differential phase and gain curves should be set comparable with the standards of the networks and the telephone company for their individual pieces of equipment such that any one unit adds little to the total of a number connected in tandem. Similarly, the demodulator is in tandem with a number of other pieces of gear and should contribute insignificantly to the total differential errors. A phase margin of no greater than one degree and a transfer linearity within one percent should be the goal.

In concluding, it should be remarked that none of the working models with which the experiments were conducted represent manufacturing prototypes by any stretch of the imagination. The work was undertaken and tried idea by idea through the use of demountable plate experimental chassis wherein each stage was built as a separate entity on a small metal plate and held in place on a special assembly such that a succession of circuits could be constructed to operate in normal fashion. The ease with which new ideas could be tested proved indispensible, however usually the only remains of earlier trials were a few sordid parts tacked upon many small metal plates scattered about the workbench. This is a little demoralizing to associates who seem to sense that progress just might be going backwards as plate after plate is discarded.

It is hoped that the ideas presented here may prove rewarding to others who are bent upon the same road and that the ultimate outcome will eventually prove to be a "Stable Precision Television Demodulator".

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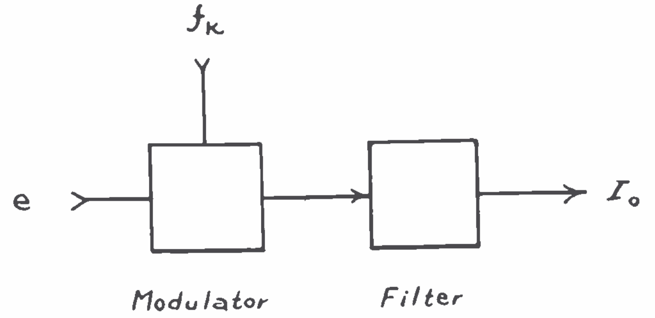
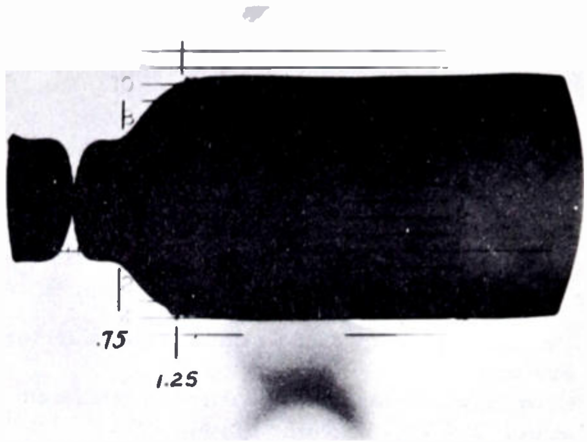


Fig. 1. Block diagram of product detector.



Step Filter Response
Fig. 2

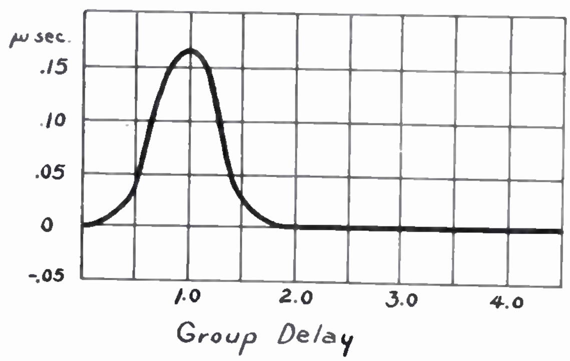


Fig. 3. Group delay of filter having response of Fig. 2.

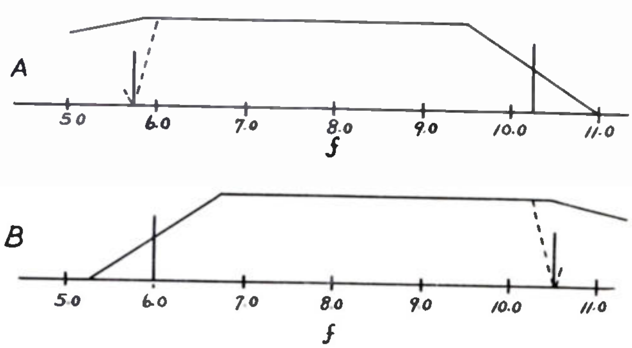


Fig. 4. I.F. Response, ideal amplitude curve.

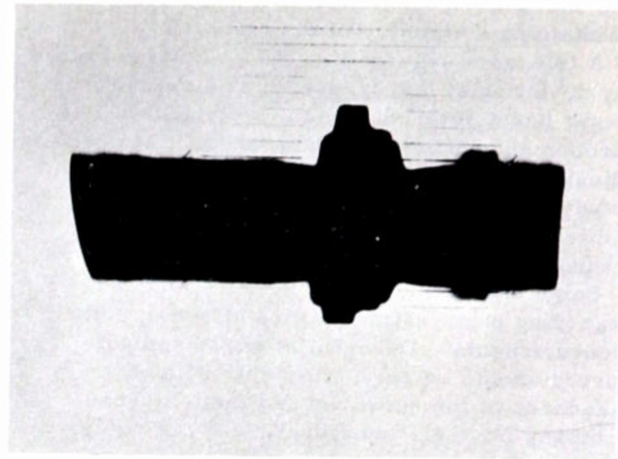


Fig. 5. Modulation envelope of I.F. amplifier, prior to skew filtering.

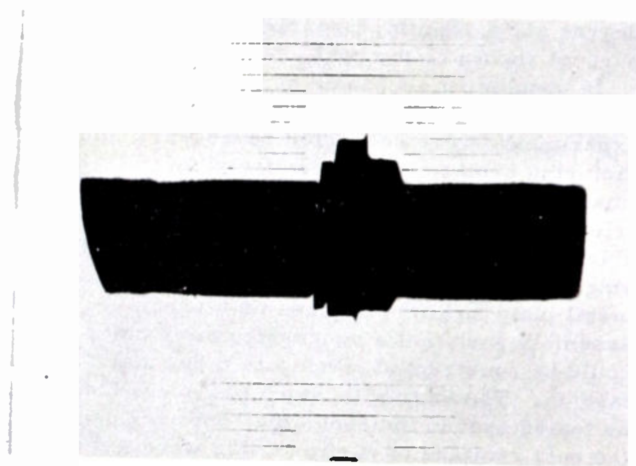
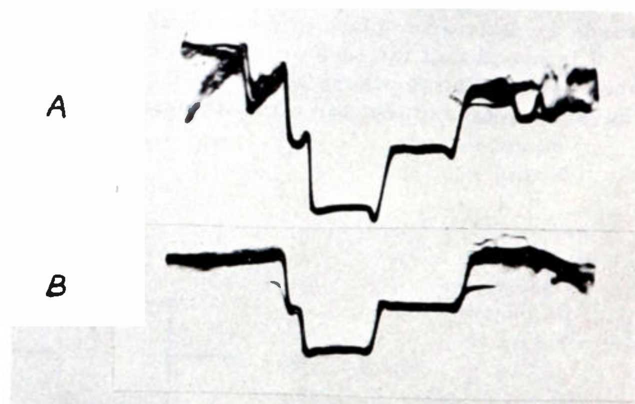


Fig. 6. Modulation envelope following IF skew filter.



Detected sync pulse, (A) before, (B) after phase equalizer.
Fig. 7

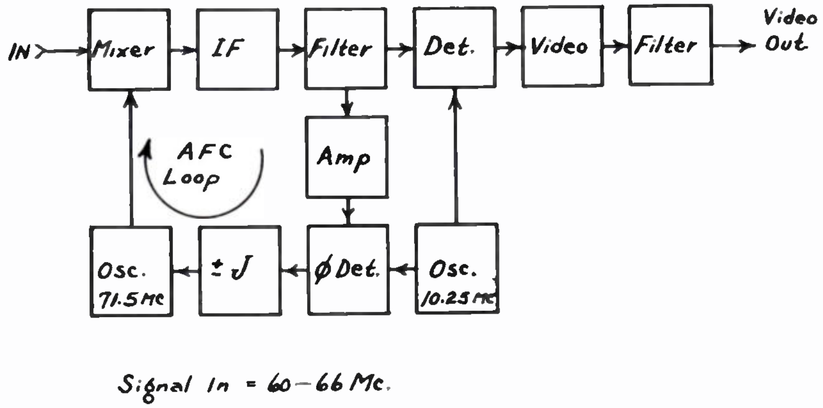
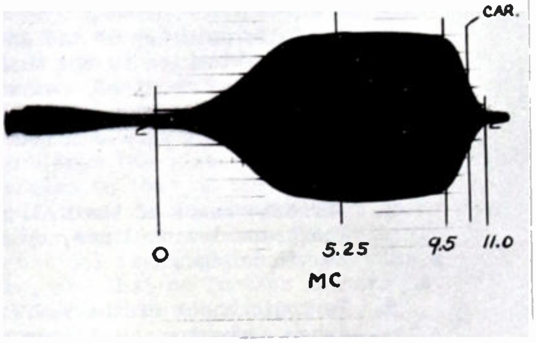
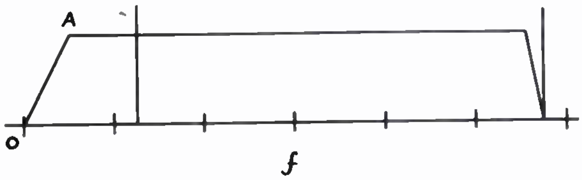


Fig. 8. Block diagram of complete demodulator.



I.F. Amplifier Response.

Fig. 9



Transmitter Spectrum Response

Fig. 12

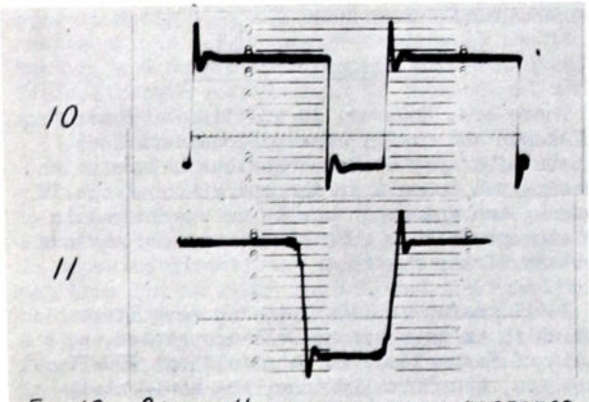
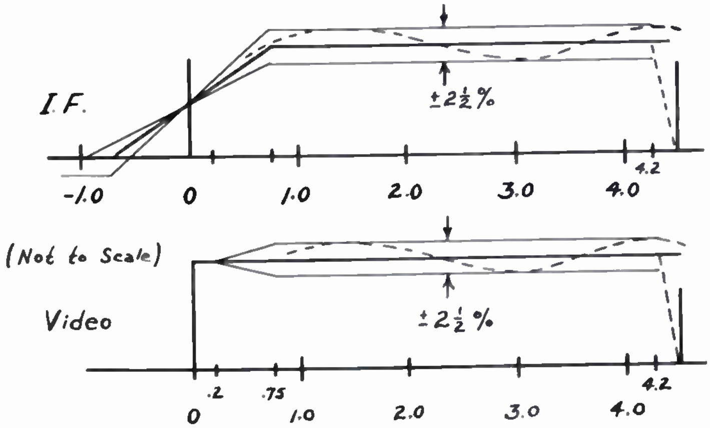


Fig. 10 Step filter square wave response.

Fig 11 Phase equalizer impulse response.



Proposed Tolerances

Fig. 13

OPERATION, MAINTENANCE AND FIELD TESTS OF QUADRATURE-FED ANTENNAS

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Summary

A resume of the more commonly known points involved in the routine operation and maintenance of the "superturnstile" type antenna is given.

The more common methods of measuring the electrical characteristics, V.S.W.R., leakage and short-circuit resistance are listed.

Several additional factors of a more nebulous nature are discussed in detail, with regard to their effect on antenna performance, and with regard to means of proving these antenna characteristics. These factors include horizontal and vertical radiation patterns, batwing power distribution and phase relationships throughout the antenna system.

A measurement technique to check these factors is described in detail and a vectorial method of predicting antenna vertical plane patterns is given, with typical examples of antennas designed to accommodate specific coverage problems.

Introduction

In the routine operation and maintenance of a television broadcasting station, probably the most difficult and, at times, frustrating element of the system is the transmitting antenna. This piece of equipment is exposed to the roughest treatment of wind, rainstorms and heat, and in many cases, highly corrosive atmosphere, and is expected to perform over long periods of time with only superficial, and often inadequate inspection and maintenance.

Since, by far, the great majority of present VHF antennas are of the "superturnstile" or batwing type, I will confine my remarks to this antenna. It should be understood, however, that much of the discussion applies to other multiple-feed antennas, such as the "super-gain" type.

The various manufacturers of TV antennas make certain recommendations regarding operation and maintenance of their products, and these suggestions, which are usually fundamental in nature, if followed, will provide the broadcaster with reasonably dependable and trouble-free service over a long period of time.

General Considerations

Among the usual suggestions regarding operation and maintenance are the following:

1. Periodic mechanical inspection of the complete antenna coax wiring harness for signs of electrical breakdown, arcing, loose or corroded connections, spot heating or any other discrepancy.
2. Inspection of the transmission line system for signs of heating, mechanical wear, maladjustment of spring hangers, or dents in outer conductors.
3. Periodic painting of the antenna, with strict attention to the masking of insulators as required, and with proper surface preparation of any excessively worn areas by wire brush and spot priming.
4. Periodic check of the D.C. resistance of the transmission lines, by means of a Kelvin bridge.
5. Periodic check of the V.S.W.R., using a sweep generator, oscilloscope and well-known techniques.
6. Continuous monitoring of the gas pressure and flow in the system, and the V.S.W.R., as indicated on the reflectometer associated with the transmitter.
7. Periodic check of electrical leakage of the system, with all feed points disconnected, using the "megger" procedure.

These items are all, more or less, of a routine nature, and are well-known to the TV transmitter engineer. Therefore, we will not discuss them further here.

The Specific Problems

There are, however, in addition to these well-known and easily checked considerations, factors of a more or less nebulous nature which are of vital concern to the operation of the TV antenna, and which can easily be overlooked in the attempt to keep abreast of the more obvious problems of the system.

I will refer in detail to two such items with which we have become well-acquainted, as a result of "experience in the field" at KGO-TV. These are, specifically: one, the horizontal; and two, the vertical field patterns of the batwing antenna.

The primary objective of the TV antenna is to provide the necessary power gain in a specified direction in the vertical plane, usually, but not always, toward the horizon, while at the same time providing the necessary pattern shaping to give a strong ghost-free signal in the region near the antenna, at an angle of as much as 10° to 15° below the horizon, depending on the characteristics of particular installation, with regard to population density and antenna elevation above ground.

In general, it can be stated that, as far as the horizontal radiation is concerned, an essentially circular pattern is the normal pattern of this type of antenna, and while the F.C.C. permits up to 10 DB of variance from circular, for coverage purposes, most of the antennas in operation today utilize a circular pattern.

This brings us to the first factor which is to be considered in assessing the operation of a particular antenna system. The phasing of the North-South and East-West elements, in order to provide a circular pattern, must be in quadrature. Any considerable deviation from 90° will create a figure-eight pattern, and the configuration of the conventional diplexer is such that the figure-eight from the visual transmitter will lie at right angles to that of the aural transmitter. It would be normally assumed that the 90° phase relation between the bays is "built-in" by means of identical transmission lines, plus a phasing section, and that no further concern be felt over this item. This is, in general, a valid assumption. As an example, however, of a discrepancy which can develop to change this relationship radically, I will cite a case in which a defective miter elbow at the top end of a 450' run of 3-1/8", 51.5 ohm coax caused a phase shift of approximately 18° in the supposed quadrature relationship.

Due to an error in assembly, the inner conductor of this miter elbow was forced off center ¼" to ½", causing deformation of a teflon insulator, but otherwise making no noticeable change in the system. The lumped reactance of this discontinuity was such that it partially cancelled the residual V.S.W.R., which was approximately 1.1, and thus made little or no change in the reflectometer reading at the transmitter, and only became obvious as a result of field strength measurements of both visual and aural carriers.

Figure (1) shows the resulting ratio of visual to aural field strength which is produced by the error in quadrature phasing cited above. Since the phase rotation of the two carriers is in opposite directions, the two figure-eight patterns lie at right angles, and the resultant ratio of visual to aural field strength takes on a figure-eight pattern as shown. This data is from field measurements taken at random locations in all directions from the antenna.

Field measurements, in attempting corrective measures for a problem of this type, are cumbersome, and so we spend considerable time setting up facilities for direct checking of the antenna.

In order to measure the relative phase of the r.f. voltages at the antenna junction boxes without resorting to field strength measurements, we devised the slotted line arrangement, shown in Figure (2). We were able to produce results reproducible to within 1°, as follows:

1. Two lengths of RG-11/U were run from the antenna to the transmitter room, where all the test equipment involved in this and subsequent measurements was located.

2. Adapters were provided to fit the junction boxes, using end fittings of the type used on the batwing feed lines. Thus, we were able to disconnect batwing feed lines and probe the junction boxes for this test.

3. First, both probes were placed into one junction box, and a 1,000 cycle modulated signal from the visual transmitter was detected by the slotted line and detector. A reference null was thus determined on the slotted line.

4. By moving one of the probes to each of the remaining junction boxes, a new null for each box is determined. The electrical distance between these nulls, measured on the slotted line, is obviously half the difference in phasing between the various test points.

This approach definitely proved an 18° discrepancy in the electrical length of the feed lines, and showed the correct 90° relation following the elbow replacement.

By carrying the above procedure still further, using double shielded cable, (K-125 Federal) and suitable probes with clip connectors, it was possible to check completely the phasing at each batwing feed point with very acceptable accuracy. These measurements require considerable attention to stray radiation pickup, since the voltage at the batwings is quite low, compared to the large fields encountered by the probe lines adjacent to the batwings. This factor is particularly important in measuring the uppermost bays in a 12-bay array, where 3-1 voltage variations are expected between certain bays, with respect to others, for vertical pattern-shaping considerations.

This pattern-shaping aspect, commonly called "null-fill" and/or "beam tilt", is a function of the relative batwing phasing and the relative batwing power level. These parameters, along with the physical spacing of the batwings, determine the vertical pattern characteristics, and therefore the measurement of relative voltage at the junction boxes and at the batwing feed points is a natural follow-up of the phase measurements.

Calculations indicate that nominal discrepancies in V.S.W.R. on individual batwing feed lines can, under certain conditions, create impedance irregularities at the junction boxes which will upset the specified power distribution, and therefore the vertical pattern characteristics.

Following are examples of two typical 12-bay antenna specifications built specifically for superior null-fill characteristics:

Bay Number	Relative Phasing (Electrical Degrees)	Relative Voltage
1 (Bottom)	0°	1
2	0°	1
3	0°	1
4	0°	1
5	-20°	3
6	-20°	3
7	+20°	3
8	+20°	3
9	0°	1
10	0°	1
11	0°	1
12 (Top)	0°	1

Bay Number	Relative Phasing (Electrical Degrees)	Relative Voltage
1 (Bottom)	-70°	1
2	-44°	1
3 thru 11	0°	1
12 (Top)	-70°	1

You will note that the first of these two antennas uses both phasing and unequal power distribution to accomplish the pattern shaping, while the second uses only phasing for this purpose. The phasing is done by simply using different lengths of feed lines to various bays. The unequal power distribution is a little more complex. This requires impedance transformers at the junction box distribution points. It should be noted however, that low V.S.W.R. on each feed line is a prime requisite to proper power distribution in all cases, and that variations from normal can create considerable discrepancies in the supposed "null-fill" pattern.

We were able to accurately measure the batwing feed point voltages, corresponding to the relative fields indicated in the above charts, through the use of an RCA BW-7 field intensity meter.

An r.f. signal of constant magnitude was fed to each of the two main transmission lines individually, and through the use of one of the probe lines mentioned before, relative voltages at each feed point were directly measured by the BW-7 meter.

Figure (3) shows the vertical field pattern of the two antennas cited above, as calculated by the manufacturers, assuming that all phasing and voltages are as specified.

It is possible, by means of a rather simple vector construction, to produce curves corresponding to the above curves for any condition of phasing and voltage distribution, and to visualize the effect of changes in either of these parameters on the antenna characteristics.

Figure (4) is a sketch of the typical turnstile antenna, showing the space phasing between batwings and its effect on the field at a distant point, which is the vector sum of the fields from each radiator.

It is obvious that the amplitude of the received signal at an angle depressed from the horizontal, is dependent on both the phase lag between successive bays, and the built-in phasing by means of unequal feed lines, as well as the amplitude of the signal from each bay.

By adding vectorially these elemental phase voltages for each radiator, a resultant vector can be obtained for the field in any given direction. Then, by plotting the resultants of a family of such curves, it is possible to reproduce the expected vertical pattern with great accuracy. This family of curves, for the two subject antennas, is shown in Figures (5) and (6). The resultants of these curves give us the vertical field patterns shown in the previous Figure (3).

You will note from the vector diagrams that slight changes in power distribution and/or phasing can change the value of the field in the null regions considerably, and it is this factor with which we are greatly concerned.

Conclusion

It is well-known, among TV broadcasters, that there exist many problem receiving areas in relatively close-in locations, where ghosts are prevalent, accompanied by weak signals. This is a problem even under line-of-sight conditions in areas such as Mt. Wilson, near Los Angeles, and San Francisco, where obstructions are at hand to create multi-path.

In our specific case, the extensive antenna measurements and trial-and-error changes which we made prove that, while a 30% fill-in at the first null produces a good TV picture in the problem area at approximately 7° below the horizon, a reduction to 20% creates an almost unusable condition in parts of this area.

The execution of measurements of this nature, where considerable dependence is placed on a rigger to carry out instructions over an intercom, five hundred feet in the air in the black of night, is not the easiest task in the world. However, by careful attention to detail, and with adequate preparation and equipment, much worthwhile data may be obtained on the operating condition of the entire antenna system.

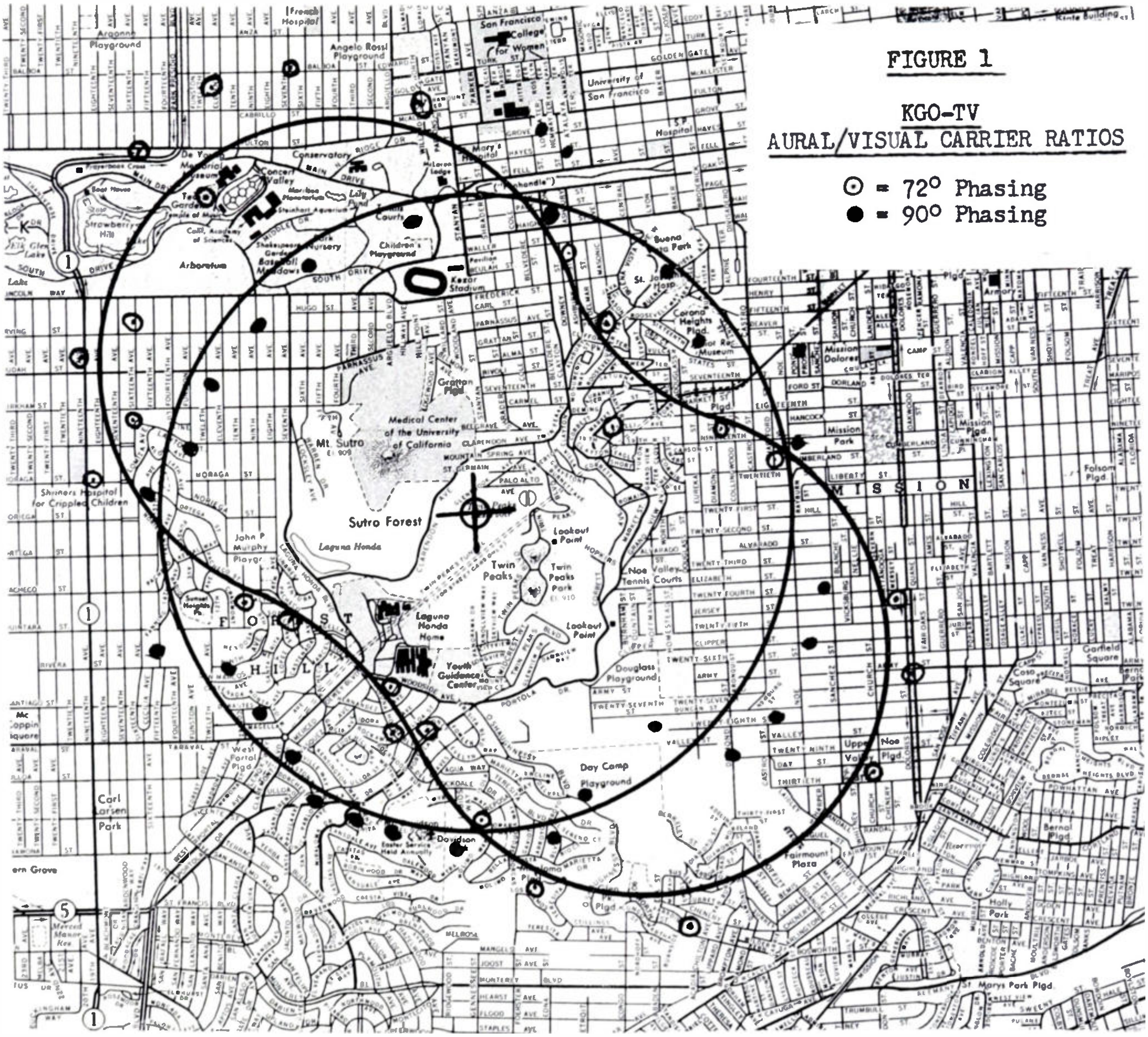
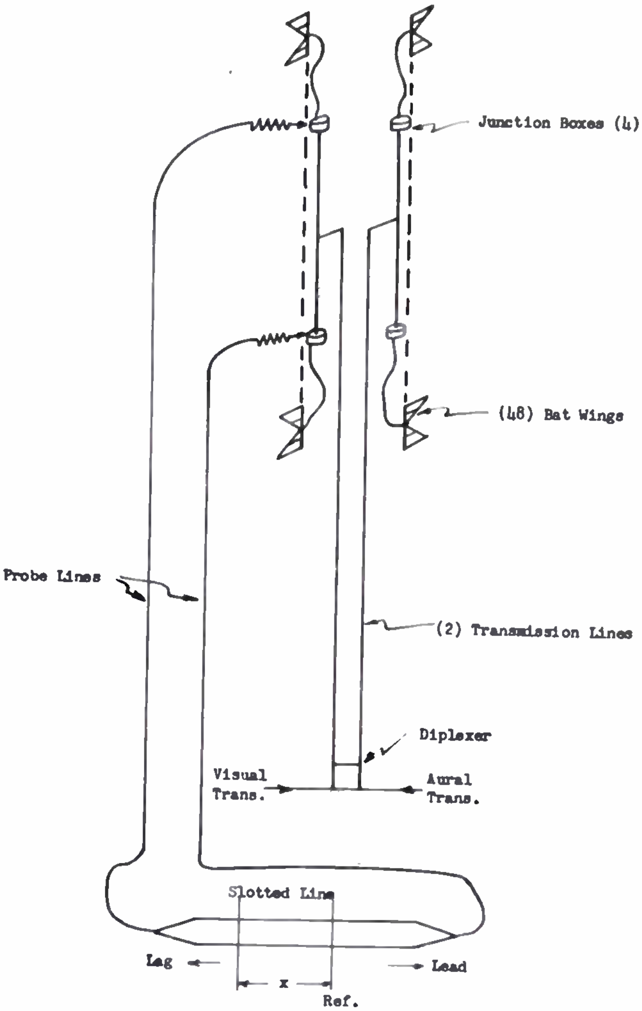


FIGURE 1

KGO-TV
AURAL/VISUAL CARRIER RATIOS



ANTENNA
PHASE MEASUREMENTS

FIGURE (2)

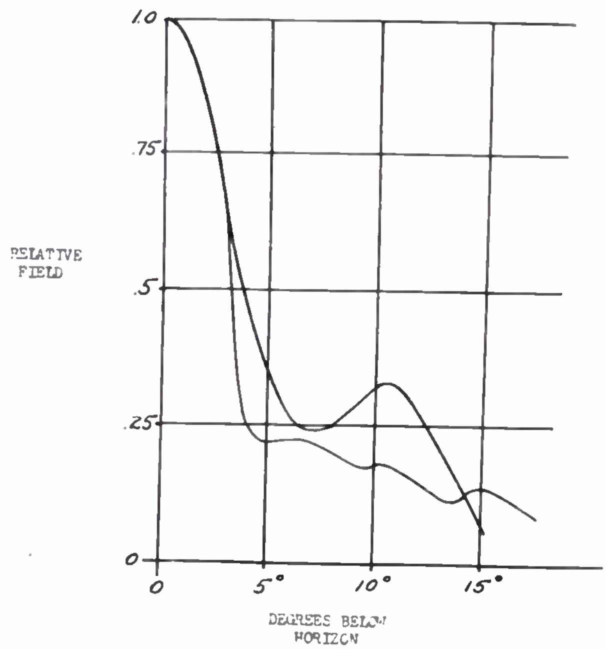
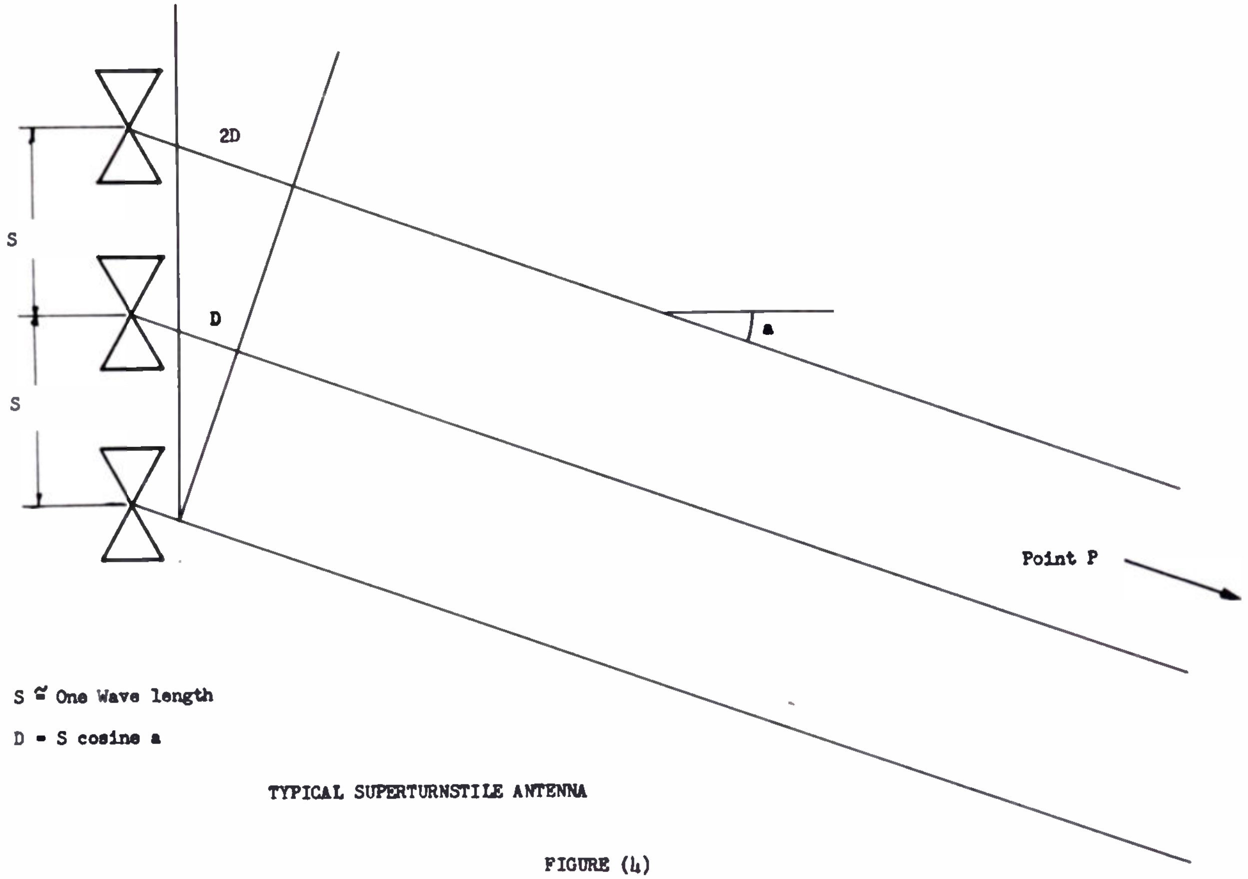


FIGURE (3)

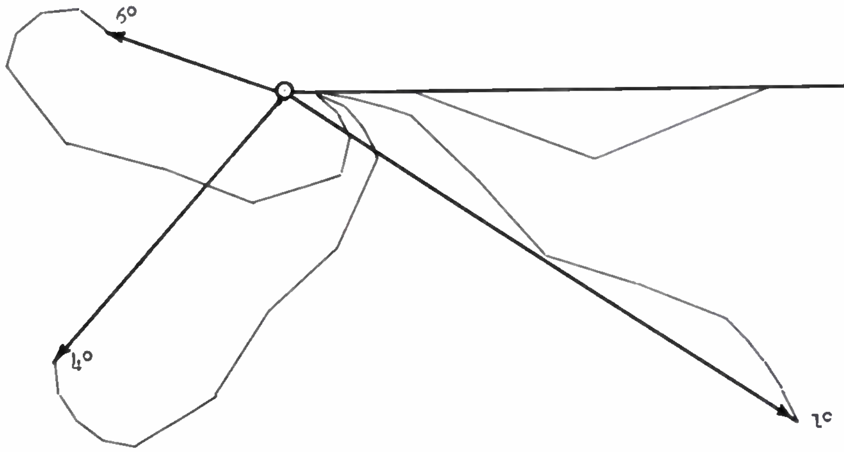


S ≅ One Wave length

D = S cosine a

TYPICAL SUPERTURNSTILE ANTENNA

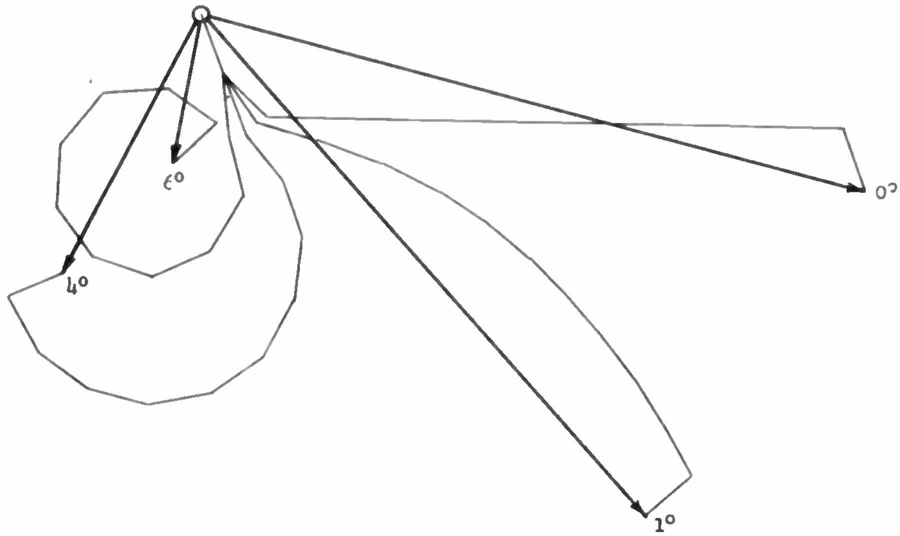
FIGURE (4)



BAY	PHASE	AMPLITUDE
1 (Bottom)	0°	1
2	0°	1
3	0°	1
4	0°	1
5	-20°	3
6	-20	3
7	+20°	3
8	+20°	3
9	0°	1
10	0°	1
11	0°	1
12 (Top)	0°	1

APPROXIMATE VERTICAL FIELD DISTRIBUTION - ANTENNA #1

FIGURE (5)



RAY	PHASE	AMPLITUDE
1 (Bottom)	-70°	1
2	-44°	1
3-11	0°	1
12 (Top)	-70°	1

APPROXIMATE VERTICAL FIELD DISTRIBUTION - ANTENNA #2

FIGURE (6)

TIME AND SPACE PHASING OF TWO-WAY LOUDSPEAKERS

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This paper discusses general considerations involved in the design of two-way loudspeaker systems and reviews the history of its origin.

Fletcher and his group at Bell Labs. demonstrated in 1933 a loudspeaker system which incorporated sufficient frequency range and acoustic power that was capable of reproducing the sounds of a symphony orchestra. It was a two-way system with a single crossover around 300 cps.

The low frequency horn was of the folded exponential type, the air column being 11 feet in length.

The high frequency energy was radiated from a multicellular type horn approximately 3 feet in length. The system was mounted so that the mouth of each horn was in the same vertical plane. This system after being used in the Philadelphia-Washington demonstration was made available for sound motion picture experimental work.

We found that on a tap dance routine, two separate and distinct sounds of the tap could be heard--one coming from the high frequency horn and the other from the low frequency horn. After some study, we realized that there was a path length difference of 8 feet between the two sources of sound. We then moved the high frequency horn back 8 feet so that in effect the time axis of the sound to a point in front of the system was zero. The double tap disappeared and there was a solid tap without echo.

This to my knowledge was the first appreciation of a time delay of this magnitude being important. It was approximately 8 milliseconds and there was a general impression that as much as 50 milliseconds could not be observed.

Today in our theater systems, we have reduced the length of the low frequency to less than 2 feet by using a large throat. In our hi-fi approach, either short horns or flat baffles are used for the low frequencies.

If we will concede that 2 milliseconds becomes marginal in detection, we can arrange the high frequency diaphragm to be back of the flat baffle and reverse the phase on a system using a 500 cycle network since this becomes one-half wave length difference, and if we use a constant resistance dividing network, the phase shift is 180 degrees. If we prefer, we may use an M derived dividing network, and if the M is 0.6, the phase shift is 220 degrees. Thus by combining these principles, we can have a high frequency horn of 18 inches or so mounted in the same cabinet with a flat baffle and meet our overall objective of less than 2 milliseconds.

We have briefly discussed time phasing and also included electrical phasing. Obviously with two sources of sound as obtained from a dividing network system, the power at crossover is divided in two equal parts also. Hence if one unit is out of phase with the other, complete cancellation exists at the crossover frequency. This is easily demonstrated in an anechoic chamber or its equal, a free field. The dip can be on the order of 30 db at crossover but since interference also exists at adjacent frequencies, a fraction of an octave becomes suppressed. This is apparent in a listening test by providing a reversing switch. When the system is out of phase, the sound of white noise or hiss is changed in character. Applause also can be used for the test.

The ability of the listener to detect this change ranges from difficult to being obvious on the first switch, but certainly any discriminating listener has no problem after a few switches to identify the correct position.

Now for some details about arrangement. First, there is such a thing as an acoustic center for loudspeakers.

In the case of a cone speaker, the cone is in a form of a "V". At frequencies up to 1500 cps, the acoustic center is at a point approximately midway between apex and rim. In a 15 inch speaker, this is roughly 1½ inches to 2 inches back of the rim. In the high frequency unit, the diaphragm is much flatter and can be considered to be at the voice coil.

Then there is the problem of distribution. A 15 inch cone becomes directional at 300-500 cps. Thus the beam width for constant pressure is decreasing and the total power radiating is decreasing at 6 db per octave.

This consideration dictates a crossover at 500 cps. If this is not done, there becomes a change in distribution pattern at crossover since the high frequency unit will have a wider pattern.

This is why the larger system or where space is not the all-important limitation, we use low crossover frequencies.

Now a word about shelving. The term "shelving" is used to describe the amount of attenuation that is required usually in the high frequency loudspeaker so that the acoustical efficiency of the high and low frequency units are identical. High frequency units in general when used in conjunction with typical horns require 2 to 6 db attenuation. This usually takes the form of a "T" or "L" attenuator arrangement so that the proper impedance will be presented to both the network and to the high frequency driver. The shelving adjustment is determined at the time the loudspeaker is designed and set so that equal acoustic outputs are present above and below the crossover region. For this reason, the shelving attenuator should not be thought of as a tone control.

In summary, it is the author's opinion that two-way loudspeaker systems can offer the highest quality of reproduction when using dynamic type loudspeakers where accurate design is incorporated in the dividing network, and when an exact exponential flare is maintained on the high frequency horn.

A WIDE ANGLE LOUDSPEAKER OF A NEW TYPE

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The Problem

One major problem encountered in designing high frequency loudspeakers is obtaining a uniform distribution of sound power over a large solid angle. At low frequencies no problem is encountered, the sound covering a wide angle almost despite anything done for it. At high frequencies, however, it is not at all unusual for the power radiated to go to zero at an angle of 20° or 30° from the axis of symmetry of the speaker if there are no reflections. This directionality is a wave length phenomenon. It emphasizes the great variation of wavelengths in the audible range. Between 20 cps and 20,000 cps the wavelength of sound varies from about fifty feet to approximately 0.6 inches. All practical wide-range speakers have dimensions somewhere in the middle of this range of wavelength. For this reason their performance is quite different at different parts of the frequency range. At very low frequencies they behave as point sources with uniform radiation patterns. At wavelengths near the largest dimension of the speaker, however, the radiated sound becomes highly directive and in most cases stays directional as the frequency is further increased. A number of effects combine to cause this directionality. Typical ones are: differences in path length for sound from different parts of the speaker, phase differences between different parts of the speaker diaphragm, and transverse vibrations at the speaker mouth.

Possible Solutions

Any attempt to eliminate directional effects from the output of a speaker must involve creating a roughly spherical wave front leaving the speaker. Several schemes are presently in existence for doing this: one common method is shown in Figure 1.[1] This figure shows a multicellular horn consisting of fifteen exponential horns arranged in five vertical columns and three horizontal rows. All of the horns have a common throat and the mouths form a portion of a spherical surface. The mouth of each is eight inches square and subtends an angle of 17°, this gives a solid angle of 51° x 85° for the entire array of horns.

Figure 2[2] shows the directional characteristics in a horizontal plane for this array of horns. These curves were taken in a plane containing the axis of the center horn and centered vertically at the mouths of the second row of horns.

These curves show a wide variation in the directional characteristics as the frequency is varied. The low-frequency pattern, especially around 500 to 1000 cycles, is very poor. This is an important frequency range for two reasons: the crossover frequency is usually in this range and the average sound intensity in this range is high both in speech and in music.

The directional characteristics of these horns may be explained fairly successfully as the pattern of one large horn at low frequencies and as the sum of the radiation from several highly directional horns at high frequencies. At 8000 cycles, especially, it is easy to see the individual horn patterns. You will notice that the width of this pattern is about 85°, the same as the width of the array of horns. The pattern in the vertical plane is similar but narrower at high frequencies. Figure 2 shows that a speaker of this type radiates power uniformly over a wide angle only when the dimensions of its surface are large compared with the wavelength. The purpose of this paper is to describe a speaker that avoids this size requirement by using concentric horns instead of an array of horns adjacent to one another.

An Alternate Solution

The original idea for this speaker was to use a design like that shown in Figure 3. This is a half cross-section view. The actual horn is the figure of revolution generated by revolving the drawing about the lower straight line as an axis. It was hoped that a spherical wave could be generated by introducing appropriate phase differences between the various horns. Using the Bessell function equations for the output of the various segments of the speaker, the effect of various phase differences was calculated. It was found that a spherical wave could be generated at any one frequency but not over a wide range of frequencies.

A simple modification of the above design was obvious and is shown in Figure 4. As this figure shows, the horns in this version terminate normally to a spherical surface. The solid angle covered is about 165° both horizontally and vertically. If the sound arrives at the mouths of all segments in phase, it should generate a very nearly hemispherical wave front going out from the speaker.

In designing a speaker of this type two problems at once became evident. Using ordinary exponential horns it was impossible to preserve the same ratio of areas for the horns at all points along the axis and also have them terminate normally to a spherical surface at the mouth. The exponential horns gave only one parameter to vary while two parameters would have to be varied to satisfy the two conditions. The second problem was getting the sound to arrive at the mouths of all horns in phase with a variation in path length greater than 25 percent between the inner horn and the outer annular ring.

Preserving Area Ratio

The most obvious way of preserving the correct ratio of areas at all points while still coming out perpendicular to a spherical surface was by using a more general horn equation. With proper variation of parameters the equation

$$y = y_t \left(\cosh \frac{2\pi f_o x}{c} + T \sinh \frac{2\pi f_o x}{c}\right) \quad (1)$$

(the Salmon horn equation) may be used to generate an ordinary exponential horn, a catenoidal horn, a conical horn, or many intermediate shapes. You will notice that this equation involves three parameters; y_t the radius at the throat, is virtually fixed by the size of the driver, f_o is the cut-off frequency for the horn and is limited by the crossover frequency for the speaker. This means that T is the only variable which can be varied with complete freedom, although some variation in f_o is possible as long as it is kept below the crossover frequency. To get the desired shape would be difficult, however, with this many limitations on the parameters.

As the simplest way around these difficulties a slightly different type of horn was designed. This horn is based on the Salmon horn equation but is modified to make a set of concentric horns terminate normally to a spherical surface. This set of horns is illustrated in Figure 5, which is identical to Figure 4 save for the addition of some construction lines.

Circular arcs approximately representing the wave front at various points are drawn on this diagram. The center for all of the circles is on the axis of symmetry of the horn, but the location along this axis varies. Each horn has been constructed so as to intersect all of the circles very nearly at right angles. In designing the horns, the circles were first drawn then curves were constructed to intersect the circles at right angles. A wide variety of horns can be designed in this way by varying the radii and location of the centers of these circles.

An attempt to derive a general equation giving the radii and the location of the centers of these circles showed that only one Salmon horn would be normal to a given set of circles. It proved to be an acceptable approximation, however, to draw all the horns normal to the same set of circles, and this was done. The circles were drawn according to calculations for the inner horn. This means that only the inner horn is a true Salmon horn. The outer sections of the speaker are differences between two approximations to a Salmon horn.

Admittedly, the horns as designed are only approximations to the true Salmon horn, the outer horns especially deviating an appreciable amount from the correct shape. In the completed speaker, however, this does not matter. The equations for horns usually break down when end effects are considered, so that almost any smoothly expanding horn will perform virtually as well as a true mathematical horn.

Obtaining Velocity Shift

The second problem mentioned above was changing the velocity of the sound in the various horns so that the sound will arrive at all parts of the mouth in phase. This is clearly necessary to prevent interference and cancellation effects and give a reasonable approximation to a spherical wave front. Some effective method of changing the velocity is necessary since the path length in the outer annular ring is about 75% of the path length in the inner horn, the path length in the other sections being intermediate. To obtain this velocity change some acoustic delay lines consisting of notches in the walls of channels were tested, but were found to give only about a 10% velocity shift. Since this was insufficient a method discussed earlier but rejected as impractical was revived. This method was to introduce various mixtures of different gases in the different sections of the horn. The velocity of sound in various gases varies, being approximately inversely proportional to the square root of the molecular weight. The velocity of sound in helium, as an example, is about three times the velocity in air. By mixing appropriate amounts of helium and air any intermediate velocity can be realized.

To seal the gases in the appropriate channels, thin acoustically transparent mylar diaphragms were to be used. By making the diaphragms sufficiently thin and putting corrugations in them to keep them from being under any tension, resonances can be avoided. Also, the corrugations can be made deep enough to allow for atmospheric pressure variations.

An obvious problem with this arrangement is leakage of the gas through the mylar film. Fortunately, a study of permeation of gases through solids reveals that no rare gas permeates any metal. This does not mean that the rate of permeation is very slow; there is no permeation. This does mean that a thin film of metal evaporated on the mylar diaphragm will effectively seal the gas in if a rare gas is used to give the velocity shift. The three-to-one velocity shift available with helium was mentioned above. Happily, helium is a rare gas and does not permeate metal. Therefore, it is the gas used in the first model with evaporated aluminum on the mylar sealing it in.

As the sound travels from one gas to the other some sound is reflected at the boundary. By keeping air in the outside section and using as little helium as possible in the other sections this reflection can be kept within 1 db.

Performance of Completed Speaker

Although many more potential problems affecting the performance of the completed speaker were apparent it appeared to be better to build one and test it than to compute the performance any further. This is one of the more common design techniques for acoustic devices, which seldom perform exactly according to the equations anyway.

Figure 6 is a photograph of the assembled speaker. No diaphragm was placed over the outer ring as pure air is needed there.

Tests are presently under way on this speaker and only qualitative data is available at this time. Until more tests are run, the performance can only be described in general terms. It compares favorably with a multicellular horn speaker, although the directional pattern is almost twice the width of that in Figure 2. Some resonances are present, but they should largely be removed by better diaphragm construction.

Acknowledgments. The authors would like to express their appreciation to George Brettel for the original suggestion, and to Nathan Ballard, Jay McKnight, and Walter Selsted for assistance in manufacturing the engineering model.

References:

1. From: Acoustical Engineering, 3rd Edition. H. F. Olson, Copyright 1957. D. Van Nostrand Company, Inc., Princeton, New Jersey.

2. Ibid.

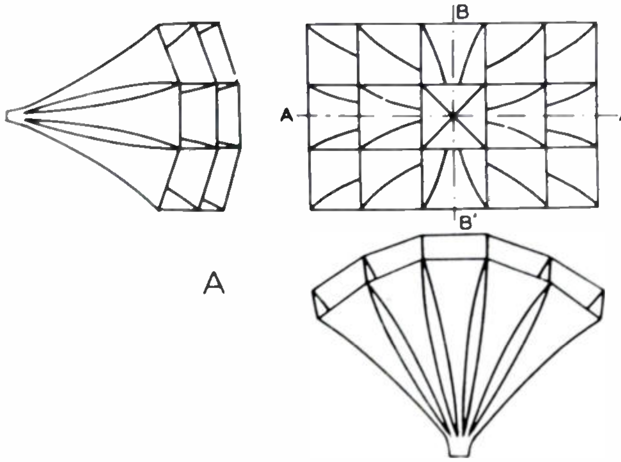

Fig. 1. A spherical radiating surface consisting of 15 individual exponential horns. The mouth of each horn is 8 inches square. Each horn subtends an angle of 17 degrees, both horizontally and vertically.

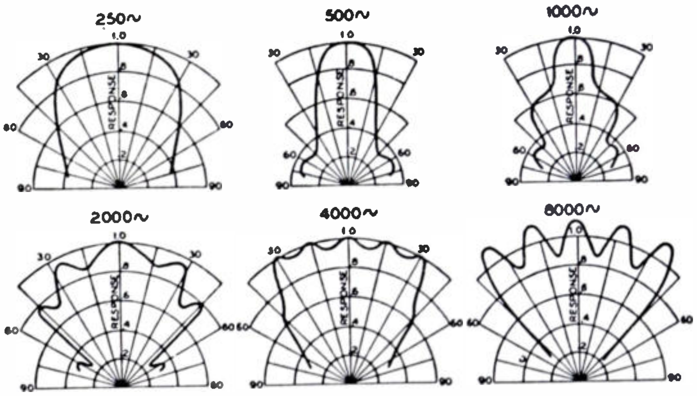

Fig. 2. Directional characteristics of the 15-cell horn shown in Fig. 1 in a plane containing the line A-A and the axis of the center horn. The sound pressure for the angle 0 degrees is arbitrarily chosen.

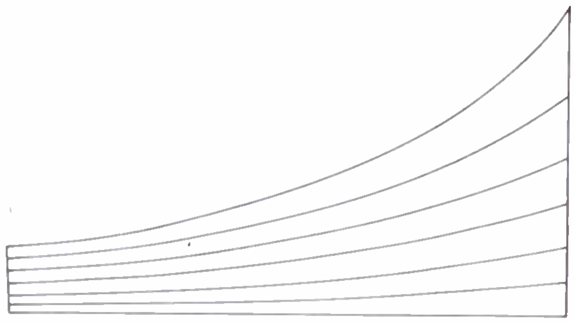

Fig. 3. Half cross section of early version of horns, showing plane surface at mouth.

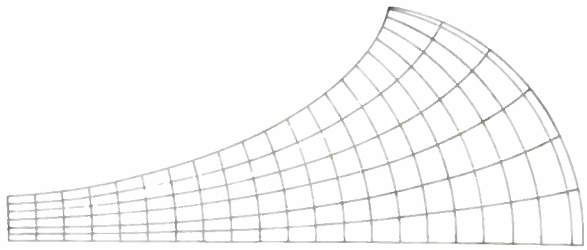

Fig. 5. Revised version of horns showing construction lines used in drafting. The horns are identical to those in Figure 4.

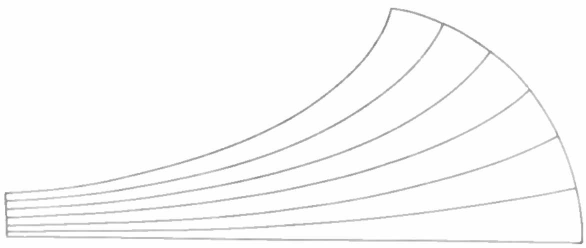

Fig. 4. Revised version of horns, with all horns terminating normally to spherical surface.

Fig. 6. Photograph of assembled set of horns.

POLAR IMPEDANCE EVALUATOR (PIE)

by

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ABSTRACT

In a small laboratory it is often desired to measure the magnitude and phase angle of an electrical impedance at audio frequencies without the sometimes unnecessary accuracy and expense of a bridge. This requirement may be met by a type of device dubbed a Polar Impedance Evaluator (PIE). It is easily assembled from an oscillator isolating matching transformer, electronic ac voltmeter, resistors of known value, and a few potentiometers (volume controls). The PIE is not a bridge, since the voltage across the "detector" position is measured and used in the simple calculations required. Polar components of impedances up to 10,000 ohms are readily obtained to 2 or 3 figures. PIE circuits have been constructed and used for measuring loudspeakers, coils, carrying dc, and for locating resonant frequency as defined by zero phase angle.

Figure 1 gives the basic circuit of PIE on the left. A signal generator is attached at the top and a voltmeter is used to measure volts E_{12}, E_{13}, and E_{14} at the respective terminals 1, 2, 3, and 4. R_0 is a reference resistor, and R is the adjustable resistor (volume control) for an adjustable reference point.

When setting up the bridge, R must be about 10 times R_0 for reasonable accuracy, and the measuring voltmeter (using VTVM for high input impedance) should have an input impedance about 10 times R. For higher frequencies, above audio, an isolating transformer for the signal generator would be necessary.

In operating PIE, E_{14} is minimized by adjusting the tap on R. This minimum is generally broad unless the unknown impedance is near resonance. Then E_{12}, E_{13}, E_{14}, and the frequency are recorded.

On the right of Fig. 1, a voltage vector diagram is shown with the points 1, 2, 3, and 4 referring to the terminals 1, 2, 3, and 4 on PIE. Since the current I flows through both R_0 and Z, it can be eliminated from the two equations $IR_0 = E_{13}$ and $I|Z| = E_{12}$. Dividing the first by the second,

$$\frac{R_0}{|Z|} = \frac{E_{13}}{E_{12}}$$

and with an additional algebraic step,

$$|Z| = R_0\,(E_{12}/E_{13})$$

The minimization of E_{14} assures that E_{14} is perpendicular to E_{23}. This divides up triangle 123 into two right triangles, 124 and 134. From geometry, $\theta_1 + \theta_2 = \theta$ (derived from the fact that the sum of the interior angles of a triangle is 180 deg). Since triangle 124 and 134 are right triangles, $\sin\theta_1 = E_{14}/E_{12}$ and $\sin\theta_2 = E_{14}/E_{13}$. Using the inverse sine and summing θ_1 and θ_2 which equals θ, the phase angle of $|Z|$ is:

$$\theta = \sin^{-1}\left(\frac{E_{14}}{E_{13}}\right) + \sin^{-1}\left(\frac{E_{14}}{E_{12}}\right) .$$

The necessity of looking up inverse sine can be eliminated by calibrating a meter scale in the inverse sine. This is shown in Fig. 2. To measure θ_1, E_{12} is adjusted at the oscillator to read 90 deg, then the meter is switched over and E_{14} is measured and θ_1 is read off in degrees. The same system is used to measure θ_2, and then, θ_1 and θ_2 are added to equal θ.

As you might have noticed there is an ambiguity in the sign of the phase angle θ. The easiest way to determine the sine of θ is to shunt R_0 with a capacitor whose value is approximately $0.6/R_0\omega$. Figure 3 shows the effect on the voltage vector diagram by dotted lines when the capacitor is in parallel with R_0. Both an inductive and a capacitor impedance triangle are shown. For the inductor impedance, upper triangle, E_{14} is increased when the capacitor is added, and for the capacitive impedance, E_{14} is decreased.

GENERAL

The concept of PIE is to make the circuit and the calculations as simple as possible, three measured voltages and two equations to plug into. All that is needed is an AC voltmeter and since all the voltage measurements appear as ratios, there is no need for absolute voltage. PIE is a modern version of the old 3-voltmeter method. Other devices which read θ directly are more expensive and complicated.

The idea for PIE originated at Jensen Manufacturing Company, Chicago, in 1947.

It is easy to take the basic circuits of PIE and adapt them for particular measurements and techniques.

Measurements of incremental inductances (Fig. 4) can be made by putting in a DC supply and a by-pass capacitor, C, in series with R, where $R\omega C >> 1$. The inductive part of Z is perpendicular to R_0, and is shown as $E_{25} = I\omega L$. Triangles 352 and 314 are right triangles. Therefore, $\sin \theta_2 = E_{14}/E_{13} = E_{25}/E_{23}$ by using relationships that $E_{13} = IR_0$ and $E_{25} = I\omega L$ it is easy to obtain that

$$L = \frac{R_0}{\omega} \frac{E_{14} E_{23}}{E_{13}^2}$$

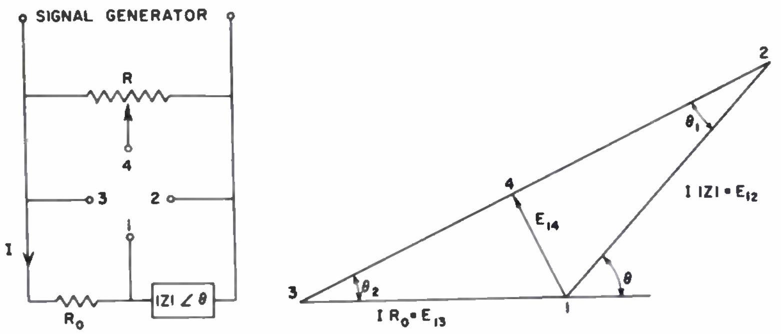
A precaution to take is that the voltmeter must be floating for measuring E_{23}. This procedure can be used to measure chokes, magnetic amplifiers, variable reluctance devices, loudspeaker fields, and similar devices. Also possible is the measurement of incremental capacitances. The measurement of a-c and d c voltages can be made across R_0 to set conditions.

Another modification of the basic PIE circuit is shown for direct reading of the magnitude Z in Fig. 5. Here R_0 is changed until $E_{13} = E_{12}$ and then the magnitude of Z is read off of the calibrated pot R_0. (A hellipot does not work well for R_0 since it is not purely resistive over the audio band.) To obtain decade switching, additional resistors, R_1, R_2, R_3, R_4 shunting R_0 are used. These resistors are chosen so that the total resistive changes by ratios of 10.

The value of R should be much larger than any other circuit impedance and that of the VTVM much greater than R. The phase angle can also be obtained in this circuit by using
$\theta = 2 \sin^{-1} (E_{14}/E_{12}) = 2 \sin^{-1} (E_{14}/E_{13})$.

Figure 6 shows an electronic adaptation of PIE. Cathode followers are used to obtain high input impedance, and thus increase the range of the high impedance end. This circuit is essentially the same as the one in Fig. 5 and therefore does not need additional description.

Figure 7 shows some measurements made using PIE of a coaxial loud speaker. The speaker was in a bass reflex cabinet and shows the usual double peak impedance curve. The little bump on the second peak is caused by cabinet resonance. The speakers were crossed over at 2.5 kc with a capacitor which accounts for the impedance coming down at the high frequency end. The phase changes show up nicely by this method, and, as they should be, they are steep at the resonance peaks of the base reflex cabinet.

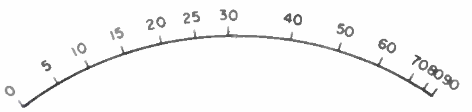


$|Z| = R_0(E_{12}/E_{13})$

$\theta = \theta_1 + \theta_2 = \sin^{-1}(E_{14}/E_{13}) + \sin^{-1}(E_{14}/E_{12})$

POLAR IMPEDANCE EVALUATOR (PIE)

Fig. 1



LINEAR SCALE

$\theta = \arcsin E_{14}/E$

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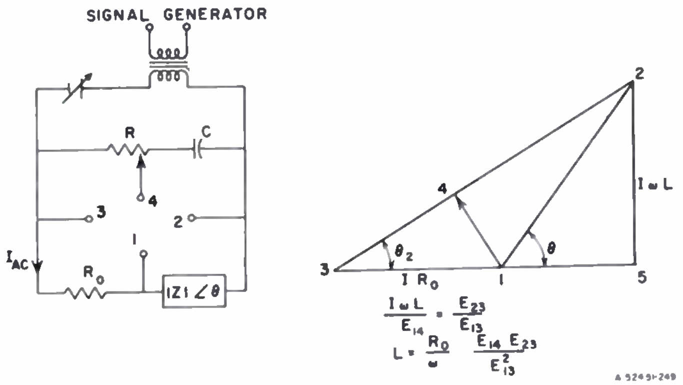
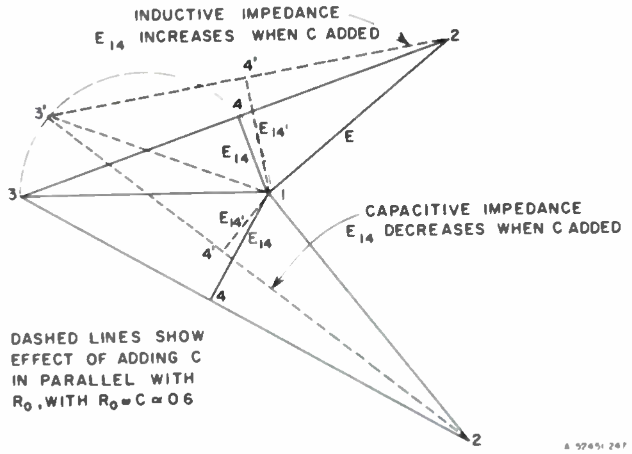
ANGLE SCALE FOR METER

Fig. 2



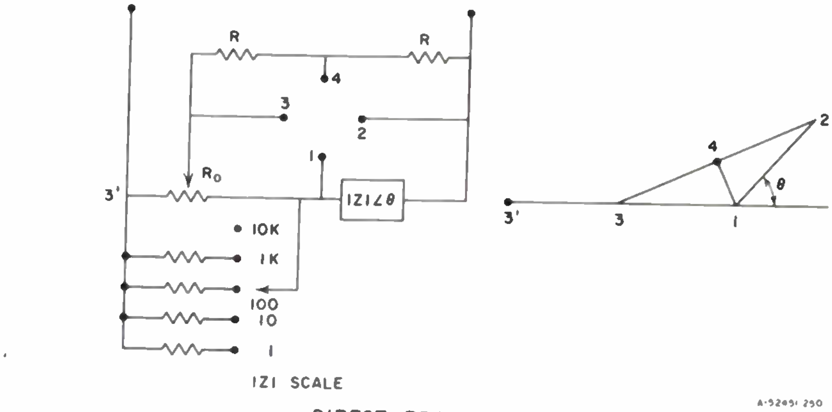
MEASUREMENT OF INCREMENTAL INDUCTANCE

Fig. 4



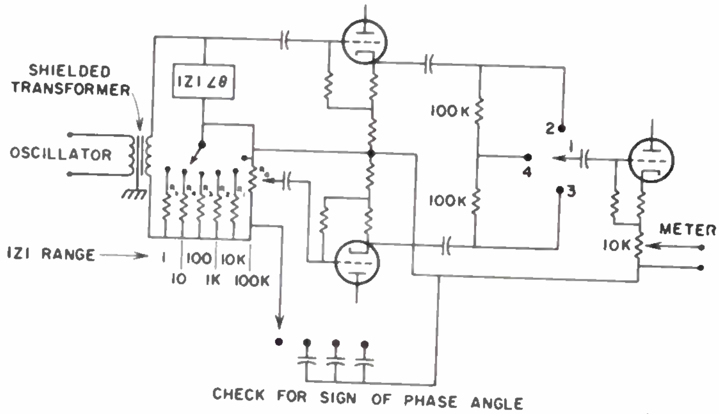
DETERMINATION OF SIGN OF θ

Fig. 3



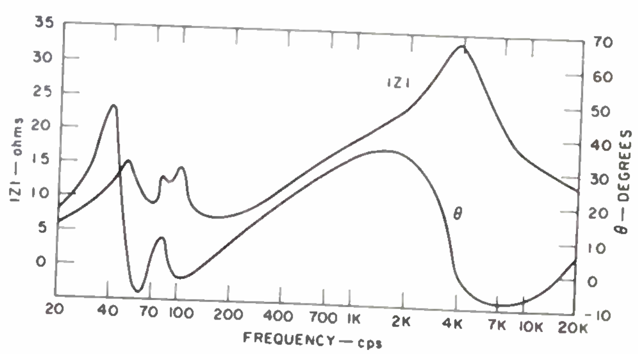
DIRECT READING FOR |Z|

Fig. 5



IMPEDANCE-PHASE METER

Fig. 6



|Z| ∠θ OF AN 8-ohm 12" COAXIAL SPEAKER IN A BASS REFLEX CABINET

Fig. 7

MULTI-CHANNEL AUDIO RECORDERS

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The recent accent on commercial stereophonic recordings has resulted in a demand for two-channel, three-channel(and even more channels)recording instruments, both for making masters and for home use. The design considerations, operational techniques, advantages and disadvantages are discussed.

The unique use and operation of an eight-channel audio recorder for a well-known figure in the entertainment world is described.

If you were faced with the problem of creating a new product for your company, how would you go about it?

Les Paul and Mary Ford, the well-known and very popular entertainers, had such a problem. Guitar playing and vocalizing is old stuff, you might say. But consider the problem of an entertainer who must come up with something new to intrigue his listeners--that is, a new product. In their case, a new product would be a new and intriguing sound. One may consider such an abstract and transitory thing as sound as being a new product.

How does one get a new sound? After all, sound is sound. But Les and Mary invented a new recording technique to intrigue their listeners. This technique required a new type of recorder with a different head arrangement. The normal recording function on a tape recorder uses the head arrangement shown in Figure 1. The erase head is followed by the record head and the playback head in that sequence. Their new machine had a slightly modified head arrangement shown in Figure 2. Here, the playback head is placed before the erase head and followed by the record head. With this special head arrangement, they were able to listen to the playback of a previously recorded signal while they simultaneously mixed it with the new and additional signal to be recorded. This is known as "sound-on-sound". The functional arrangement for accomplishing this task is shown in Figure 2. This process of mixing a new signal with a previously recorded signal was then repeated a number of times until the desired new sound quality was achieved.

Their system as such had a major disadvantage. The previously recorded signal is erased and the resulting master is a composite one. Thus, if there are any errors made during any one of these subsequent re-recordings, the master becomes useless and the entire process must be started over again. Such incidents actually happened to Les and Mary. On a particular six-step recording, the master was spoiled on their sixth and final re-recording process. Unfortunately for them, it was not because of an error on their part but the misfortune of having a telephone situated near the microphone in their converted studio-garage.

Because of such frustrating experiences, it led them to the purchase of two more recorders with a normal head arrangement. Thus, they were able to play back the previous recording, mix it with a new signal and record it on the second machine. This process was then continued back and forth between the two machines until the desired number of re-recordings were made. In this manner, the master in each step of the way was preserved. Any errors made during any one re-recording process could be done over any number of times without starting from scratch. With this system, reverberation techniques and other sound effects were used. This system was adequate, although there was still much more that could be, and was done later. Their experiments with these two systems paved the way to a very special multi-channel audio recorder to be described shortly.

Another special use for an audio recorder is in a machine appropriately called a Language Recorder--a machine used in the teaching of foreign languages. Reproducing facilities are provided on both channels of a two-channel machine. However, recording facilities are available on one channel only. On a separate special half-track recorder, the instructor records the lesson material on the channel which does not have the recording facilities on the language recorder. The instructor's tape is then placed on the language recorder and the student is then able to listen to the instructor. The student may at any time record on the second channel repeating the instructor's lesson material. He can then play back the tape listening to either or both channels, separately or simultaneously, to get his enunciation as precisely as he desires. This method of teaching languages has become highly effective.

Let us look at another new product--stereophonic recording. Stereophonic sound is not a new idea in the field of audio perception. Experiments and ideas were carried out as far back as 1893. However, the practical method of bringing stereophonic sound to the homes of millions is a new product. Also, the practical method of recording and retaining this sound may be considered a new product.

Multi-channel audio recorders are tools to be used in the creation of new and different sounds to meet the demands of the public. To create new and different products or results, one sometimes must modify a basic tool in some manner. The standard monaural or single channel audio recorder may be considered our basic tool. From this mechanism have come about a great many audio recorders of greater complexity used to achieve the desired end result. The advancements made in multi-channel audio recorders and recording techniques have been quite rapid, but by no means has it reached the limit of capability and advancement.

It is generally agreed that three channels are the optimum minimum number necessary to create the auditory illusion of depth and position. Such a system could be afforded by the fortunate few in the higher income brackets. But in an effort to bring stereophonic sound into the home of the average listener, two-channel systems were investigated. After considerable experimenting in the two-channel recording and reproducing techniques, satisfactory results were obtained. Figure 3 illustrates the two-channel stereophonic system. Such a two-channel stereophonic system has been accepted by the public as a practical and adequate means of reproducing stereo sound.

Stereophonic sound can be recorded and reproduced by two independent tape recorders; and for that matter any other type of medium of recording. However, the synchronizing problem with two separate machines is very critical and not very practical. The next step, of course, was to put these two channels on one machine. This is the "Staggered-Stereo" system. In this system, one head records the upper track and another head, spaced about 1-1/4 inches away, records the lower track.

This arrangement led to a practical system. However, the problem of maintaining the same head-to-head, or gap distance from one machine to another was critical. Head spacing is critical because it is necessary to keep the time error between tracks to a minimum. This time or synchronizing error due to inconsistent gap spacing gave rise to what might be called "Floating Sound", sound that produces the illusion that the instruments are wandering from side to side. This illusion is caused by phase change of any given sound arriving at the two recording mikes, as the effective head spacing may change due to stretch or side to side wander of the tape over the two spaced heads. Moreover, this phase shift varies with frequency. Hence a complex tone, such as that produced by a violin, creates a weird wandering effect. To eliminate this effect the gap distance must be maintained on the order of a few ten-thousandths of an inch from machine to machine for proper reproduction. This may not sound too fantastic a figure when we look at the fact that a 15 kc signal at 7-1/2 inches per second has a wavelength of one-half of one-thousandths of an inch. Thus, if the gap spacing error is one-fourth of one-thousandths of an inch, the sound emerging from the two speakers will be exactly opposite in phase at 15 kc. It would therefore seem impractical to build a "Staggered-Stereo" recorder-reproducer because such a system would require the use of five heads--one erase head, two record heads, and two playback heads.

To eliminate the criticalness of the gap spacing adjustment as well as to be able to build a practical stereo recorder-reproducer, multi-channel recorders were designed, whereby the heads are stacked one above the other. The gap spacing error was then a function of the techniques used in the manufacture of the heads rather than a critical adjustment that the operator must attempt. This is the "Stacked-Stereo" or preferably "in-line" head construction, which has now been adopted almost universally for stereophonic recording and reproducing in this country.

Since the advent of the two-channel stereophonic system, the demand for two-channel audio recorder-reproducers has increased; first for the professional recordist and more recently for the "Hi-Fi" hobbyist. These machines are built with a full-track erase head, which erases both of the channels with one head acting over the full width of the tape. Two-channel machines are also built with individual erase heads for each channel if the customer so specifies.

A problem was initially encountered in the design of such a separate erase machine. As you know, if the two separate erase-bias oscillators are not generating approximately the same frequency, a beat note is developed between the two signals which is recorded on the tape. An obvious solution that has proven successful is in the use of a "Buffer" amplifier in each of the erase head circuits; all such circuits are then fed from a common oscillator.

These two-channel recorders were adequate for a majority of the recordings attempted. However, for the professional user in the recording of tape masters, a third channel was sometimes desired for the recording of a soloist, dubbing a foreign language, or for special sound effects, etc. One method of using the third channel is illustrated in Figure 4. The third channel, used for the recording of a soloist, for "filling in the middle", or for other sound effects, is mixed into the other channels to produce the two-channel stereophonic tape sold to the customer for his two-channel reproducing system. This method of recording produces a more vivid and well-balanced musical interpretation more pleasing to the listener.

The first three-channel recorder was made for a quarter-inch tape with a full-track erase head. The individual track width of this machine naturally had to be made narrower than the two-channel machines so that all three channels could be accommodated on the quarter-inch tape. This narrower track width degraded the signal-to-noise ratio as well as increased the percentage of "drop-outs". "Drop-out" is the momentary loss of signal resulting from imperfections of the tape. This problem becomes increasingly more annoying as the width of the track is decreased.

To preserve the high order of signal-to-noise ratio demanded by the recording industry, half-inch wide tape was resorted to. This eliminated the undesirable features of the three-channel, quarter-inch tape machine and has become, so to speak, the "standard" of the professional recording industry. These three-channel, half-inch tape machines, like those of the quarter-inch variety have been made with both a full-track erase head and separate erase heads and sometimes with no erase heads, depending on the customer's wishes. The distortion, signal-to-noise ratio, frequency response, and recorded signal level have been maintained on each of the three tracks as on a standard single channel recorder.

The use of a full-track erase head as against erasure for each channel is dictated by the individual user's preference. Both have been used successfully. Some have eliminated the erase head entirely, claiming that the erase head gives rise to noise and distortion. This is a matter of opinion--and the tape used. Even without the erase head, bias signal must still be supplied for the record head, which can cause similar results. However, if the source of the erase signal is properly designed and adjusted--whether it be from the oscillator or a combination of oscillator and buffer--it should give rise to very little additional noise and distortion, if any.

Let us get back to Les Paul and Mary Ford's exploits in producing their new sound. Their basic recording technique is known as "Sound-on-Sound" recording--the method of overlaying one sound upon another sound on a single track. Not only did they mix the previous recording with the new signal, but their technique required the feature of several previous recordings mixed in various combinations in order to obtain the desired sound quality. Thus, with their original machines, it can be seen that synchronizing problems were inherent. Also, in progressing through the re-recording process, sometimes as many as 24 times, degradation of sound in both frequency response as well as noise and phase errors caused the final recording to be rather muddy--not as crisp and clear as they wished it to be.

Les Paul felt that there might be other ways and means of accomplishing better end results. This led him to dream up a new machine for his particular purpose of overcoming the difficulties he was experiencing with his original system. It was at this time that Les Paul came to us with an idea for an eight-channel multi-track recorder-reproducer, utilizing a one-inch tape. Figure 5 shows the completed machine. Each channel was to have its own erase head so that he would be able to re-record on any channel he desired without affecting the others. At the same time, the inter-channel synchronizing of each new track recorded one at a time had to be maintained so that the composite playback of all tracks, mixed together, would be in unison. One must remember that the playback head follows the recording head position-wise on the tape. Therefore, if Les and Mary monitored the last recorded track made so as to keep the musical beat, while recording on the next track, the composite playback of the two tracks would be out of step by the time interval equal to the spacing between the record head and the playback head. This could become disastrously cumulative on eight tracks! The basic question in the synchronizing problem for us as machine builders was in figuring out how to make a record head think it was a playback head. It may sound strange but the answer was obvious--hook up the record head and use it as a playback head. This was done by designing the heads and switching circuits with care so as to minimize noise pickup in these very sensitive low-level circuits.

Various other special features were incorporated in this machine which Les Paul has nicknamed the "Octopus". Among them is a special switching panel for the pre-setting of any of the eight channels for recording. Also incorporated were manual or remote monitoring of the proper head for synchronizing, and complete remote mechanical control facilities.

This machine has given him the tool to execute some new recording techniques to yield new and crisp sound. The possibilities for such an eight-channel machine are very great. He can produce additional sound effects by proper mixing of the various channels. Reverberation techniques can also be used as well as other gimmicks to put his new sound across to the public. The possibilities are so numerous that it is doubtful whether even Les Paul has discovered all of them yet--even with his great knack for finding new and better ways of producing his sound.

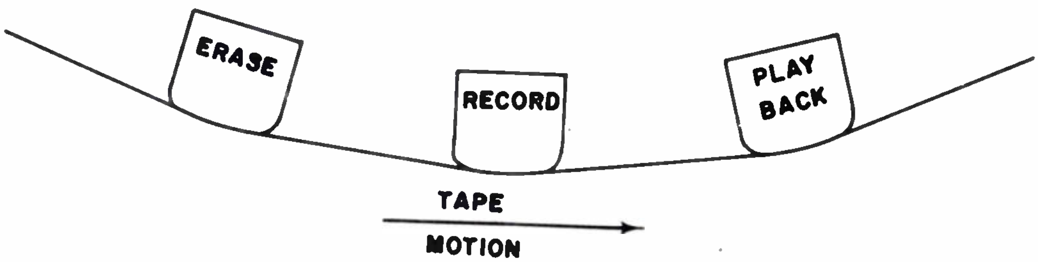

Fig. 1. Normal head arrangment.

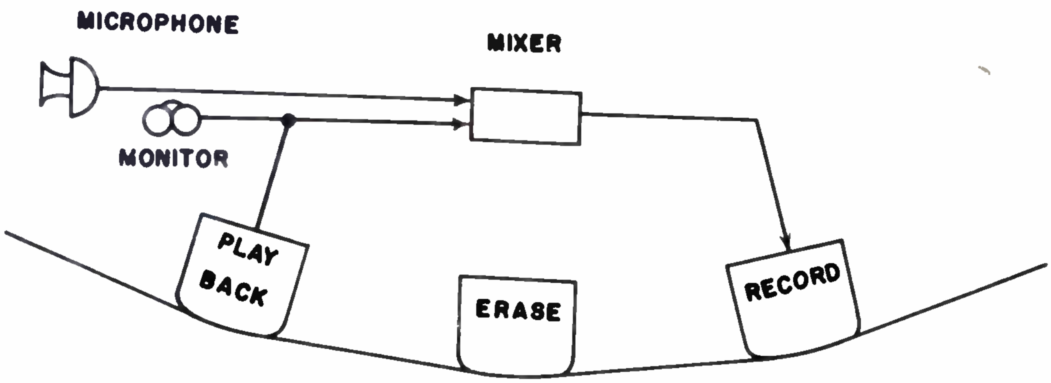

Fig. 2. Modified head arrangement for sound-on-sound recording.

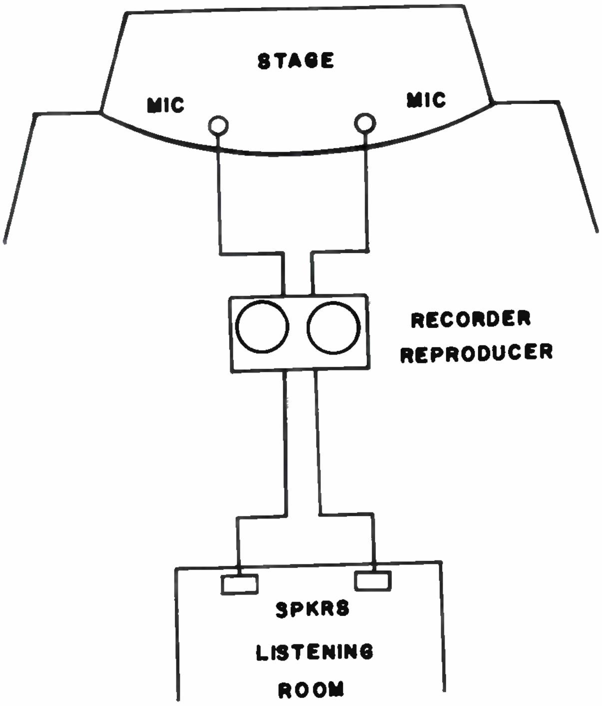

Fig. 3. Two-channel stereophonic system.

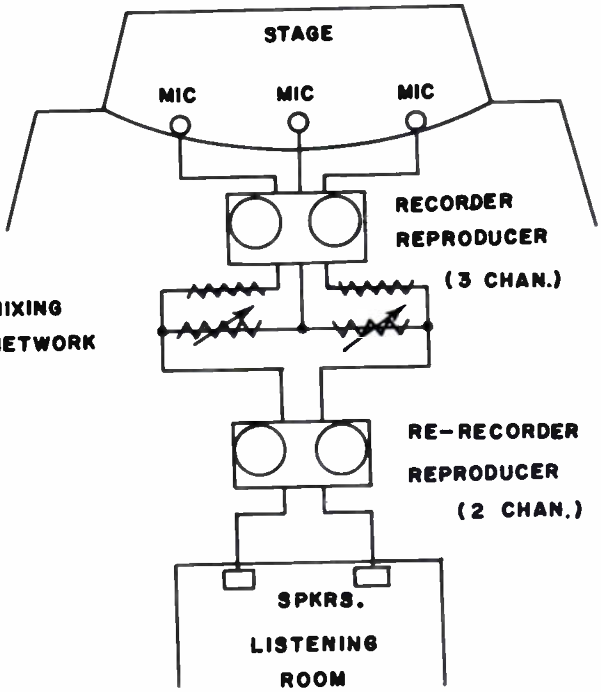

Fig. 4. Method of using three-channel recorder for two track stereo releases.

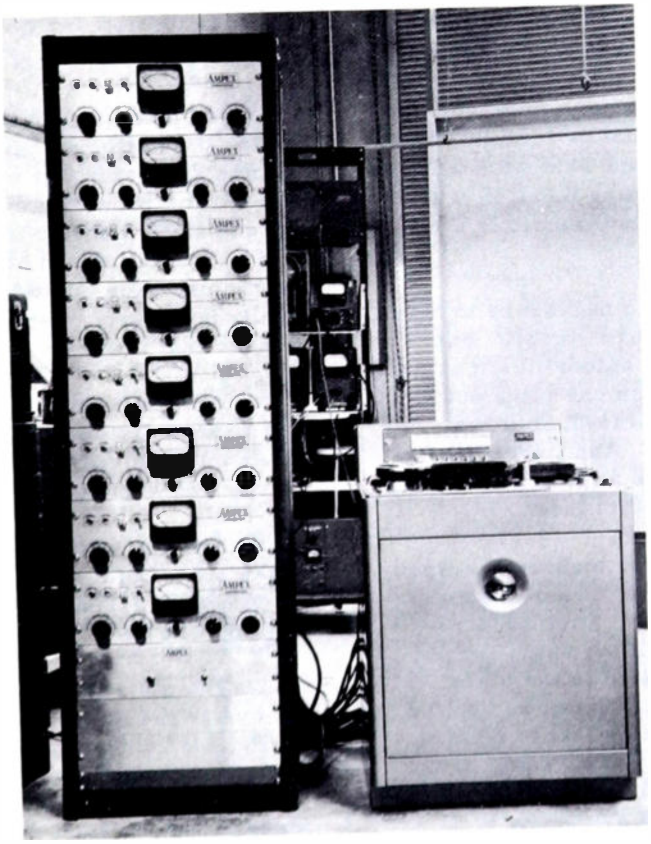

Fig. 5. Les Paul & Mary Ford's "Octopus".

METHODS OF RECORDING COMMERCIAL STEREOPHONIC MASTERS

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The problem involved is to provide the listener with the illusion of auditory perspective. The first work of this nature was the celebrated Bell System transmission from Philadelphia to Washington in 1933. Since 1950 the use of these techniques has resulted in commercially obtainable tape recording and has also formed the basis of certain new LP techniques.

A quiet controversy is raging among many engineers, both in this country and in Europe, as to what is the proper(?) stereophonic recording technique. Various terms such as "binaural," "stereophonic," and "auditory perspective" are defined.

If someone were to tell you that you could make a stereophonic recording with one mike, would you believe it?

Those of you who are somewhat familiar with the problem will answer that question with some word such as "preposterous"! Those of you who are less familiar with the problem may have no reaction at all and start yawning. But all of you know that with just one mike you can get a certain kind and amount of realism, but in comparison with stereo--and what well informed person hasn't heard something labelled stereo--single channel reproduction of music is a bit dead.

Everybody knows it is necessary to have more than one mike to record stereo. But according to some English authorities it is not necessary to have more than one mike position to record stereo. This is known as the "crossed mike" technique. And it works. Whether it is good or not is a matter of subjective opinion. But more of this later.

Stereophonic sound, or auditory perspective, as it was called by Bell Telephone Laboratories in 1932 and '33, is an attempt to re-create in the listening room the same sort of acoustical pattern that existed in the place where the sound originated. It can be applied to every type of sound such as that of an airplane flying by, a symphony orchestra, or a piccolo player performing in the Painted Desert. Most often it ranges from a symphony to a string quartet.

The main objective of commercial stereo recording is to bring enjoyment to the listener in his own home by endeavoring to re-create something resembling the sound that this special listener connects in his own mind with a live performance he once heard somewhere.

This assignment can be quite a task!

Of primary importance is the space in which the customer has set up his stereo reproducing apparatus. It may be in a reverberant recreation room, an overstuffed living room, or an odd shaped room. Or the apparatus may be set up incorrectly in acoustically the very finest of rooms. The commercial recordist has virtually no control over this situation. Only an educational program can help.

The customer, moreover, may not know for what he listens, or if he does, how to get it. Over this subjective situation the recordist has no control whatever. All he can have is hope.

The professional recordist must therefore make some bold assumptions. This he has done. To see what assumptions must be made some history will help.

The main proposition has always been to re-create the sound with life-like realism. J. P. Maxfield, of Bell Telephone Laboratories, devised a complex mathematical formula about 1937 for the placement of two microphones in a room to secure what was felt to be the best possible single channel pickup. It involved specific numbers for such things as room size, reverberation time, number of performers, plus some numbers for some abstract, subjective factors as well. He recognized that the single channel listener obtained an illusion of presence through two means: the relative loudness of the various instruments in an orchestra, say, and the apparent distance from the instrument to the mike. These factors created a feeling of position over a single loudspeaker without actually revealing it as can be achieved in a stereophonic pickup and reproduction using more than one channel.

In the Maxfield pickup one non-directional mike was placed fairly well away from the performers at a point about where the direct sound was the same loudness as the reverberant sound. This picked up the overall sound including generous quantities of room reverberation. Sound from this mike lacked considerable "definition". Therefore an accent mike was placed closer to the source than is usual for a one mike pickup. This yielded the sharp intonation or definition necessary. The output of these two mikes was mixed and judged over a monitor speaker.

It is well to realize that Maxfield was endeavoring to compensate for a most important item. The human being with two normal ears can concentrate his attention on any given sound to the relatively high exclusion of all other sounds. This is a combination of brain and ears acting as a discriminator. But take away the use of one of those ears by placing a finger in the outer ear canal and the discriminator suddenly fails to function as a concentrator. Maxfield's accent mike was a monaural substitute for this human attention-focusing ability.

Much of this work in listening and devising ways of creating acoustical illusions was the outgrowth of the earlier stereophonic work at Bell Telephone Laboratories in 1932.* A three-channel transmission was considered to be the optimum minimum for the best auditory perspective. A larger number of channels yielded to the law of diminishing returns. Two channels gave a stereo effect but was not too definitive. Three channels was the number chosen for their demonstration with the Philadelphia Symphony at that time. Cost was no object for the demonstration.

But cost is certainly important when one is considering a home installation. And two channels are less costly than three. It would be easier in some respects to work with three, but two can be made to perform with satisfaction. Time was, and only a short time ago, when any stereo system of any description in the home was virtually out of the financial question. Another assumption, then, is that the recordist must satisfy a two-channel home system.

But the professional may use any trick he wishes in the preparation of the tape (say) that the customer buys. All he has to do is make it come out "two-channelled". Oh, yes. One more thing: to stay in business, the results must sell.

It might seem obvious, therefore, that the first such recordings were made on two-channel tape recorders. (Figure 1) This is true, but not for the seemingly obvious reason. The first recordings (of recent times) were made for listening via a pair of earphones--one channel for each ear. It was "discovered" that a microphone spacing of about the width of the human head was just about right for this binaural method of listening. When the mikes were spaced too far apart one heard two distinct sets of sounds. Actually it seemed when listening with headphones to sound from widely spaced mikes that one's ears were out on the ends of two long sticks!

And a converse error was made at first. Believing that because we had two ears spaced about six inches apart, the mikes under all conditions should therefore be spaced a similar distance apart. Reproduction over loudspeakers rather than headphones was an obvious must. But when the speakers were placed well apart in a listening room and the mikes picked up the sound from points only six inches apart we had the converse of the ears seeming to be out in space. Imagine someone speaking as he walked across the stage from one side to the other past a pair of closely spaced mikes. In the listening room he would seem to leap suddenly from one side of the room to the other.

All sorts of separation dodges were tried. A sheet of hard, or soft material was placed between the mikes. Even a bag of sawdust about the size of the human head was tried.

Finally the law of reciprocity, and the rereading of the Bell Telephone Laboratories report on their 1932 experiments, caught up with our modern experimenters one generation later.

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* Electrical Engineering, Jan. 1934, Vol. 53, No. 1; six articles.

In a two mike system (in this country at least) the mikes were placed in the same relative position in the recording room that one might hope that the speakers would be placed in the listening room, say at the one-third points across the width. Now comes the juggling and the out-guessing. If the mikes are spaced too far apart, there seems to be a "hole" in the middle of the reproduced sound, and we seem to hear two separate monaural sources from our two reproducing speakers. This is especially true if the listener happened to put his speakers in the adjacent corners of his listening room. At first, this seemed to be a good thing. "Look", he'd say, "I've got a new kind of sound reproduction. They call it stereophonic. It requires two separate channels. Listen; you can hear each one!" True, he could hear each separate channel. One had violins; the other had brass. It was showy. But the instruments didn't blend very well as they should have and as they normally do in a concert hall.

While this separation between the two channels is "showy" and helped initially to sell stereophonic sound to the public, because it was different, the average listener soon tires of this abnormal condition. It is much more satisfactory to listen to a well-blended system; that is, a system which does not have this "hole-in-the-middle" effect.

The considerations which overcome this effect in the listening room are substantially twofold. First, the correct spacing of the speakers within the room. And second, the actual reflection of the sound from the speakers by the walls and ceiling, etc., of the room. In the concert hall under live listening conditions we depend a great deal for our listening enjoyment on the reflections from the various boundary surfaces of the hall. Subjectively and psychologically we believe that we are listening directly to the performers. But do you recall the effect of listening to a symphony orchestra performing outside in possibly a poor band shell or perhaps no band shell at all? The effect is entirely different, isn't it? Likewise in the living room we depend on the time-delayed reflections from the room boundaries to aid in creating the desired stereophonic effect, thus enhancing our enjoyment of the music. Advocates of corner speakers for stereophonic reproduction please take note.

Obviously if the speakers are too far apart we do get a separation in the center. Pulling them closer together eliminates this under normal circumstances. Also, if the original recording was made with the microphones spaced too far apart, then the hole actually exists in the middle of the recording. The obvious cure is for the professional recordist to have his microphones properly spaced. Again, it is the professional recordist's responsibility to monitor the results of his microphone setup in a room which approximates the dimensions of a living room, and preferably a room which resembles very closely the acoustics of the average listening room at home.

That this "hole-in-the-middle" actually exists is apparent from the ways in which professional recordists have endeavored to solve the problem. In the United States one solution has been to use a three-channel stereophonic recording system. (Figure 2) As was mentioned previously, a three-channel system was found to be more efficient and definitive in the transmission of stereophonic sound. With a three-channel system we have twice as many time differences of sound arriving at the various microphones, and loudness differences at the same points as we do in a two-channel system. Apparently it is easier for the ear of the listener to integrate this additional information over a three-channel reproducing system. But the professional recordist must translate this three-channel information into a two-channel reproducing system so that his recordings are salable on a commercial basis. One of the ways in which the professional accomplishes this is shown in the figure. The center channel is split and mixed proportionately with the two outside channels. Judgment of how much of this center channel to mix with the outside channels is subjective. In other cases the center channel may be used for entirely different purposes, such as, for example, a soloist, or to enhance some featured portion of the total ensemble. This arrangement and technique results in commercially acceptable recordings.

In England Dr. G. F. Dutton of E. M. I. has developed a different technique. This consists of using two ribbon microphones, one above the other in the same vertical line, but aimed with their maximum responses at a divergent angle toward the two different sides of the stage. The typical figure eight pattern of each microphone then covers the sides as well as the center of the recording ensemble. The two-channel separation comes from the reduced pickup from the side of the mike toward the opposite side of the orchestra. This technique tends to eliminate the "hole-in-the-middle".

John Moseley, of Pye Records, Ltd., also of England, has tried a similar technique using uni-directional microphones mounted one above the other and facing at a divergent angle of from 90 to 120 degrees.

To those of us in this country who are used to listening to the results obtained from widely spaced microphones, this English system, while it does fill in the center very effectively, seems to be more like a single channel pickup reproduced over two spaced speakers, and lacks some of the specific separation one seems to expect in a stereophonic reproduction. It may be considered somewhat like the pickup obtained from our two mikes placed about six inches apart as was mentioned previously. However, the crossed mike technique does yield a definite stereo effect which the side by side arrangement does not. The subjective reaction to the results of these two different methods of recording varies with the individual auditor, as we might expect.

Of course, the best stereophonic reproduction is achieved when the reproduction is in the same room as that in which the original recording was made. Such a situation was experienced a little over a year ago in two separate places. One of them was a two-channel recording and reproducing session before a full audience at the Philadelphia Academy of Music with the Philadelphia Orchestra and Eugene Ormandy. A short time later a three-channel system was used here in San Francisco with Enrique Jorda and the San Francisco Symphony Orchestra in the Opera House. A selection had been previously recorded using a three-channel tape system. The orchestra members came on stage, picked up their instruments and apparently started to play. Part way through the selection the musicians placed their instruments in their laps, yet the music continued. The audience was dumbfounded at the completely life-like reproduction. The playback was done over three theatre type speakers mounted on the stage in the midst of the orchestra.

The optimum type of reproduction will probably never be achieved in the home because of the entirely different physical acoustics present in the listening room as compared with the recording room. Perhaps someday we may discover a less expensive way of providing three or more channels of reproduction at a modest price for a home system which would improve the illusion. Perhaps we may overcome the deficiencies of our present systems and obtain even more reality in sound reproduction. Or we may even discover an entirely new way of transmitting, storing, and reproducing such three-dimensional sound phenomena as we have been discussing here. From such problems are our future achievements derived.

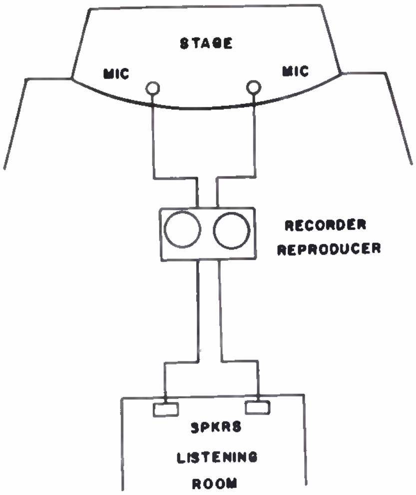

Figure 1

The normal two-channel stereophonic recording and playback system uses two microphones, and a suitable two track recorder. The playback uses an appropriate two track reproducer and two separate loudspeakers in the listening room.

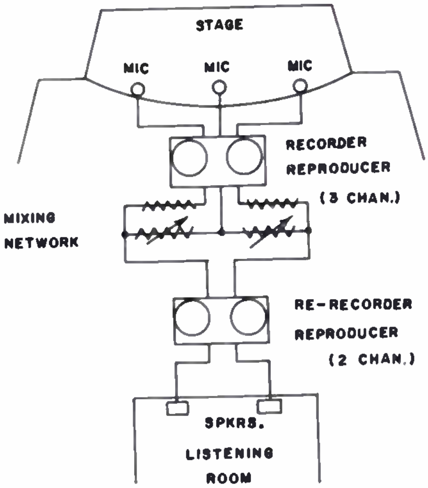

Figure 2

The more commercial stereophonic recording system utilizes three microphones and a three track recorder. The third mike is often used for soloists or special effects. Dubbing of a three track master to a two track commercial release is accomplished through a suitable network. Playback is over a normal two track system.

SECURING 110° SWEEP FOR THE PUBLIC

W. D. Schuster[(1)], E. O. Stone[(2)], and C. E. Torsch[(3)]

Summary

A radically new, compact television receiver has been marketed throughout 1957, containing the first commercial sweep system based on new 110° sweep angle, 1-1/8 inch neck diameter picture tubes. Efficiencies achieved in the subject sweep system permit this set to consume little if any more power than if the set had been designed for bulkier and heavier 90° picture tubes with standard 1-7/16 inch neck diameters.

The new 110° yoke developed for this project is no heavier than 1955 design 90° yokes. The related horizontal sweep output transformer for 110° is smaller than most 90° transformers for 16 kv operation and resorts to cancellation of the usual d.c. flux component to secure 110° sweep system input closely approximating previously commercial 90° sweep system input.

Vertical sweep is particularly economical, and introduces the Sylvania 10DE7 miniature dual triode for oscillator/output combination, operation directly from 250 volts without resort to boost voltage from the horizontal sweep.

Development and adoption of "neck shadow" limit standard yokes for use at the widely separated tube supply, set manufacturing and yoke manufacturing plants was accomplished in 1956 before set manufacture began, with complete success. Nearly identical standards were later adopted by JETEC.

Introduction

For the past five years the transition to sweep angles for television picture tubes in excess of 90° has been imminent. Entry of commercial 110° sweep receivers into public use has been sudden after several oblique approaches were evaluated in various laboratories. The sales appeal of progressively shortened and lightened television receiver cabinetry has been stressed, to cause engineers to design sets featuring greater portability, with picture tubes of the 14 inch and 17 inch-diagonal screen variety. Even greater cabinet length savings are realized with wide-angle sweep on 21 inch and 24 inch picture sets, usually of the console cabinet class, although 21 inch, 110° portable receivers have been built and marketed since June, 1957 by Sylvania.

(1) Television Receiver Design Engineer, Sylvania Electric Prod., Inc., Batavia, N. Y.

(2) Television Picture Tube Application Engineer, Sylvania Elec. Prod., Inc., Seneca Falls, N.Y.

(3) Chief Television Engineer, The Rola Co., Inc., div. of The Muter Co., Cleveland, Ohio

Following extensive engineering development, commercial 17 inch, 110° television receivers have been manufactured continuously since the latter part of 1956, by Sylvania, incorporating for the first time in quantity production, the new 110° picture tubes and a related system of sweep components developed for high efficiengy and economy in mass production.

The overall characteristics of this new set, basic to the Sylvania "Slim Jim" receiver series, have been described elsewhere[(4)] by Mr. Derek Swaine, Supervising Design Engineer and W. D. Shuster. The objective of the present paper is to chronicle the planning of the sweep system and the picture tube application work involved in securing a practical design foundation without which the complete commercial receiver could not have been erected in proper balance.

Picture tube - Yoke Ecology

The evolutionary priority of the egg versus the chicken is a familiar problem confronting television systems engineers at each new frontier of sweep angle increase. In the past, known contours of picture tubes and tube bases have existed considerably in advance of an intended set production date; yoke development followed gradually. Frequently, a series of developmental sweep tubes and sweep components is offered to set designers simultaneously with the new picture tube, to allow evaluations of size, weight and power consumption for a selected picture size and brightness.

In this instance, the picture tube contours were undergoing minor final adjustments at the time the first complete receivers were leaving the production line. The fact that the original yoke design was flexible enough to accomodate to the picture tube adjustments, deserves, it is felt, some note. These mutual readjustments of the picture tube, receiver circuitry and sweep components, particularly the yoke contours, without sacrifice of of the early efficiency objectives, required considerable inter-plant visitation, fluent communications and steady accomodation to new factors and conditions.

That coordination had to cross company and state as well as departmental boundaries, is noteworthy, in view of the emerging equipment as first produced passing the mechanical and "shadow limits" adopted by JETEC at a much later date.

(4) D. Swaine and W. D. Schuster: "A Portable with the 110° CRT", Radio and Television News, May, 1957

Preliminary Stages

Advanced development groups had, early in 1955 investigated the possibilities of a 120° included sweep angle for picture tubes, still retaining the previously standard 1-7/16" nominal neck diameter. It was apparent that equipment sale was even then limited by sweep energy costs for 90° sweep on the 1-7/16" neck size. To achieve the same picture size, at 120° (at a given anode voltage), a high cost, heavy core 90° yoke of more than usual efficiency was found to consume excessive energy for the new trace, even before shortening core and coils to reduce neck shadow effects. This special 90° yoke in question weighed nearly 2.5 pounds, relative to the conventional one pound 90° yoke familiar to most manufacturers during 1955 - 1956. This obvious cost increase in yoke material without approximating commercial 90° yoke intake for 120° tube screen coverage discouraged further efforts on large neck tubes in conjunction with wider sweep angles.

It had been considered, since 1952, desirable to concentrate small neck picture tube developments then in progress, on a 1-1/8 inch nominal neck size. A small-neck 90° yoke was developed (5) and also a small-neck 120° yoke predicated on a 2.828 inches external flare radius on developmental picture tubes. The volume of space within these two yoke cores was sufficiently smaller in diameter and axially longer than conventional 90° yokes that gains were achieved in sweep efficiency up to a point where the 120° tube application was still slightly limited by neck shadow fringes on the screen corners.

At this juncture, during the fall of 1955, two other flare contours were proposed, and enthusiasm for the extreme change to 120° waned. The sweep angle objective swung to approximately 110° and two camps formed to advocate either a flare radius of 2 inches instead of the previous 2.828 inches or, alternately a parabolic flare based on the equation $y = .615\,x^2 + .570$. These shapes were both proposed to enclose a minimum diameter of waste space around the extreme positions of the cathode ray beam within the picture tube neck, and allow for normal errors in gun configuration, mounting and glass irregularities. Sufficient work was done to indicate that either of these two contours allowed a decrease in yoke material and some energy reduction relative to the original concepts for the 2.828 inches radius proposal. The parabolic contour appeared to allow even less yoke material to consume no increase in energy relative to the 2 inch radius systems for comparable sweep lengths.

In January 1956, the Rola Company representative proposed to JETEC consideration of the use of a limit gauge on tube flares of the parabolic variety, defined by an equation $y = .615\,x^2 + .584$. This departure from the nominal glass mold contour was sufficient, that, even though this precise gauge equation was not adopted, use of the shape permitted yoke designs to proceed and coil assemblies to fit the bulb flares. Gauges of the initially proposed shape still fit current production yokes of several competitive designs extremely well. Ultimately, JETEC Standardization adopted the 110° sweep angle, a parabolic limit gauge $y = .58\,x^2 + .576$ above a 1.242 inch neck diameter, to the height where the gauge internal diameter is 4.25 inches.

While the tube and its gauges were being finalized, general objectives were being formed by Sylvania based on preliminary yokes and flybacks provided by Rola. In 1956 model Sylvania 90° chassis, 250 volt B supply enabled these production sets to fully sweep the 110° tubes with little if any power demand increase, using the standard tube complement, together with the developmental 110° sweep components. With this system performance in view, Sylvania formulated the following objectives:

1. Concentration on the parabolic-flare 110° picture tube.
2. Completion of yoke development to match the contour of the 110° parabolic tube alone, "neck shadow" clearance to be only slightly inferior to 90° experience in adjustment ease.
3. Develop the yoke to fit commercially producible ferrite cores tooled early in 1955 by three major suppliers.
4. Minimize yoke material and maximize magnetic efficiency, particularly in the horizontal windings.
5. Reduce flyback material and relative to commercial 90° flyback design as proposed through improved circuitry.
6. Operate the 110° picture tubes at approximately 16 kv at cutoff, to ensure adequate safety factors for electrostatic-focus guns under extremes of operating voltages.
7. Drive the flyback transformer with a 6DQ6A tube operated within all ratings.
8. Use the 6AX4 type damper tube within all ratings.
9. Improve vertical sweep system efficiency to require low input to the matching transformer.
10. Secure an improved miniature bulb dual triode for economical vertical oscillator and output use, later released as the Sylvania 10DE7.
11. Retain the 250 volt B supply concept to permit use of selenium-stack rectifiers in a compact voltage doubler system for maximum portability.

(5) "High Efficiency Yoke for Small Neck Tube" -C.E. Torsch, Tele-Tech and Electronic Industries -July 1955

12. Accomodate low power line voltage input with full sweep at proper aspect ratio.

13. Quickly establish "standard" yokes to define limit tube and yokes within Sylvania and its yoke suppliers.

14. Above all, design for ease of manufacture and assembly, with a sweep system cost objective approximating or bettering the existing 90° system costs and picture performance, including driver tube comparisons.

Sweep System Concept

It was immediately recognized that all known means of aiding magnetic efficiency in the yoke flyback and vertical output transformer were urgently needed in the new set, to minimize weight, size and realize low power supply drain and system cost.

If a single item of the chain were to be neglected, the economy of the other elements would be neutralized.

Realization that a thorough understanding of the picture tube geometry and gun orientation problems was essential was clear to all groups involved. The ultimate in system efficiency can be obtained only by establishing and cross-relating standard "shadow-limit" yokes which accurately define a mutual boundary between production yokes and glassware and gun-mount variations in tube manufacture. The satisfactory experience with 70° and 90° tube standardization (6), even after the manufacture of over 50 million tubes, led to adoption of planning to locate and define the limit tube and yoke combination as rapidly as the tube industry could be seen to converge on dimensional standards. Duplicates of this limit combination were to be maintained at both the tube and receiver factories for instant reference.

Details of the compact yoke and core design, the concept of flyback miniaturization and improved coupling through "desaturation" of the core, vertical output tube and coupling transformer development, picture tube gun orientation and "neck shadow" control are treated in groups hereafter.

Compact 110° Yoke

The accompanying sectional view cut across the screen diagonal of the new picture tube illustrates the shapes and relative size of the 21 inch bulbs of the conventional neck 90° tube superimposed on a 1-1/8 inch neck 110° tube (see figure 1).

(6) "Yoke Development for Standardization of 70° and 90° Deflection Angle" - C. E. Torsch - I R E Transactions on Broadcast and Television Receivers Volume BTR-1, Number 4, October 1955.

Figure 2 is a restricted view, flare only, on the 110° version with a hatched area indicating the flare of the JETEC 110° limit gauge. Superimposed on the latter is a sectional view of the ferrite core employed in this yoke design.

The JETEC #126 limit gauge ultimately adopted defines the maximum space displacement and contour allowed for glass and provides an index for reference of the distance of this area from the screen, which establishes external maxima and minima for tube dimensions. The tube base and straight-neck region is allowed to rise to 1.168 inches diameter. A short region of 4° taper in the gauge over the neck to funnel glass splice ends at 1.242 inches diameter, followed by a parabolic section described by the equation $y = .58 x^2 + .576$ where $x = 0$ locates a plane .92 inch closer to the tube base than the "reference line" of the tube. The "y" ordinate values, in inches, describe the internal radius of the gauge from an axis of revolution. This gauge is limited to an "exit diameter" of 4.25 inches.

This yoke core configuration was developed early in 1955 for small-neck tube yokes, through the cooperation of the Allen-Bradley Company, Stackpole Carbon Company and the General Ceramics Corporation, with yoke design planning of The Rola Company. Dimensional standards of the Metal Powder Association were introduced and conventional molding considerations were allowed for in providing a flexibility in a choice of length of core. All three ferrite sources tooled this core shape as a quadrant with "ship-lap" edges (for reduced magnetic leakage). No restriction was placed by Rola on the distribution of the molded parts to encourage the early evaluation of small-neck picture tube sweep systems in general. It is believed that this is the first time that such a widespread tooling of a yoke core design has been unrestricted in distribution.

The flat back of the core requires more ferrite to be employed than strictly necessary, in exchange for conventional, safe tooling, as known in 1955, to guarantee steady availability. This shape was considered to give the minimum practical inside diameter at this state of the art of yoke design, 1.6 inches diameter being an MPA preferred dimension step, allowing an annulus of space at least .208 (7) thick for coils and intervening insulation. With this inside diameter, a tangent, smooth radius of .860 inch was judged to produce a good degree of parallel to the bulb gauge surface. The outer diameter was selected to cut the core thickness to .4 inch, where the front core radius was producing relatively little benefit, in further containment of the yoke field. Subsequent core designs have increased the exit flare diameter from the 2.4 inch value for the design described herein, to over 2.8 inches, however this increases core cost and reduces flexibility in the choice of length. The original subject core can

(7) "Trends in Ferrite Core Design for T-V Yokes and Flybacks" - C. E. Torsch - Proceedings Eleventh Annual Meeting Metal Powder Association - Volume II, May 1955.

be made from the same tool to a maximum length of approximately 1-5/8 inches to below 1 inch, in an infinite variety of possible increments. At this time the subject yoke contains a core 1.3 inches long to secure optimum sensitivity vs. beam clearance.

The coils are nested between this core and the flare gauge diameter in a snug sandwich, the upper layer containing the vertical windings, the lower layer next to the glass, the horizontal coils. The sandwich "meat" is formed by a molded sheet of vinyl, containing a central rib to separate the horizontal coil edges for half of the series-connection assembly voltage.

Both horizontal andvvertical coils have an approximately cosine distribution of thickness to match a constant thickness insulator for a mixture less than the maximum thickness indicated previously.

While a variety of impedances can be wound in the available space, it was found desirable to employ a series connection of horizontal coils to a total inductance of 30 millihenries in the horizontal windings, and 32 - 36 ohms in the total of the two series connected vertical windings. One horizontal coil is reversely wound relative to the opposite coil in series with it to secure a raster more free from velocity modulation. (8)

In addition to this reverse winding of the horizontal coil half, both vertical coils are wound in reverse to secure a.c. "ground" at the inner or start turn of the left hand coil, facing the yoke flare. This assists the elimination of vertical-position modulation at the left edge of horizontal lines by draining off electrostatic current from the high-potential edge of the horizontal coil on the coil start nearest to the beam position early in horizontal tract times.(9) The vertical windings of this wire size (#28) do not greatly limit the Q of the horizontal magnetic circuit, as would an extremely low resistance o coil using #22 wire for example, due to eddy current loss at the horizontal sweep rate in the vertical coil conductors in the path of horizontal flux.

A choice of series-connected coils for both windings was based on greater productivity of yokes from the new winding machinery. This choice was fortunate in view of the early demand for large quantities of yokes.

(8) "High Efficiency Low-Copper Sweep Yokes with Balanced Transient Response", C. E. Torsch, I R E Transactions on Broadcast and Television Receivers, PGBTR-6, April 1954.
(9) "Extention of the Balanced-Transient Response Principle to Color T V Yokes", C. E. Torsch, Transactions of the I R E Broadcast and Television Receivers PGBTR-6, April 1954.
(10) U. S. Patent 2,568,471 -Torsch and J.B. Coullard, also, "High Efficiency 90° Cathode-Ray Sweep System" -Torsch, Tele-Tech and Electronic Industries- June 1953.

Relatively powerful cunife-strip magnets were attached to flexible arms to aid the horizontal sweep sensitivity of the assembly, as well as reduce the "pin cusion" pattern distortion at the right and left hand vertical edges resulting from the effort to emphasize focus quality of the deflected spot on the screen.

The yoke is encased in a Tenite case of flame-retardant acetate, of high temperature stability. The case is made to clasp, without resort to rivets or eyelets, a fibre terminal board at the front face, keeping the usual connections away from the area required for access to the magnet bars and centering magnet and clasp Coloring the case fluorescent red warns the adjuster against grasping the yoke beyond its protective case flange at the front, where the terminals are located. A post and plastic ridge on the case interior, index the core segments and vertical coils against rotation in the assembly. A plastic apron on each case half at the rear protects the assembled coils from handling damage when the assembly is slid along conveyors or tables, and during shipment.

A detachable plastic cap with dual disc centering rings is mounted at the rear of the yoke. To this is mounted a plastic funnel which is clamped to thè picture tube neck by means of a sturdy spring of the type used on auto radiator hose fittings. The plastic funnel is coated internally with a friction-paint to reduce the risk of shift on the glass neck during shipment. Ring cores of ferrite have been made for this yoke, cracked in halves for assembly ease. Very little performance is to be gained relative to the ship-lapped quadrants due to the extreme thickness of the core in this design, tests of the cracked ring show.

The spring-steel strap which ties the case and core together is magnetized after assembly to pr-duce a weak field causing a shift of the entire raster to the right to compensate for a combination of earth's field (in North America), tilt limitations.

The entire yoke assembly weighs approximately one pound - no more than usual 90° yokes for 16 KV operation.

Flyback Transformer "Desaturation"

Once an economic limit is reached in the yoke, the flyback in conjunction with it must be matched and whittled. Two basic factors have already been covered in the literature (10) but have not generally been commercially used together thus far. These two items are: first, consistent mechanical arrangement of coil spacing to secure critical coupling to the high voltage extension winding; second, elimination of the d.c. core flux component inherent in practical flybacks wherein the damper tap is below the driver plate connection to the full primary winding.

arrangement of coil spacing to secure critical coupling to the high voltage extension winding; second, elimination of the d.c. core flux component inherent in practical flybacks wherein the damper tap is below the driver plate connection to the full primary winding.

Figure 3 illustrates the circuitry involved, based on the sectionalizing of the transformer primary; use of a width control inductance (adjustable, in this instance). Use of this coil is as an isolation choke through which the cathode current of the driver tube flows into an isolated section of the primary and out to the damper plate. The reverse flow of d.c. flux in the lower section of the primary neutralizes the d.c. flux set up by the plate current flow from damper cathode to driver plate. The ampere-turn balance is established, then the conventional air gap between core halves is eliminated from the flyback. This raises the shunt impedance of the flyback and permits a reduction of ferrite core cross section and weight reduction of approximately 35% compared to 1956 model 90° flybacks. Further core reductions are practical.

The tuning of tertiary high voltage winding leakage reactance, self inductance and capacitance to secure the "flat-top" primary voltage-pulse wave form during retrace is vital to minimize insulation stresses at the driver tube plate, damper cathode, flyback primary coil, in each yoke horizontal coil and in horizontal to vertical insulation. This 13 per cent reduction in peak voltage of a well tuned wave may actually represent a 20 to 40 per cent reduction in in peak voltage of a well tuned tertiary compared to a poorly tuned unit in which the harmonic frequency, phase and amplitude actually increase the fundamental crest.

Vertical Output System

Considerable effort was applied within the Sylvania organization to secure an improved output circuit and adequate miniature output tube to drive the 110° yoke windings. Figure 4 shows this this circuitry.

A large portion of the work secured an excellent vertical output autotransformer without resort to grain-oriented steel in the laminations, tight coupling, low resistance of windings, with sufficient shunt inductance to keep magnetizing current low relative to load demand. Little attention has been given to this important item in current 90° equipment.

The miniature dual triode output tube, Sylvania 10DE7 was developed to consume an average drain of about 30 milliamperes and operate from 250 volt B supply without resort to boost voltage derived from the horizontal sweep, and allowing for cathode self-bias voltage loss. This tube readily delivers 80 milliamperes peak current for this application, at 21 volts total bias. Only 580 volts peak-to-peak plate swing results at full scan. The oscillator section in the same bulb is a lighter triode and runs from filtered boost voltage but requires only a milliampere average drain. The output and oscillator sections together operate as a multivibrator.

This output tube and transformer combination will have considerably more reserve for 90° standard yoke applications than most economical systems in use. Note the output tube current characteristic in Figure 5.

Use of pentode output tubes would require more sockets and a tube to complete the oscillator, also require a screen supply resistor and by-pass capacitor, before sweep linearity is considered (adverse to most pentode circuits).

"Neck Shadow" Controls for Yokes and Picture Tubes

The following limitation on design of efficient yokes for economical sweep systems is termed "neck shadow". To this is closely related the recently complained of "contrast reduction" in scenes reproduced on picture tube screens when overscan of sides or even corners of the screen is necessary (to refill either the width or height of the screen at correct aspect ratio of reproduced objects.).

The cathode ray beam will always, in covering the full picture tube screen (which seldom approaches a proper 4/3 aspect rectangle), impinge on the inside of the neck to funnel flare region (which is round), resulting in a projected image (or "neck-shadow") of the flare circle on the bulb funnel, on or near the screen corners. The degree with which this always present shadow on the bulb walls actually reaches or encroaches on useful screen area has historically been a touchy subject, shrouded in mystery and suspicion.

This overscan would have been most serious on a full-round screen of the 16GP4, 70° type of picture tubes if longer yokes designed for efficient scan of the rectangular screen 70° had been used on the round screen tube. As most 90° tubes were nearly rectangular in screen shape, less loss of picture area and resultant contrast loss was noted. The overscan of tube neck or bulb walls beyond the screen liberates secondary electron from the aluminum film, aquadag coating or even the glass itself. These secondaries are more abundant where the primary beam is of high current density. The cloud of secondaries scatters over the screen, adding to the total illumination, most markedly where no primary excitation of the screen would leave a dark or dimly lit area.

Even more significant in causing contrast reduction (than secondary electron emission from from walls) is the abundant, higher velocity primary cathode ray beam ricochet from the neck and bulb walls, by elastic collision and grazing reflection from the overscanned areas. This random reflection spray scatters over the entire screen and further decreases the illumination contrast partially degraded by the secondary electron spray mentioned above.

This condition of contrast reduction need not be present if the screen is not overscanned. All presently-designed 110° bulbs contain screens of incorrect aspect ratio (for reproduction of 4/3 aspect transmitted signals) and the screen edges are considerably removed from spherical chords. It has been found that even the shortest field-yokes now available contribute to neck or funnel overscan, even though not to visible shadows on the screen itself.

The reduction or near-elimination of this defect in picture reproduction can be obtained, retaining the presently standardized parabolic neck flare contour, if the aspect ratio of the screen is corrected and the screen edges straightened to minimize overscan at the corners. Bulbs of this revised shape could have larger screen diagonals for the present screen height, but a longer cone would result, to retain 110° sweep to the maximum screen diagonal.

To retain the present screen diagonal dimension and bulb length steps now set for 14, 17 and 21 inches 110° bulbs, the screens would be wider and less high, with small corner fillets and flattened screen edges. It is presently considered unlikely that the contrast loss will be judged worth redesign of glass molds and receiver masks and mountings to correct present overscan conditions. Presently, 21 inches, 110° bulbs are thinned in the neck flare region by "paddling" to enhance contrast.

The objective of lighweight, efficient sweep systems with full utilization of the greatest possible yoke length at a minimum core diameter, is only met by matched controls on the maximum bulb thickness and minimum spacing of the limiting orifice region to the plane of the maximum projected screen diagonal span, to locate the maximum conic angle to be swept in a well aligned sweep yoke and bulb axis.

The yoke design cannot be efficiently finalized, therefore, until orderly definition of the limiting glass geometry is secured. To this end, the bulb glass supply companies have agreed to establish tests of the funnels prior to fact plate sealing.

This test is the measure of penetration of a second gauge (JECTEC #125 plug gauge) into the open bulb funnel, on the outside of which is simultaneously held the external reference line gauge (#126 gauge) discussed previously. A measure of the overlap of the two gauges is an indirect measure of the maximum glass wall thickness separating the gauges.

By choice of the plug gauge surface to approximate the exit trajectory of cathode rays leaving the influence of an ideal, economically designed yoke field, toward the corners of the screen for clearance of the bulb walls under most unfavorable conditions, equal clearance can be predicted on a well-aligned electron gun, yoke field and glass neck axis, if the neck axis is the perpendicular bisector of the maximum screen plane. The plug gauge ultimately established is a conic surface described by the parabolic function $y = .49 x^2 + .420$ from a tip diameter of .928 inch to a diameter of 3.26 inches. Prior to the face plate seal, this plug tip must, in all satisfactory bulbs, penetrate the #126 gauge at least to within .3 inch of the plane of the y axis, or at least .62 inch beyond the bulb "reference line" plane.

With the external limit dimensions of the bulb established and readily recheckable at any time, it is desirable to allow tube manufacturers and set manufacturers to repeat the glass companies' initial pre-seal mechanical measurements for limit glass-ware, by the only simple expedient presently visualized, by electronic means. The construction of a special sweep yoke must be undertaken, even before final yoke design concepts are complete, to produce a magnetic field in which the corner cathode rays (at 5 microamperes, 16 kv) will just skim the walls of the limit sweep angle bulb, containing the mixture of adverse dimensions defined by JETEC. The symmetrical formation of the neck shadow on a limit tube must just clear the screen corners as viewed from directly in front of the screen parallel with the neck axis, with the stipulated low beam current.

Selection of limit glassware was expedited by cooperative efforts of the major glass suppliers, Corning Glass, and Lancaster Lens, and processing many of these bulbs by Sylvania at Seneca Falls. Rola Engineering developed an encased yoke, at the invitation of JETEC subcommittee 6.9, during November 1956. This special yoke assembly was checked on the series of limit 17 inch bulbs by then completed into picture tubes by Sylvania. The most severe mixture of dimensions encountered on the series was established as a "limit tube" on which the special yoke operation on the 5 microampere beam current required careful beam centering and ion trap adjustment to secure complete avoidance of symmetrical diameter shadow in all screen corners.

By use of this yoke which determines the effective "limit tube", tubes can be classified as within or beyond practical operation limit. The rejection of completed tubes focuses attention upon marginal or reject glassware before any appreciable quantity can reach the receiver manufacturer.

Conversely, the commercial yokes to be used with the same tube series should exhibit less neck shadow than the standard yoke on any tube.

The original special yoke was dispatched for tests by each of the other tube manufacturers represented on the standardization subcommittee, JETEC 6.9, with other bulbs in the other factories. During December, 1955 and January, 1956, six different companies had observed the operation of this proposed standard yoke for shadow limitation.

Meanwhile, three other yoke assemblies were built to match the first unit in performance, and adjusted by comparison with it to give the same shadow characteristic. One of these was established as the Sylvania Tube Department's reference standard in subsequent relationship of shipment of picture tubes to the Sylvania Receiver plant at Batavia, N. Y. The second yoke of the trio of secondary standards was brought to the Sylvania Receiver plant, together with a carefully selected "twin" to the "limit tube" established at the Sylvania Tube plant. Rola retained the last of this trio of secondary standards.

The first effect of this arrangement was to locate the condition till then undetected, namely, that the 17 inch bulbs were not entirely matched in mechanical proportion to the 14 and 21 inch types of 110° bulbs. Readjustment of this was accomplished by the glass companies during December 1955 and agreed to as final size-limit recommended standards for maintainence of interchangeability of glassware by JETEC subcommittee 6.4 during January 1956.

Final adjustment of the commercial design of the yoke coil and core length was then possible during December 1955, pending the readjustment of 17 inch bulbs to match the 14 and 21 inch glassware proportions in the 110° series, using the tentative standard reference yokes provided. It was decided, in view of the hazard that bulb readjustment might not become a reality in time, to maintain the three reference yokes as controls between the Sylvania plants and between the receiver plant and the yoke manufacturing plant to bridge the 330 mile geographical span between these related operations.

As a final contribution of the close tie of this development review, evaluation of picture tube gun orientation was expedited. It was found to be advantageous to rotate the ion-trap type gun in the 17 inch tubes to a position in which at optimum trap magnet flux and position, the necessary net beam drift was directed slightly to the left of screen center, in a magnetic field-free test area. The earth's magnetic field vertical flux component in North America tends to deflect the beam from a gun in any horizontal plane, toward the left. In opposition to this, some fixed raster shift toward the right is needed to compensate the horizontal sweep non-linearity due to screen geometry. This non-linearity is in part due to the trace velocity distortion inherent in cathode ray beam motion at nearly constant angular velocity across any spherical screen, the radius of curvature of which is not concentric with the apparent center of beam deflection in the yoke (not a fixed center for all portions of the screen). Correction for this common distortion is possible by current wave modification in the yoke, departing from a sawtooth during trace. (11)

(11) U. S Patent 2,510,027 - C.E. Torsch

It is common to secure this waveform by connection of a relatively high reactance capacitor, .033 microfarad, for example, in series with the horizontal yoke reactance directly across the flyback transformer secondary terminals. This stretches center sweep-trace but tends also to compress the right edge and slightly stretches the left, on 110° tubes, more than noted on the 90° bulb series, which were relatively more bulbous <u>in proportion to the sweep angle.</u> This does <u>not</u> mean that the unconsidered appearance of the 110° bulb series, externally, is flatter than the 90° series; the opposite is obvious. Indeed, if a less bulbous screen had been adopted on the 110° series, as was initially planned for 21 inch 110° tubes originally sealed with 90° face plates, more serious sweep distortions and deterioration of beam focus would now be noted. This is one serious present limitation of wider angle sweep than the 110° level selected.

With the beam position shifts noted, it was found desirable to continue the practice common in some areas on 90° yokes, namely, to magnetize the core retaining strap after yoke assembly, to bring the position of the undeflected spot nearer to screen center. The farther forward toward the picture tube screen this strap is placed the more easily the stray flux from the strap accomplishes this result, to minimize "neck shadow" formation on the screen.

Figure 6 compares the conventional 90° yoke and flyback with the subject 110° yoke and flyback in size and shape; together with a 21 inch 110° picture tube, and a 110° yoke plug gauge adjacent to an equivalent section of the glass flare of a standard 90° picture tube.

Conclusion

In meeting the established objectives with a steady flow of picture tubes, yokes and pioneer chassis into a matched combination which appears to have gained public acceptance, without the usual "neck shadow" epidemics recalled in 50°, 70° and 90° system introduction, a milestone of orderly progress seems to have been reached.

The authors wish to acknowledge the cooperation of the other JETEC - represented tube manufacturing organizations and the related glass suppliers named above. The achievement of the goal of commercialization of 110° sweep from a laboratory concept (considered uneconomic a few months before) is largely due to the vision of Mr. Robert Thalner, Chief Engineer of the Sylvania Receiver plant, Mr. Ross Gessford, Chief Engineer of the Sylvania Picture Tube plant, Mr. William Dickinson, Section Head, Sylvania Picture Tube Design Engineering, and Mr. A. W. Keen, Manager of Sylvania Picture Tube Application and Field Engineering.

Execution of the complete receiver design under the direction of Mr. Derek Swaine has had, in the sweep transformer finalization, further assistance from Mr. F. T. Henry and others of the

Sylvania Coil plant at Williamsport, Pennsylvania.

At The Rola Company, the able assistance of Mr. J. J. Brincka in flyback engineering and of Mr. N. T. Grasso in yoke engineering is gratefully acknowledged.

These facilities were effective only through fluent intercommunication by telephone, airplane, rail and auto, with a joint spirit of cooperation to meet a common goal, the reality of which was accepted with a large measure of faith by the managements and manufacturing organizations involved.

The degree of success with which this new sweep system has met in the public domain seems to have stimulated other firms to accelerate their planning to convert to 110° equipment to the point where it is predicted that a major portion of the T. V. manufacturing will be at the wide angle during 1958, particularly in America.

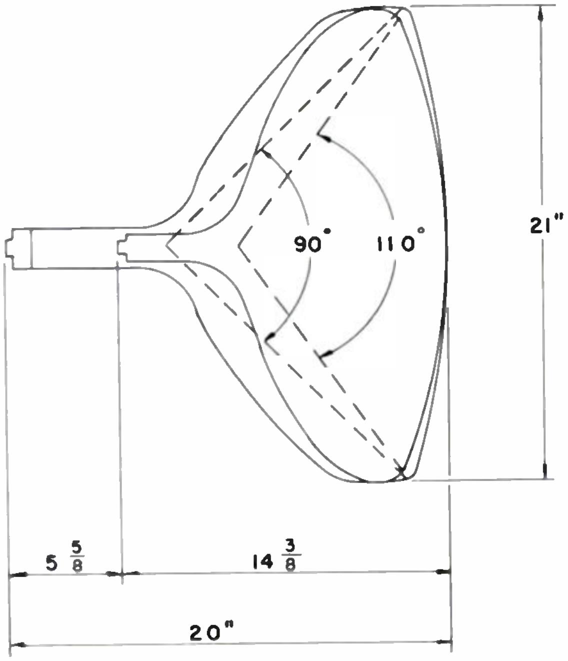

FIG. 1

COMPARISON OF 21" PICTURE TUBES (DIAGONAL SECTION), ORIGINAL STANDARD NECK 90° VS. SHORTER, SMALL-NECK 110° BULB.

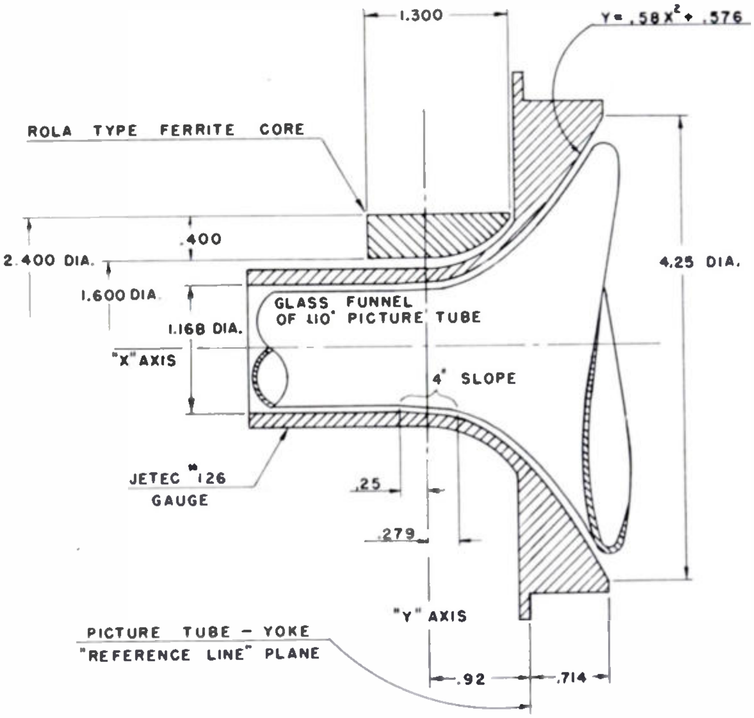

FIG. 2

SECTIONAL VIEW OF JETEC STANDARDIZED (110°-NECK) EXTERNAL CONTOUR GAUGE #126 ON TYPICAL GLASS FUNNEL OF PICTURE TUBE, SHOWING RELATION TO ROLA-TYPE FERRITE CORE SECTION.

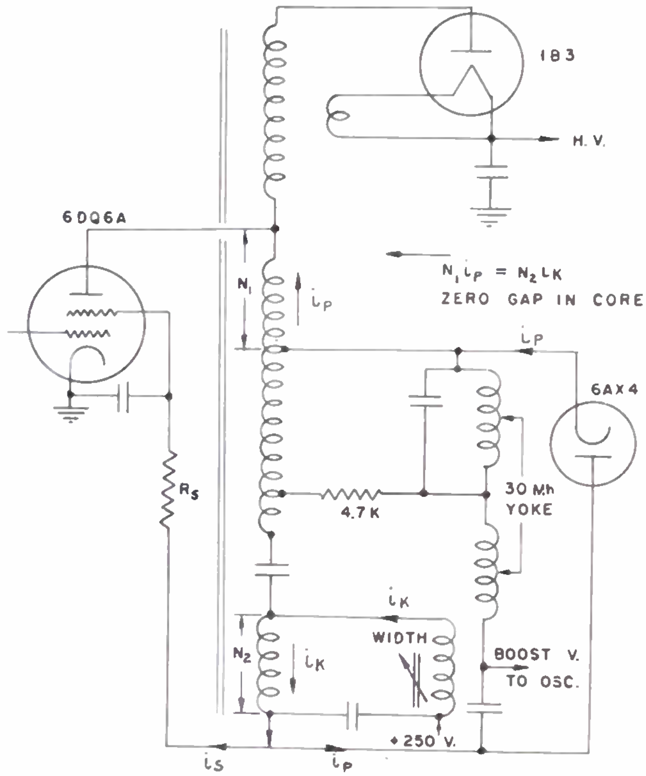

FIG. 3

110° HORIZONTAL SWEEP OUTPUT/HIGH VOLTAGE SYSTEM, SHOWING "DESATURATION" SPLIT OF FLYBACK SECONDARY FOR REVERSED D.C. FLOW, THROUGH WIDTH COIL ACTING AS VOLTAGE SOURCE - IMPEDANCE ISOLATOR.

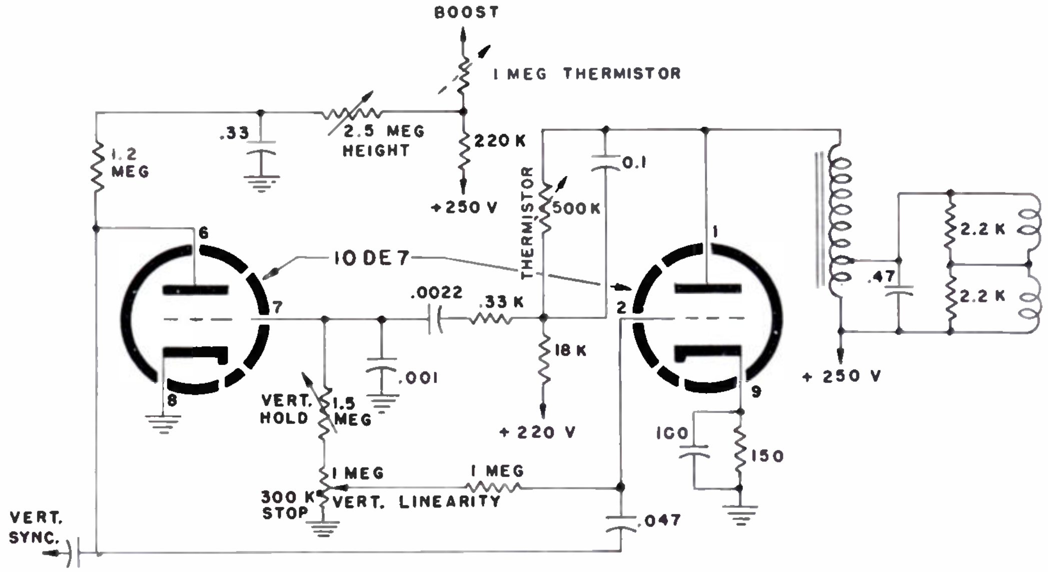

FIG. 4

VERTICAL OSCILLATOR/OUTPUT STAGE OF SYLVANIA 110° RECEIVER SHOWING THERMISTOR COMPENSATION.

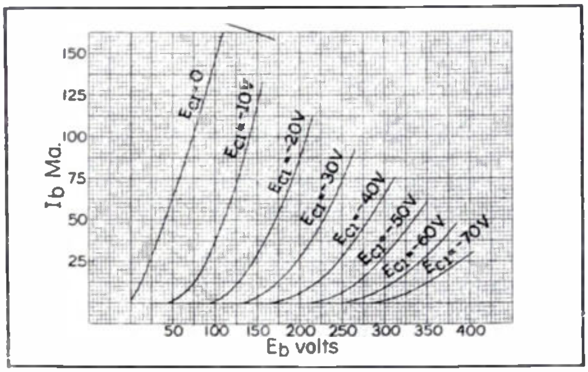

FIG. 5 Plate family characteristic curves for the Type 10DE7

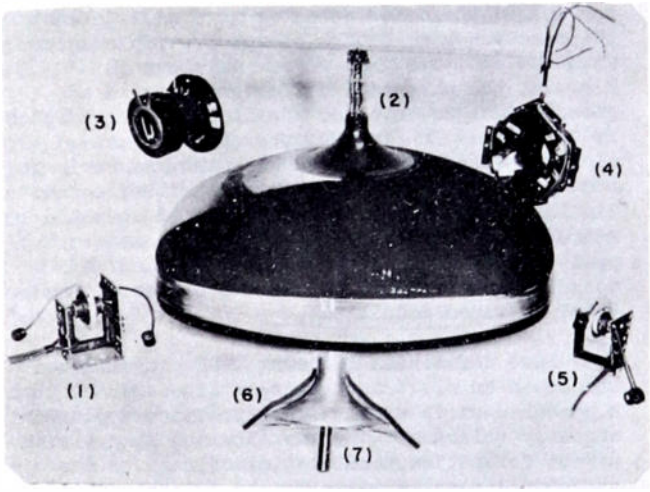

FIG. 6

(1) STANDARD 90° FLYBACK, UNDESATURATED CORE.

(2) 21" 110° PICTURE TUBE SUPPORTING BALANCE BETWEEN STANDARD 90° YOKE (3) AND ROLA 110° YOKE (4).

(5) DESATURATED CORE, 110° FLYBACK.

(6) CUTAWAY OF 90° GLASS NECK FUNNEL SUPPORTED BY (7) METAL PLUG REPRESENTING 110° GLASS NECK FLARE OF PICTURE TUBES.

A BRIGHTNESS ENHANCED COLOR RECEIVER EMPLOYING AUTOMATIC DECODING IN THE CHROMATRON

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Summary

This paper describes a prototype color receiver utilizing a unique method of decoding the NTSC transmitted color television signal.

The new decoding method, termed "quadramatic", takes advantage of the inherent capability of the CHROMATRON[1], or Lawrence single-gun color tube, to produce a color picture of high brightness. Basically a gate-off pulse is used with quadramatic decoding instead of using a gate-on pulse, with attendant space-charge blow-up, as in former self decoding methods. Highlight brightness of 40-60 foot-lamberts with good resolution has been obtained with a 22" rectangular CHROMATRON tube.

CHROMATRON Operation

The single electron gun tricolor CHROMATRON employs sets of colored phosphor strips for the light emitting screen, (Fig. 1). A color selector grid of alternately connected wires is arranged parallel to the phosphor strips.

A potential difference of proper magnitude applied to the color selector grid will deflect the electron beam to the phosphor strip aligned with the positive grid wire. As shown in Fig. 2a, the potential difference between adjacent wires is zero and the blue center phosphor is selected. In Fig. 2b, the set of wires electron optically aligned with the red strip is made positive and the red phosphor is selected. Similarly the green strip is selected when the set of wires over the green strips is made positive. Color selection is independent of scanning pattern, orientation, position of the beam, or video modulation.

Post deflection focusing, PDF, is utilized to obtain an electron beam spot size smaller than a phosphor strip width. This provides ample tube assembly tolerance. PDF is obtained by applying a post deflection acceleration voltage to the phosphor screen relative to the switching grids; in this case 18 KV, (Fig. 2).

NTSC Color Signal

The NTSC color signal consists of a high-resolution luminance component for black-and-white compatibility and a limited resolution two-phase quadrature-modulated chrominance component, (Fig. 3).

For color signal components with frequencies of less than 0.5 mc, the I and Q components are both present and the signal in terms of the color difference components is:

$$E_m = E_Y + 1/1.14 \left[(E_R - E_Y) \cos\omega t + 1/1.78 (E_B - E_Y) \sin\omega t\right]$$

The chrominance difference signals $(E_R - E_Y)$ and $(E_B - E_Y)$ are reduced in amplitude to 87.7 percent and 49.3 percent, respectively, to prevent excessive over-modulation of the transmitter. The luminance signal is not reduced in amplitude, otherwise compatibility would be affected.

Quadramatic Decoding

The one-gun CHROMATRON has a color-selector-grid interelectrode capacitance of about 3000 uuf. While square-wave switching can be used for low-frequency color-switching applications, an external inductance resonated with the grid capacity is used to provide the 3.58 mc sine-wave switching employed in quadramatic decoding. The effective Q of the resonant switching circuit is high so that a power saving of about 100 is effected by using real power rather than reactive power.

The sine-wave switching causes the beam to oscillate back and forth over the red, blue, and green phosphor strips as raster scanning action takes place. When this frequency is chosen equal to the NTSC subcarrier frequency, a low-visibility interlaced dot pattern results.

The highlight brightness of the one-gun CHROMATRON can be increased substantially by increasing the gating duty factor of the electron beam to unity. The quadramatic decoding technique approaches this criterion more closely than other decoding techniques previously utilized. The NTSC signal phase sequence, (ref. Fig. 3) is red, blue, green; and repeats at a 3.58 mc rate. Decoding of the NTSC signal in the one-gun CHROMATRON is accomplished by choosing 3.58 mc as the color switching frequency and applying the NTSC video signal to the beam intensity control grid. The R-Y and B-Y vectors are separated in phase by 90°, and G-Y and B-Y vectors are separated by about 124°. If the color difference vectors are superimposed on a diagram showing the phosphor strip sequence for a CHROMATRON with blue as the double resolution color, it is observed that the NTSC color signal difference vectors and a double blue CHROMATRON phosphor strip sequence match very closely, (ref. Fig. 4).

It will be noticed from Fig. 4 that any yellow video information present in the transmitted scene occurs at the time the electron beam is crossing the blue-emitting phosphor strip for the second time; therefore, the electron beam is gated "off" for approximately 36° during the second blue crossing to eliminate the unwanted blue during a yellow hue presentation. This gate off pulse is known as the "anti-blue" or "knockout" pulse and compared to other decoding techniques permits approximately 90% (i.e., 100 - [36/360 x 100%]) utilization of the electron beam.

Former decoding methods utilized a "gate-on" pulse to turn the electron beam on when the proper phosphor strip was being crossed. The low duty cycle imposed by color purity considerations required a high peak-to-average ratio of current pulses during the "gate-on" time for a bright display. The resultant space charge dilation of the beam imposed a limit on the resolution capabilities in highlight areas.

Using quadramatic decoding, a highlight brightness of 40-60 foot-lamberts is easily obtained from a rectangular 22-inch CHROMATRON with a 25 KV viewing screen potential.

This method of enhancing brightness by using an increased duty cycle may be compared with a decoding method employing chroma modulated third harmonic gating which was presented at the 1957 National IRE Convention[2].

Prototype Receiver

The circuit of a quadramatic decoding receiver is basically simple and requires fewer tubes than commercially available color receivers.

A developmental quadramatic decoding receiver using 25 tubes was built and demonstrated. A block diagram of this receiver is shown in Fig. 5. It is similar to earlier one-gun CHROMATRON receivers described in the literature with the exception of the decoding method.[3]

Figure 6, the anti-blue, or knockout pulser consists of a high perveance tube in a peaking amplifier circuit. Driving voltage for the pulser is derived from the color selection amplifier. Pulse phasing is adjusted in the pulse generator grid circuit. A diode clips the output pulse backswing. Pulse amplitude at the CHROMATRON cathode is 120 V; pulse width is 70 millimicroseconds. The peaked output pulse contains harmonic frequencies of the 3.58 mc subcarrier, therefore the anti-blue pulser is carefully shielded and all power supply lines to the anti-blue chassis are filtered. The 120 V amplitude pulse is sufficient to gate off the CHROMATRON electron beam for all contrast levels.

The higher brightness in quadramatic decoding displays increases viewing screen current requirements. Shunt regulation of both color selector grid and viewing screen potential is used in the quadramatic high voltage power supply. A corona discharge regulator tube has been used with marked success to stabilize the 25 kilovolt viewing screen potential.

Hue stabilization during warmup is achieved by referencing the subcarrier regenerator from the color selection grid signal rather than the crystal oscillator.

Some errors in chromaticity and brightness exist in quadramatic decoding, as they do in other forms of direct decoding. However, certain normalization functions can be performed upon the NTSC color signal to readjust the color difference coefficients to unity. Complete normalization results if the ratio of chrominance to luminance is set to 1.14 and if the B-Y axis is multiplied by 1.78, (Fig. 3).

The chrominance to luminance ratio can be normalized by increasing "chromaboost". Chromaboost is a coined term applied to amplitude control of the 3.58 mc chrominance video relative to luminance video.

A satisfactory method of accomplishing color hue normalization is to apply a second harmonic of subcarrier frequency to a control grid of the first video amplifier stage. The second harmonic modulating frequency amplitude is chosen so that gain of the first video amplifier stage varies by a factor of 2 in the linear operating range of the tube. A polar representation of amplifier gain shows the locus of the gain vector to be an ellipse; hence, the term elliptical-gain amplifier.

Subjective evaluation of quadramatic decoding displays reveals that rigorous NTSC signal normalizing is not required if phosphor luminous efficiencies and strip widths are properly chosen. In effect, normalization of the viewed color picture is accomplished mechanically within the CHROMATRON. For this case only the conventional color saturation or chromaboost circuit is necessary.

The subject of colorimetric errors resulting from direct application of the NTSC signal to the intensity control grid of a one-gun continuous color sequence display has appeared in the literature.[2,4] These prior graphic presentations utilizing the CIE diagram indicate satisfactory reproduction of critical colors.

Another method of qualitatively treating the subject of reproduced color error is by analysis of beam current distribution on the phosphor area. Beam current on each phosphor strip is plotted for a given transmitted color. For a perfectly accurate reproducing system the total beam current should excite the desired phosphor strip. Fig. 7

indicates the magnitude of color contamination present in red due to excitation of the undesired phosphor strips. (See appendix.) The green phosphor is not excited at all while the blue phosphor excitation comprises less than 2% of the total illumination. Blue has the most contamination of the pure primary colors (Fig. 8). Of the total beam current, 30% flows to the green phosphor strip and 6% flows to the red phosphor strip. However, the blue phosphor used has three times the luminous efficiency of the red or green phosphor. Thus, the effective green contamination is 16% and the red 3% when reproducing blue without hue normalization. Figures 9 and 10 show the beam current waveforms for green and yellow. Figure 11 is the rear view and Figure 12 is the side view of a prototype quadramatic color receiver.

In conclusion, the quadramatic decoding technique offers a high brightness color display with a minimum of circuit complexity and alignment adjustments.

Acknowledgment

The principle of quadramatic decoding is a contribution of Professor E. O. Lawrence. A large portion of the material in this paper has been derived from a Litton Electronic Display Laboratory* internal report (EDL 103) prepared by R. Door, J. M. Rosenberg and the author. Credit is due all members of the Litton Electronic Display Laboratory staff, especially G. Morrison and R. Wong for their work on the prototype receivers.

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*Formerly West Cost Development Laboratory, Chromatic Television Laboratories, Inc.

Appendix

Calculations of beam current waveforms. (Figures 7, 8, 9, 10)

Given:

$$E_m = E_Y + .493\,(E_B - E_Y)\,\text{Sin}\,wt + .877\,(E_R - E_Y)\,\text{Cos}\,wt \quad (1)$$

$$E_Y = .30\,E_R + .59\,E_G + .11\,E_B \quad (2)$$

Where E_R, E_G, and E_B are signals derived from the red, green, and blue signal outputs of the camera.

Assumed Conditions:

a. All the beam current is concentrated in an infinitely small spot.
b. Chrominance to luminance ratio set to 1.37.
c. System up to the picture tube is linear.
d. Transfer function of the picture tube is proportional to the 2.2 power.
e. CHROMATRON phosphors balanced for equal energy white with sine-wave color switching.
f. The effect of hue normalization techniques is not included.

Sample Calculation for a Transmitted Saturated Red at an Angle of 13.4°.

a. Let $E_R = 1$; $E_G = 0$; $E_B = 0$

b. Then substituting these values in equation (2) $E_y = .30$.

c. Multiplying chrominance by 1.37 and substituting 13.4° for wt:

$$E_m = .30 + .493 \times 1.37\,|-.3|\,\text{Sin}\,13.4^\circ + .877 \times 1.37\,|.7|\,\text{Cos}\,13.4^\circ$$

$$E_m = 1.163$$

d. Transfer function of the picture tube is assumed to be 2.2,

$$I = kE_g^{\,2.2} \;;\; I = k\,(1.163)^{2.2}$$

$$I = k\,1.394$$

The calculations are similar for any other desired color.

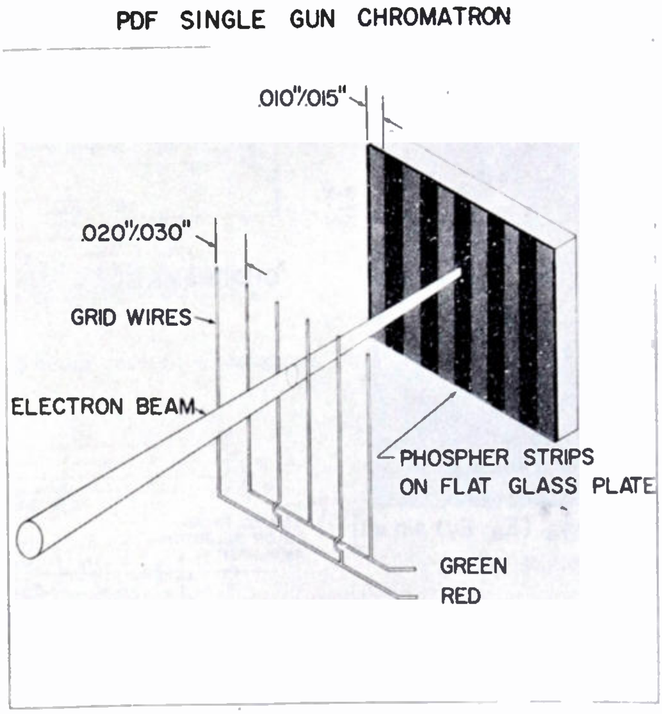


Fig. 1

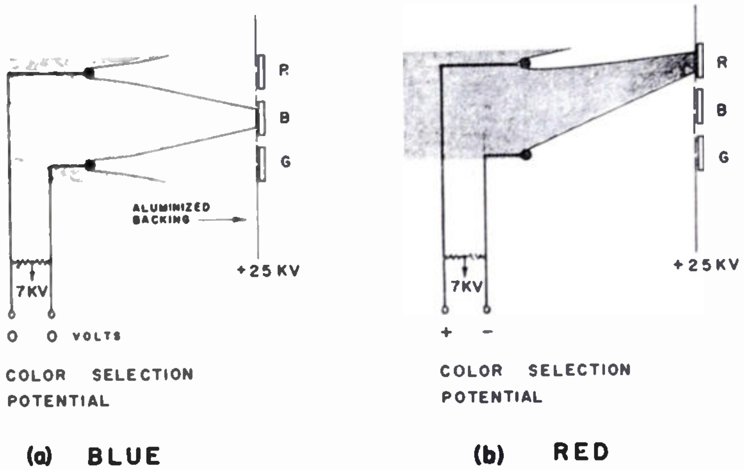
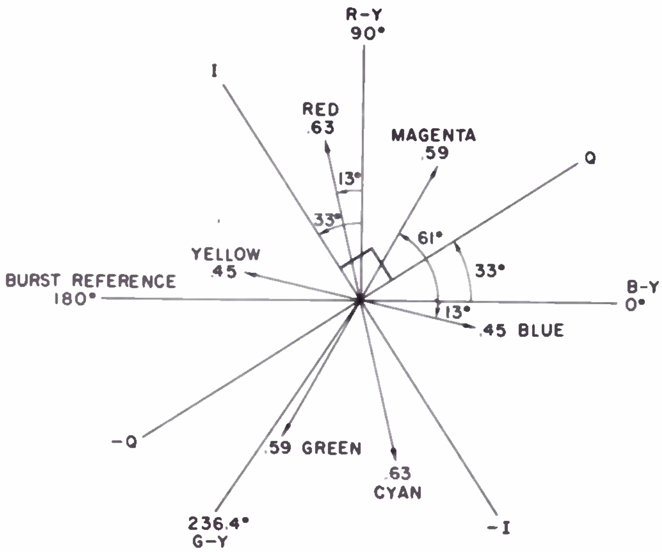


Fig. 2. Color selection in the single-gun CHROMATRON.



$$E_M = E_Y + \frac{1}{1.14}\left[(E_R - E_Y)\cos wt + \frac{1}{1.78}(E_B - E_Y)\sin wt\right]$$

Fig. 3. Polar diagram showing various phase relationships of the NTSC signal color components.

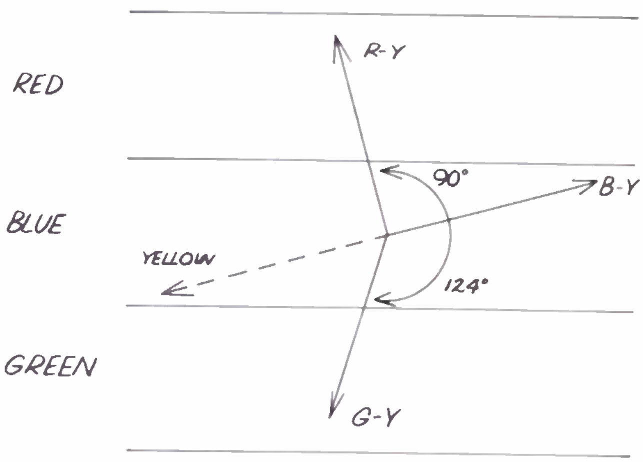
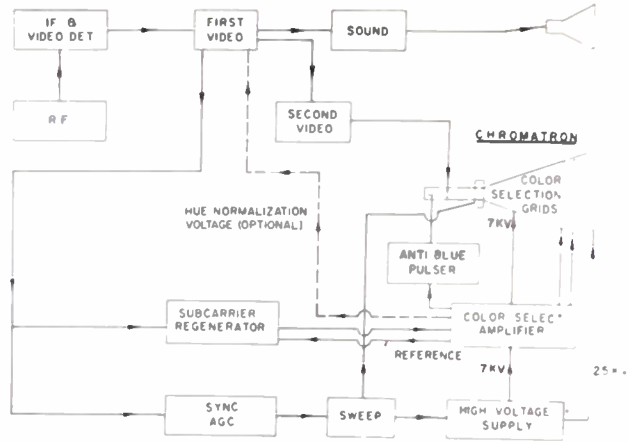


Fig. 4. NTSC color difference vectors superimposed on a double-blue sequence CHROMATRON.



QUADRAMATIC COLOR RECEIVER

Fig. 5

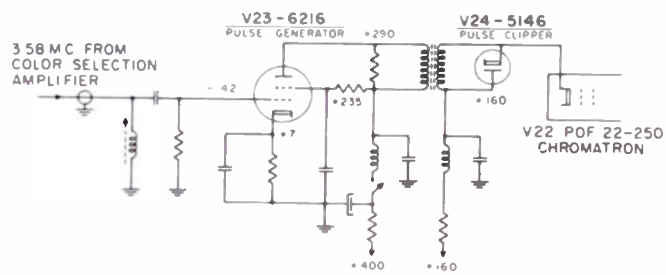


Fig. 6. Schematic anti-blue pulser.

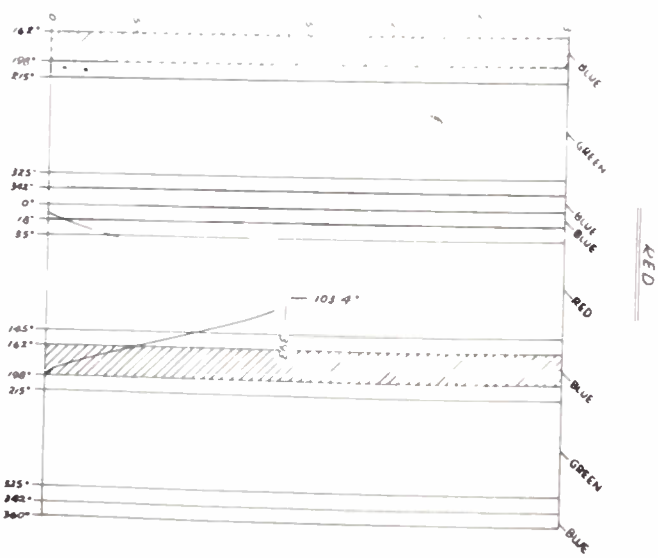


Fig. 7. Quadramatic decoding beam current waveforms.

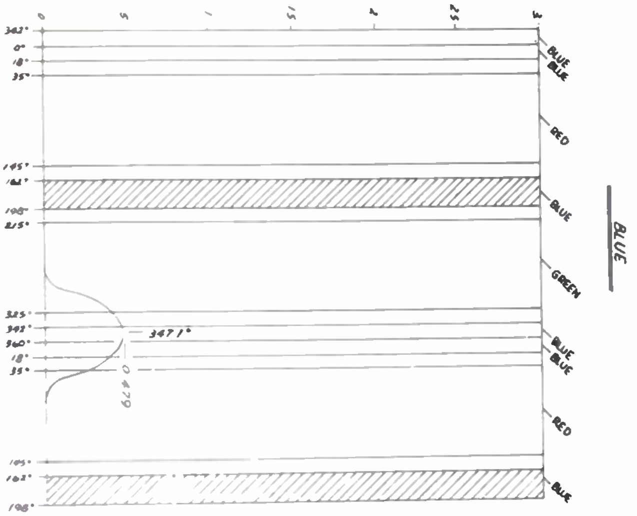


Fig. 8. Quadramatic decoding beam current waveforms.

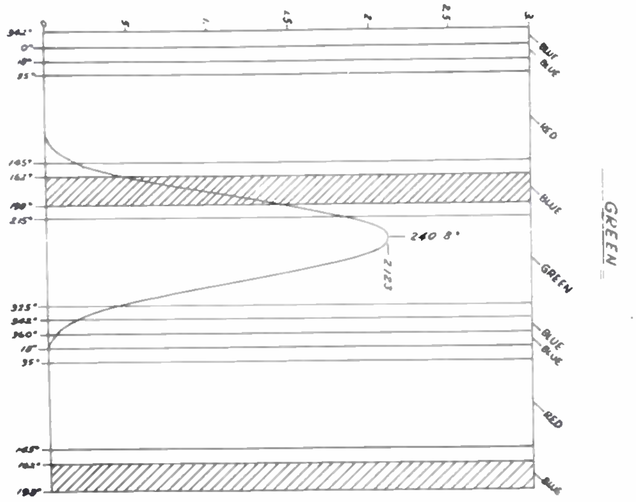


Fig. 9. Quadramatic decoding beam current waveforms.

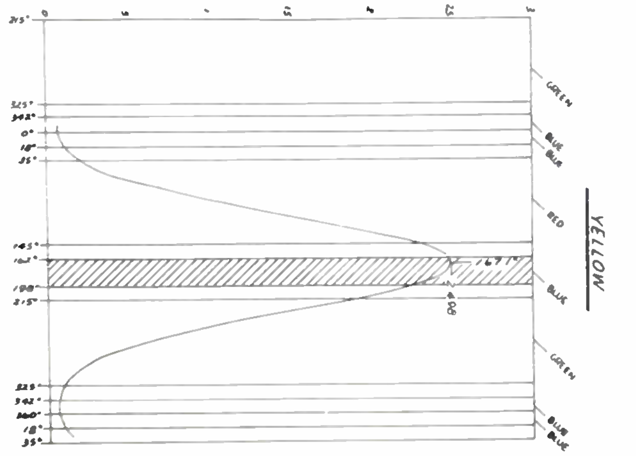


Fig. 10. Quadramatic decoding beam current waveforms.

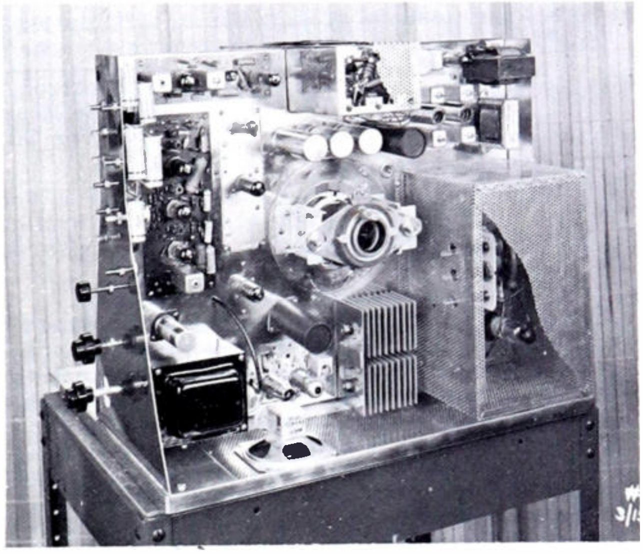


Fig. 11. Rear view quadramatic color receiver.

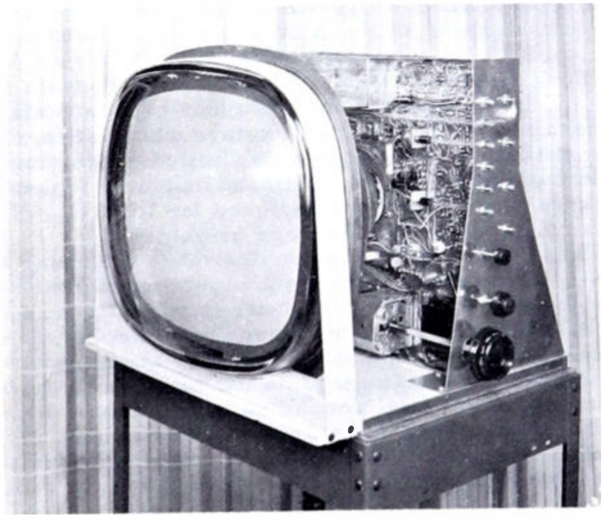


Fig. 12. Side view quadramatic color receiver.

TWENTY-ONE INCH DIRECT-VIEW STORAGE TUBE *

by

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Summary

Applications for storage tubes have increased in recent years with the development of the 5-inch direct-view storage tube. This paper discusses the development of a 21-inch storage tube. The tube utilizes a metal envelope with curved faceplate and a usable viewing diameter of about 18 inches. Externally, the tube is similar to an ordinary television picture tube. Internally, a high-velocity electron gun, a low-velocity flood gun, and storage target permit writing and storing information to be viewed after the original signals are gone. The storage target assembly is curved to match the faceplate curvature, thus avoiding a separate viewing screen inside the tube. Photoetched Cu-Ni meshes are used in the storage assembly. The mechanical and electrical problems encountered in the development will be discussed and performance figures will be given.

Introduction

Direct-view storage tubes of 5-inch size have been developed to the stage where several manufacturers have them in limited production. In contrast to the conventional indicator tubes, these tubes have been designed for long persistence at high average brightness.[1,2] These characteristics are desirable for radar and for simultaneous storage and presentation of television pictures. For many applications, including radar and processed data display, indicators having a larger viewing area are desirable. The larger tubes offer ease of viewing and more total resolution over the tube diameter. Recently a 21-inch direct-view storage tube has been under development for such applications.

In many respects, the 21-inch tube can not be simply scaled from the 5-inch tube. For example, the fine metal screens used in the 5-inch diameter tubes would not be strong enough to be supported over a large diameter without objectionable vibration. A thicker mesh material such as the etched copper-alloy mesh, used for color television shadow-mask screens, must be used in the storage assembly. The storage mesh must be curved to match the faceplate curvature of the envelope. The development of the 21-inch tube has required the solution of these and many other similar fabrication problems.

Description of Operation

Direct-view storage tubes commonly operate in one of two modes: the bi-stable, two-tone mode;[3] or the halftone mode. The halftone mode is used with the 21-inch tube because fabrication of the target is greatly simplified. This mode of operation has been fully described in connection with the 5-inch tubes[2,4], but perhaps a brief description should be given before going further. Figure 1 shows the essential components of the tube: the writing gun, the flood gun, the storage-target assembly, and the viewing screen. An enlarged section of the storage-target assembly is shown pictorially on the same figure. The assembly consists of a collector mesh, a storage mesh with dielectric material evaporated on its surface, and a viewing screen. In the unwritten state, the dielectric surface is charged negatively, thus repelling the low-energy flood electrons back to the collector. Where the high-velocity writing beam has charged the dielectric in a positive direction by secondary emission, flood electrons can penetrate the storage mesh and are then post-accelerated to the viewing screen. The amount the writing beam charges this dielectric, controls the number of flood electrons penetrating and thus the brightness of the screen. In Fig. 2 we plot a typical transfer characteristic of brightness versus dielectric surface potential. All of the voltages shown are referred to the flood-gun cathode potential. The figure shows the half-tone range and the cutoff potential.

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* This work has been supported in part by the Lincoln Laboratories of the Massachusetts Institute of Technology, Cambridge, Massachusetts. Typotron is the registered trade-mark of the Hughes Aircraft Company for its brand of character-writing storage tube. Tonotron is the registered trade-mark of the Hughes Aircraft Company for its brand of half-tone storage tube.

The flood beam is used for erasure of written information. This erasure mode depends on two factors: (1) there is a high capacitive coupling between the dielectric surface and the storage mesh, (2) low-energy electrons strike the dielectric and charge it in a negative direction, since the secondary ratio is less than one. This means that when the storage mesh is elevated by an amount equal to the cutoff potential, i.e. 8.4 volts in the characteristic of Fig. 2, the storage surface will follow by capacitive coupling. All parts of the dielectric previously at operating range potentials will then be slightly positive and flood electrons can strike the surface, charging it down to flood-cathode potential. When the storage mesh is then returned to its normal voltage, the surface is carried capacitively to the negative cutoff potential. This process can be accomplished in one pulse or in many pulses depending on the application. When one pulse erasure is used, the erase time is about 600msec for the present tubes. When a pulse train of low duty cycle is used, erasure time can be prolonged and the written information will fade down with an arbitrary time constant controlled by the duty cycle of the erase pulse.

Design

Broadly, the problems encountered in the design of this tube can be divided into mechanical and electrical problems. The mechanical problems are those of envelope modification, mesh curving, and mesh mounting. The electrical problems are those of flood gun, flood beam collimating lens, writing-gun design, and storage-target design.

Rather than to institute a program of fabricating a new envelope, it appeared desirable to make use of available television bulbs. Since the storage assembly must be inserted in the front end before the sealing operation, this narrowed the selection to color television envelopes. These come in two sections: funnel and panel. The two sections are sealed together after the storage assembly has been mounted inside the panel and the electrical leads connected. The envelope chosen was the metal 21AXP22 color bulb since the ratio of usable diameter to maximum diameter was large, and the metal panel made electrical connections easier. The electrical connections were made in the panel section through holes drilled in the panel and leads sealed in with glass. The sealed electrical leads are shown in Fig. 3.

The storage target assembly consists of the collector mesh and storage mesh. Both are photoetched Cu-Ni meshes similar to the type used in color television shadow masks. The meshes are first annealed, then curved with a punch to match the faceplate curvature of the envelope. Magnesium fluoride is evaporated as the dielectric on the storage screen; then the meshes are mounted together as shown in Fig.4. Insulated studs are used to keep the screens electrically insulated from each other and from the panel.

The flood gun and collimating lens are designed to provide 5 ma of low-energy electrons uniformly over the target area. The flood gun is shown mounted inside the funnel in Fig. 5.

The writing guns have been of two types: a conventional spot-writing design for the 21-inch Tonotron tube, and a character-writing design for the 21-inch Typotron tube. This character-writing gun is designed to give 0.200-inch characters at the screen. A choice of 63 characters is available.

Providing uniform display over the target area with a minimum of shading has been a problem in the storage-target design. Shading can be described as nonuniformity in brightness when a constant signal level is written over the entire screen. It is most pronounced at low-brightness levels where some areas may be black and others not. One can consider this as variation in μ over the storage target. The factors which affect shading are: (1) nonuniform spacing between the collector and storage meshes, (2) nonuniform spacing between the storage mesh and viewing screen, (3) variation of hole size in the storage mesh, (4) nonuniform flood-electron collimation, and (5) stray magnetic effects. These have been analyzed separately and their effects are noted in Figs. 6, 7, and 8 as cutoff potential variation in the transfer characteristics of Fig. 2.

The two spacing effects have been analyzed experimentally, and are shown in Figs. 6a and 6b, respectively. Since the two meshes are curved by punching, the variation in spacing between the collector and storage can be held to a close tolerance. The spacing variation between the viewing screen and storage mesh is more severe. The faceplate curvature itself is not uniform and varies from panel to panel. Cutoff variation from nonuniform flood-beam collimation is caused by aberrations of the collimating lens. This affects the angle of incidence of the flood electrons at the target assembly. Figure 7 shows an approximate expression for the variation of cutoff from this source. At the collector the normal component of flood-electron velocity is

$$v_n = v \cos\theta \tag{1}$$

where v is the total velocity and θ is the angle of incidence at the collector. For the normal component of kinetic energy, we write

$$e\,W_n = eV \cos^2\theta \tag{2}$$

where V is the collector potential. The cutoff variation with angle of incidence, ΔV_{cutoff}, is due to the change in this normal component of energy. Flood electrons are collimated so that θ_{min} is usually zero and θ_{max} is small. The equation then reduces to the approximate form:

$$\Delta V_{cutoff} \cong V\,\theta^2_{max}. \tag{3}$$

From trajectory plots in the electrolytic tank, a value of θ_{max} can be obtained. Shading due to variation of hole size of the storage mesh is similar to the variable μ problem in receiving tubes. However, the problem is complicated by the dielectric surface which is at a different potential from the storage mesh. Computation shows that no virtual cathode exists near this screen. In first approximation, this problem has also been studied in the electrolytic tank. Figure 8 shows a curve, cutoff vs. storage mesh transmission. Magnetic effects can be minimized by careful de-magnetizing of the tube parts and shielding against stray fields. For an estimate of the worst case of shading possible, we can expect all these factors to be additive. In the present design, the estimated maximum variation in cutoff is an appreciable fraction of the halftone range. Measurements from actual tubes give better figures, although the shading presently observed is greater than would be desired for good halftone reproduction. Of course, for applications using an "on-off" type of writing where halftones are not displayed, such shading effects can be minimized and become less noticeable.

The two types of completed 21-inch storage tubes are shown in Figs. 9 and 10. They differ only in the writing-gun design. Figure 9 shows an electromagnetically focused and deflected spot-writing tube; the character-writing tube is illustrated in Fig. 10.

Performance

Generally, the performance of the 21-inch tube is somewhat slower than its 5-inch cousin. For the spot-writing variety, writing speeds of the order of 20,000 in/sec have been measured. By the shrinking-raster method, written resolutions of about 30 to 40 lines-per-inch can be obtained with the resolution being better at the low half-tones. Over the usable diameter of the tube; this means about 500 to 700 written lines. This is in comparison with nominal 300 lines resolution of the 5-inch tubes. Retention times of the order of one minute are available and a brightness of about 50 foot-lamberts has been obtained. As an example of controlled-fade erasure, Fig. 11 shows plots of typical short and long decay characteristics. As mentioned before, the shading problem at present is serious enough to affect good half-tone presentation. However, as a final test, a slow-scan TV picture has been stored on the 21-inch Tonotron tube and is shown in Fig. 12. In comparison, a similar one from a 5-inch Tonotron tube is shown in Fig. 13.

As an example of "on-off" type of display, Fig. 14 is a stored picture on the character-writing version of the tube. Nominal height of the characters is 0.200-inches. This means that about 2000 well-spaced characters can be written in an inscribed square raster of 12.5-inch on a side. Each character can be written in 500 μsec.

Conclusion

Many of the mechanical problems and fabrication techniques have been solved satisfactorily. Future development on the 21-inch storage tube will include improvements in uniformity over the entire screen and increased writing and erasure speeds. Both the spot-writing and character-writing tubes should be particularly applicable for radar and data-processing displays, where large amounts of information need to be displayed under ordinary ambient lighting conditions.

Acknowledgement

The authors greatfully acknowledge the advice and counsel given by Mr. H. M. Smith and Dr. G. F. Smith throughout the development program. Mr. A. Anderson showed valuable skill in doing the glass work. Mr. Chester Beintema has contributed much toward the fabrication of the tube. Buckbee-Mears Company cooperated closely and supplied the screens for the storage assembly. Lincoln Laboratories of the Massachusetts Institute of Technology supplied the character stencil for the tube.

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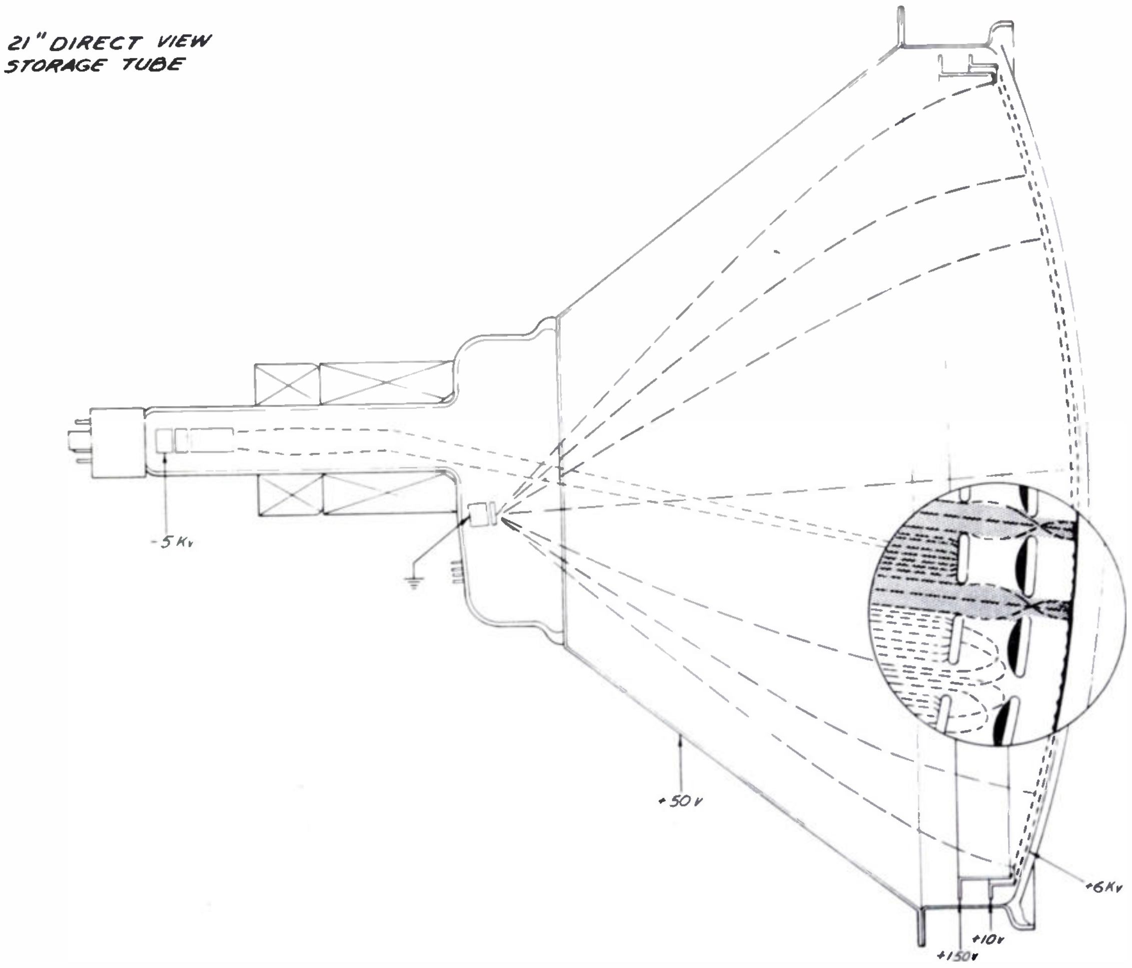

Fig. 1. Schematic of the 21-inch Tonotron tube.

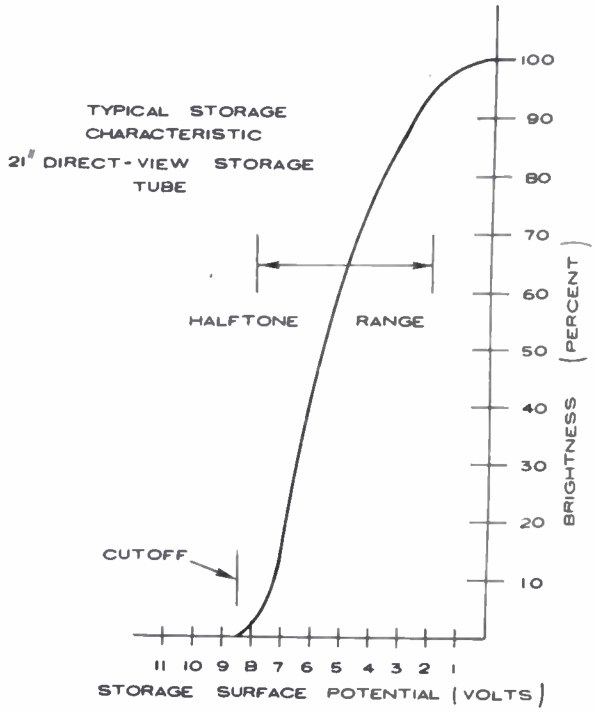

Fig. 2. Typical storage characteristics.

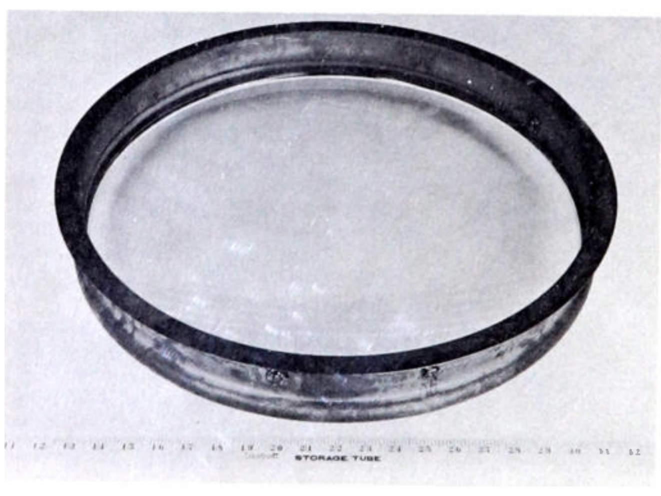

Fig. 3. Sealed electrical leads in the panel section.

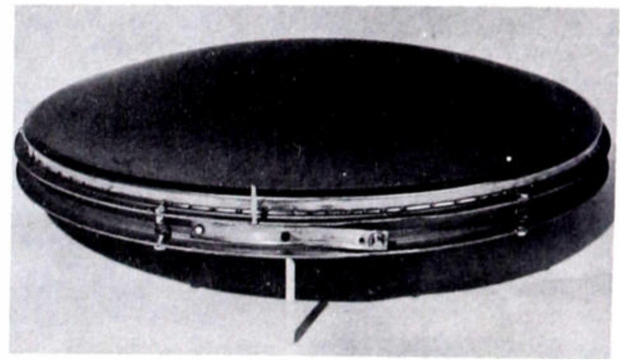

Fig. 4. Storage target assembly.

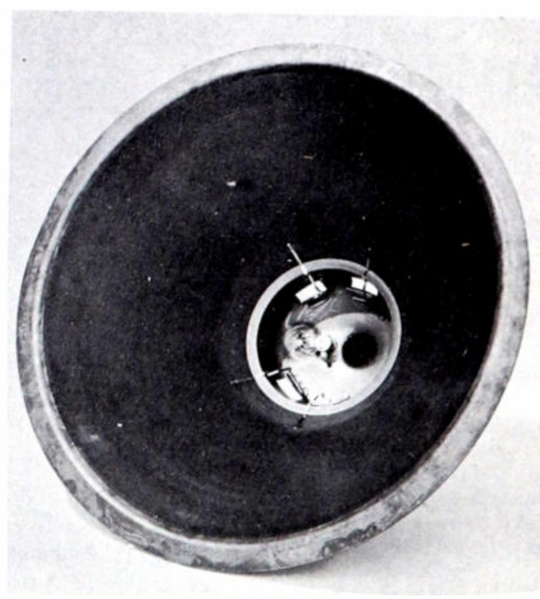

Fig. 5. Flood gun mounted in the funnel section.

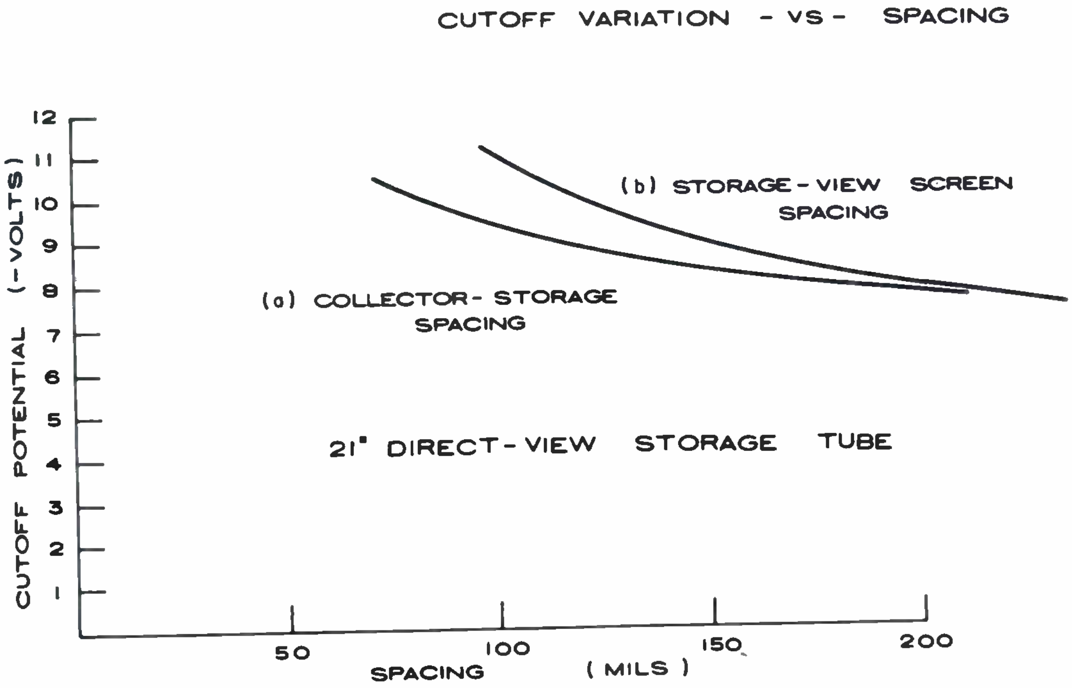

Fig. 6. Cutoff variation with spacings.

CUTOFF VARIATION - VS - ANGLE OF INCIDENCE

NORMAL COMPONENT OF VELOCITY, $v_n = v \cos \theta$

NORMAL COMPONENT OF KINETIC ENERGY, $eW_n = eV\cos^2 \theta$

$$\Delta V_{CUTOFF} = \Delta W_n$$

$$= V\left[\cos^2 \theta_{MIN} - \cos^2 \theta_{MAX}\right]$$

$$= V\sin^2 \theta_{MAX} \quad (\text{FOR } \theta_{MIN} = 0)$$

$$\cong V(\theta_{MAX})^2$$

Fig. 7. Cutoff variation with flood beam angle of incidence.

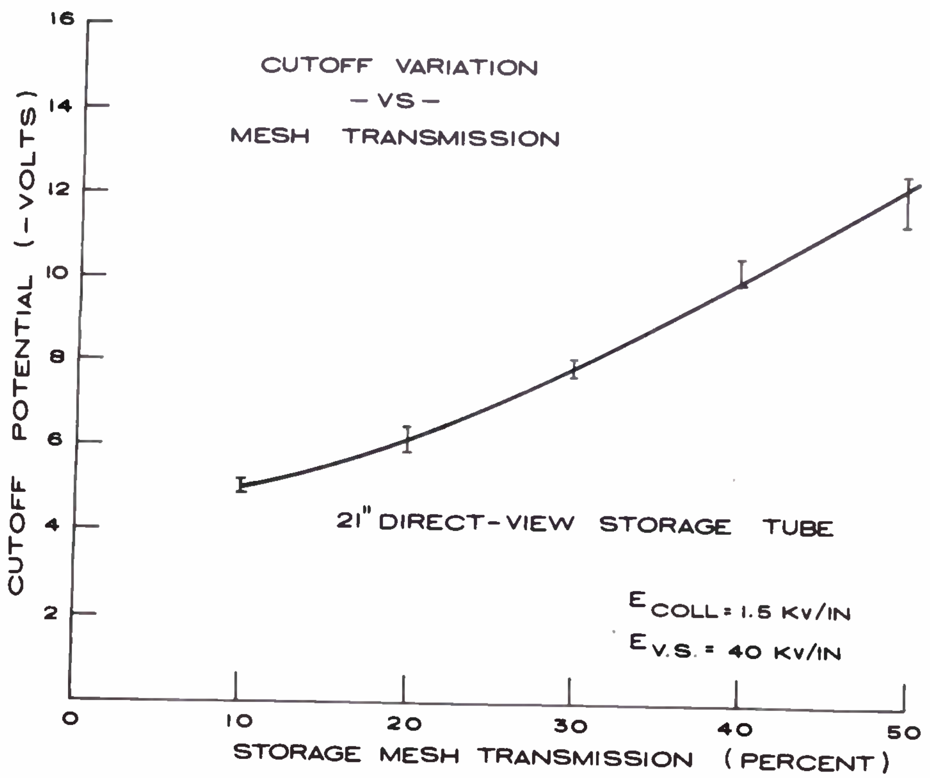

Fig. 8. Cutoff variation with storage mesh transmission.

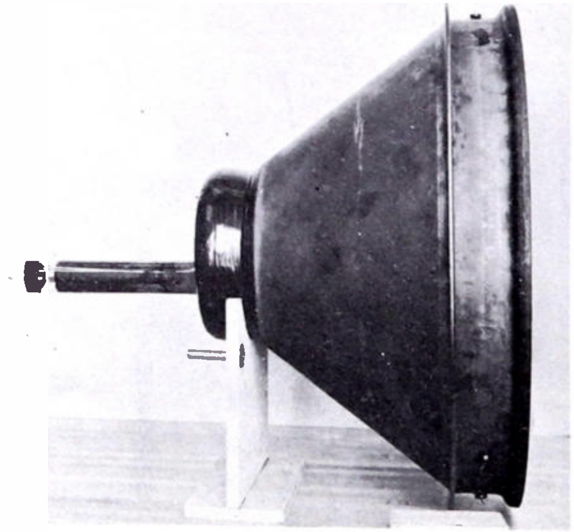

Fig. 9. 21-inch Tonotron tube.

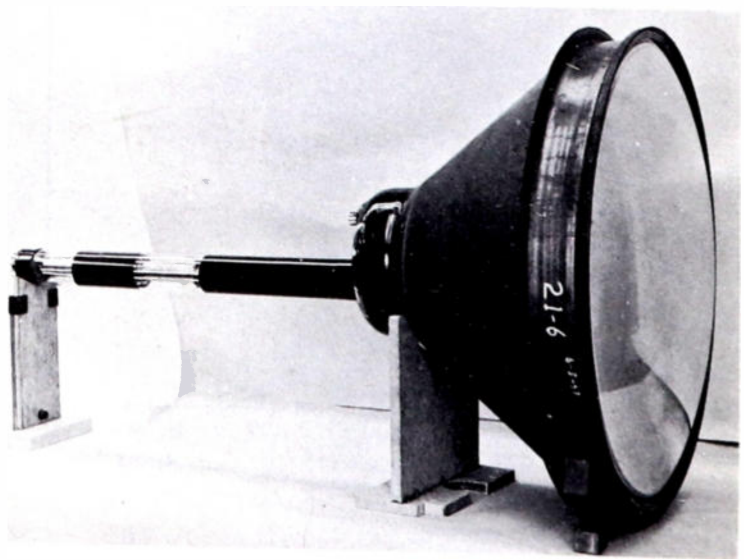

Fig. 10. 21-inch Typotron tube.

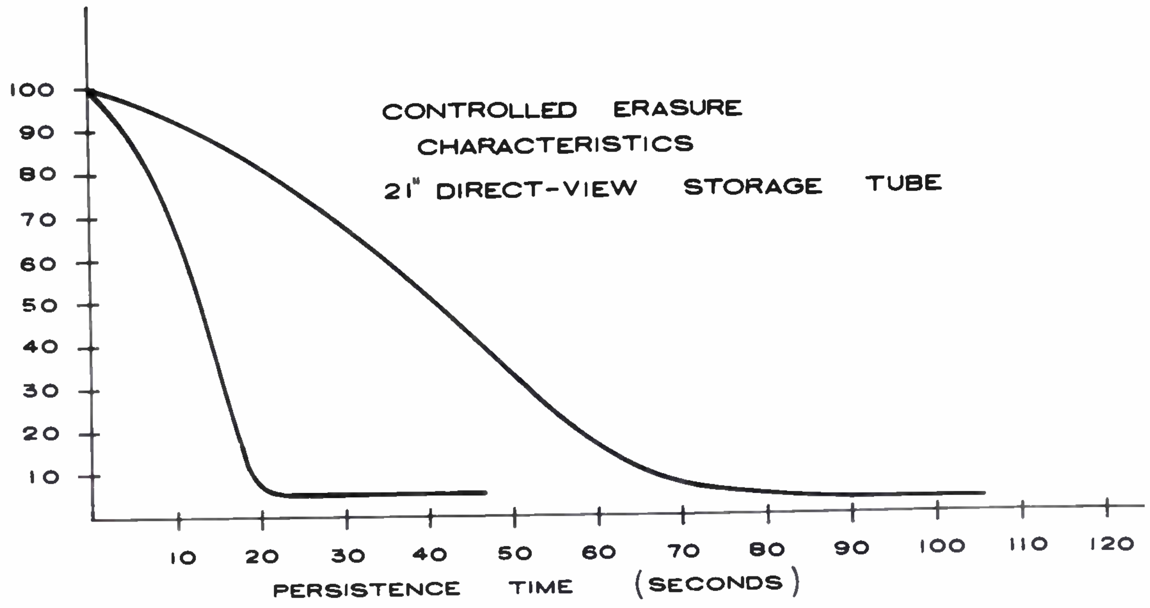

Fig. 11. Controlled erasure curves.

Fig. 12. Slow-scan TV picture on 21-inch Tonotron tube.

Fig. 13. TV picture on 5-inch Tonotron tube.

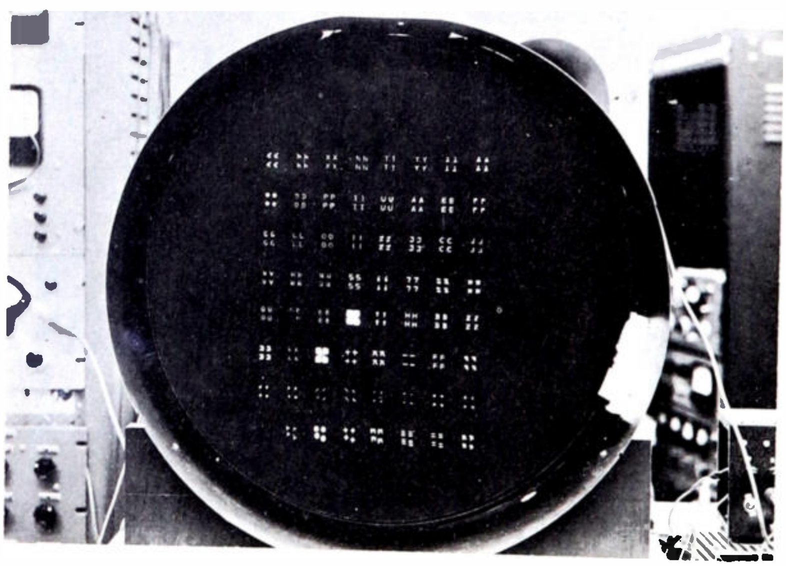

Fig. 14. Stored characters on 21-inch Typotron tube.

THE TELEVISION COLOR TRANSLATING MICROSCOPE

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and
Rockefeller Institute for Medical Research, New York, N.Y.

Summary

The television color translating microscope permits the continuous observation of specimens under intermittent ultraviolet illumination, translating differences in absorption in different ultraviolet spectral ranges into color differences in the observed picture. Thus the presence of some chemical constituents of organic cell and tissue specimens which are completely transparent in the visible and hence undifferentiable in the ordinary microscope can be readily recognized by their distinctive color. The principles of construction of the television color translating microscope will be discussed with special reference to an instrument built at the Rockefeller Institute for Medical Research. Color slides will serve to show some preliminary results obtained with this instrument.

The television color translating microscope is a new research tool of the biologist which permits him to use his color vision to distinguish between elements of the specimen differing in their spectral absorption in the ultraviolet.

It is well known that many constituents of unstained living cells, although practically transparent to visible light, have strong characteristic absorptions in the ultraviolet, particularly in the range from 2200 to 3000 A. To reveal their presence, E. M. Brumberg suggested photographing the specimen when illuminated, in succession, with radiation of three different ultraviolet wavelength ranges and using the three negatives so obtained as separation negatives for preparing a color print. Ultraviolet absorption differences would thus be translated into color differences in the final picture. Photographic color translation microscopes based on this principle and utilizing high-speed processing techniques have been constructed by E. H. Land and his coworkers and, more recently, by H. P. Hovnanian and R. B. Holt. In these instruments the final color picture is viewed on a screen, on which red, green, and blue images of the three separation negatives are projected in superposition.

In the television color translating microscope an ultraviolet-sensitive television camera takes the place of the photographic camera, a color television receiver that of the triple projector. The time delay between exposure and observation is eliminated, so that changes in spectral absorption and morphology of specimens can be observed continuously. Furthermore, large quantities of material can be examined with the same speed and ease as in visible-light microscopy. Finally, since the contrasts provided by the differential absorptions of the cell constituents are large and render staining and fixation unnecessary, the technique is particularly well adapted to the study of living cells and tissue cultures.

The basic plan of a television color translating microscope is shown in Fig. 1. A microscope with reflective condenser and objective projects the image on the photosensitive target (or targets) of a television camera, whose video signal is employed for the reproduction of an image of the specimen on the screen of a color television receiver. Illumination is provided by one or several pulsed light sources and monochromators so arranged that radiation of the three selected ultraviolet wavelengths falls on the specimen at successive pulses. The pulse source also controls the vertical deflection in the camera and receiver in such fashion that the pulses of illumination occur during vertical fly-back time, the picture signal stored by a radiation pulse of a particular ultraviolet wavelength being utilized for generating a component picture of the corresponding color in the succeeding frame period.

In the first television color translating microscope, constructed several years ago at the RCA Laboratories in Princeton, a single pulsed light source and monochromator was employed in conjunction with a standard industrial television camera provided with an experimental ultraviolet-sensitive Vidicon. A simple mechanical system was employed to select in turn radiation of three different wavelength ranges for specimen illumination.

The method of wavelength selection (Fig. 2) is based on the fact that the focal points for a prism monochromator with uncorrected collimator and telescope lenses made of the same material as the prism lie along a straight line which is oblique to the direction of the reflected pencils. If the entrance slit of the microscope illuminator is placed on this straight line, a plane mirror perpendicular to the line will project the focal point for radiation of different wavelengths onto the slit as it is displaced in a direction perpendicular to its surface.

In the instrument the plane mirror took the form of a disk consisting of three mirror sectors, recessed by a few tenths of an inch with respect to each other in correspondence with the three selected ultraviolet wavelengths (Fig. 3). The sector disk was driven by a synchronous motor so phased with respect to the vertical deflection of the television system that the specimen was illuminated with the appropriate ultraviolet wavelength in the fly-back time preceding each color frame.

While the principle of color translation could be readily demonstrated with this instrument, it had several shortcomings. Perhaps the most obvious of these was the difficulty of changing the wavelength selection. Every change in wavelength demanded the insertion of a new sector mirror or plane parallel shim to adjust the height of the mirror surface. Furthermore the lag of the experimental Vidicon caused a certain amount of signal carry-over from one frame to the next, making it impossible to obtain highly saturated colors. Finally, the variation in the angle of incidence of the illumination with wavelength made it difficult to achieve color balance over the entire field.

These drawbacks were overcome in a more elaborate television color translating microscope constructed at the Rockefeller Institute in New York. The illuminating system of the instrument is shown in Fig. 4. Three Farrand grating monochromators with individual pulsed light sources permit arbitrary selection of the illuminating wavelengths by means of calibrated knobs on the front panel. The cover of one of the sources, a Hanovia 10B1 quartz mercury arc lamp, is removed in the illustration. The use of separate light sources has the incidental advantage that sources can be selected for high emission in particular wavelength ranges.

The optical arrangement of the illuminator is more clearly evident from the diagram in Fig. 5. The source is imaged by a quartz lens on the entrance slit of the monochromator and the radiation from the exit slit is directed onto the center of rotation of a 45°-mirror which reflects it into the microscope substage, an f/3.5 aperture ratio being maintained throughout. The amplified vertical pulses from the synchronizing generator of the television system are applied through contact brushes on slip rings attached to the mirror stage to the three thyratron circuits feeding the light sources in succession, triggering a light flash at the instant when the principal ray from the monochromator and the normal to the rotating mirror surface lie in a common vertical plane. Synchronism between the mirror rotation and the synchronizing generator is assured by the fact that a 60-cycle wave derived from the commutator controls the phase of the synchronizing generator.

The circuit is shown in broad outline in Fig. 6. It shows how the synchronizing signal is generated from the output of the commutator and serves, at the same time, to pulse the several light sources and to gate, by means of a multivibrator circuit, the corresponding guns of the color kinescope.

In the new instrument (Fig. 7) a custom-made image orthicon with ultraviolet-transmissive window is employed as camera tube. With this tube no objectionable lag effects have been observed. Furthermore, the image orthicon has the added advantage of enhanced sensitivity. At the same time, we may note that certain further improvements which would reduce cost, simplify optical alignment, and minimize flicker, are under consideration.

A few color slides obtained by photographing the screen of the color kinescope illustrate the results obtained. All of the specimens examined were unstained and, so far as possible, unprepared.

The first shows a suspension of Frog blood. The ultraviolet wavelengths employed for illuminating the specimen were here so selected with reference to the three primary colors that the nuclei of the red blood cells appear orange, the hemoglobin yellow, and the smaller white blood cells purple. By changing the correlation between the ultraviolet wavelengths and the primary colors a variety of different effects can be realized.

The frog-blood specimen is also well suited for illustrating the use of the technique for obtaining quantitative data on the absorption of the specimen for specific wavelengths and, hence, the relative concentration of substances with characteristic absorptions. A particular line of the scanning pattern, which appears white in the picture of the cell suspension, is selected by a line-selector oscilloscope and the corresponding outputs for the red, green, and blue pictures, corresponding to the three selected ultraviolet wavelengths, are represented on three separate oscilloscope screens. In order to render this technique truly quantitative, we expect to make our illuminator optics achromatic throughout and to employ density step wedge calibration. The television color translating microscope will then become an instrument not only of observation, but also of quantitative measurement.

Since the major absorptions are usually due to nuclear and protein materials, the cell nuclei frequently are the only strongly colored structures in a preparation. However, when components with high absorption coefficients are widely distributed in the specimen, as for onion epithelial cells, the picture is characterized by highly saturated colors. Small black granules in the picture could be seen to undergo Brownian motion and to accumulate in the purple nuclear region after approximately half an hour of examination. Pigments such as chlorophyll and hemoglobin, which are highly colored in gross amounts but scarcely detectable in microscopic preparations, appear brilliantly colored in the ultraviolet color translating microscope, as illustrated by pictures of chloroplasts of algae. The streaming of the protoplasm in the chloroplasts could be readily observed on the color kinescope screen.

Frequently, the detection of a particular substance with a characteristic transmission or absorption in the ultraviolet is of primary interest. This may be aided materially by the slow periodic interruption of the corresponding source of illumination or color channel. The presence of flicker then supports the detection on the basis of color alone.

In summary, the television color translating microscope makes it possible to examine unstained living material in the ultraviolet with the same ease and discrimination of detail which, in the visible, is achieved by the use of specific stains and fixation, which generally result in the death of the cells. Furthermore, since any primary color may be correlated with any characteristic absorption wavelength in the ultraviolet, the range of "electronic stains" which may be employed in the color translation technique is extraordinarily wide. Thus the television color translating microscope should prove not only a powerful tool in the study of vital processes in normal and abnormal cells and tissues, but may also become a valuable aid in diagnosis.

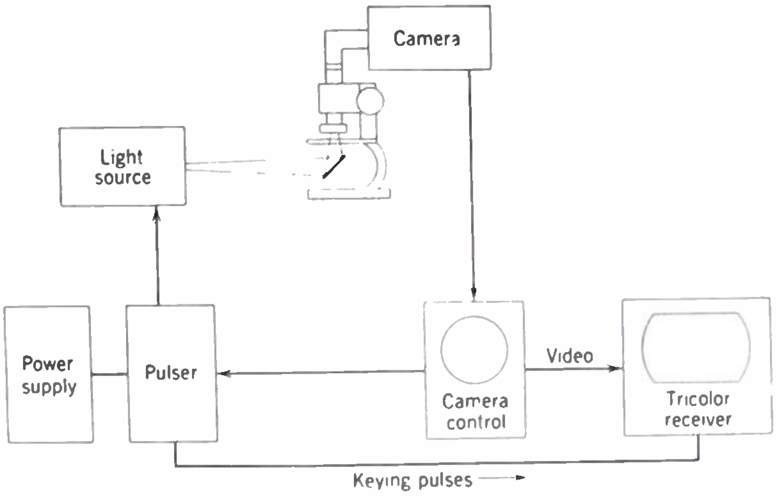

Fig. 1 General Block Diagram of Television Color Translating Microscope.

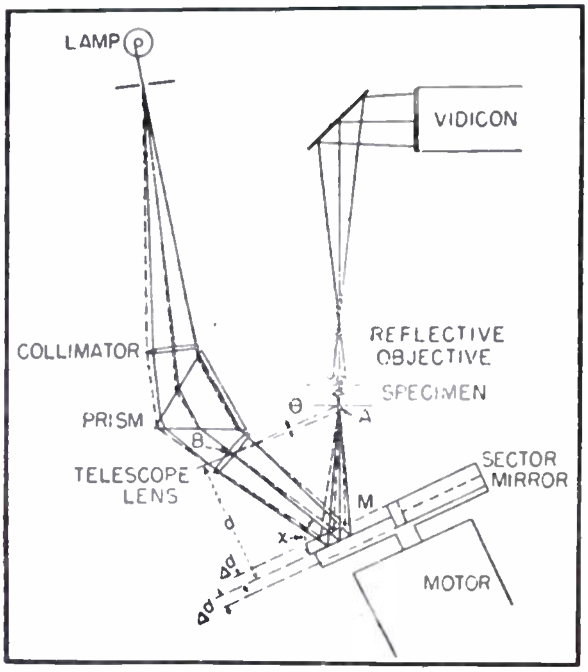

Fig. 2 Wavelength Selection in Single-Source Color Translating Microscope.

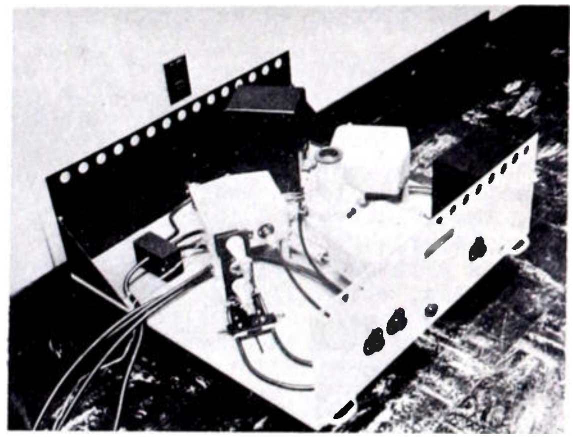

Fig. 4 Illuminating System of Television Color Translating Microscope with Three Sources

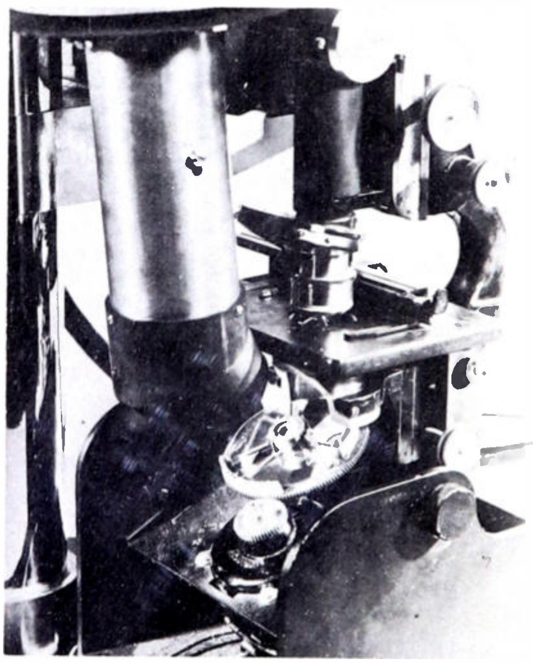

Fig. 3 Single-Source Color Translating Microscope: Construction of Illuminating System.

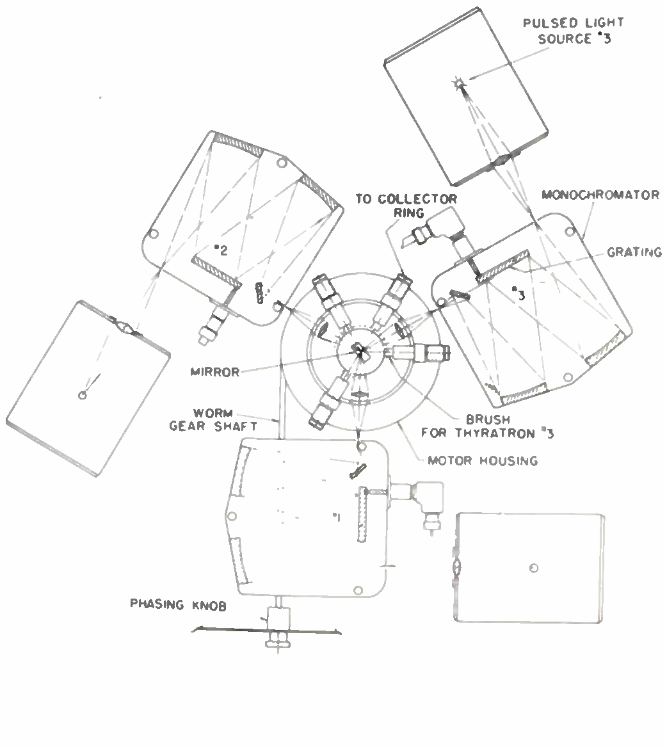

Fig. 5 Diagram of Illuminating System.

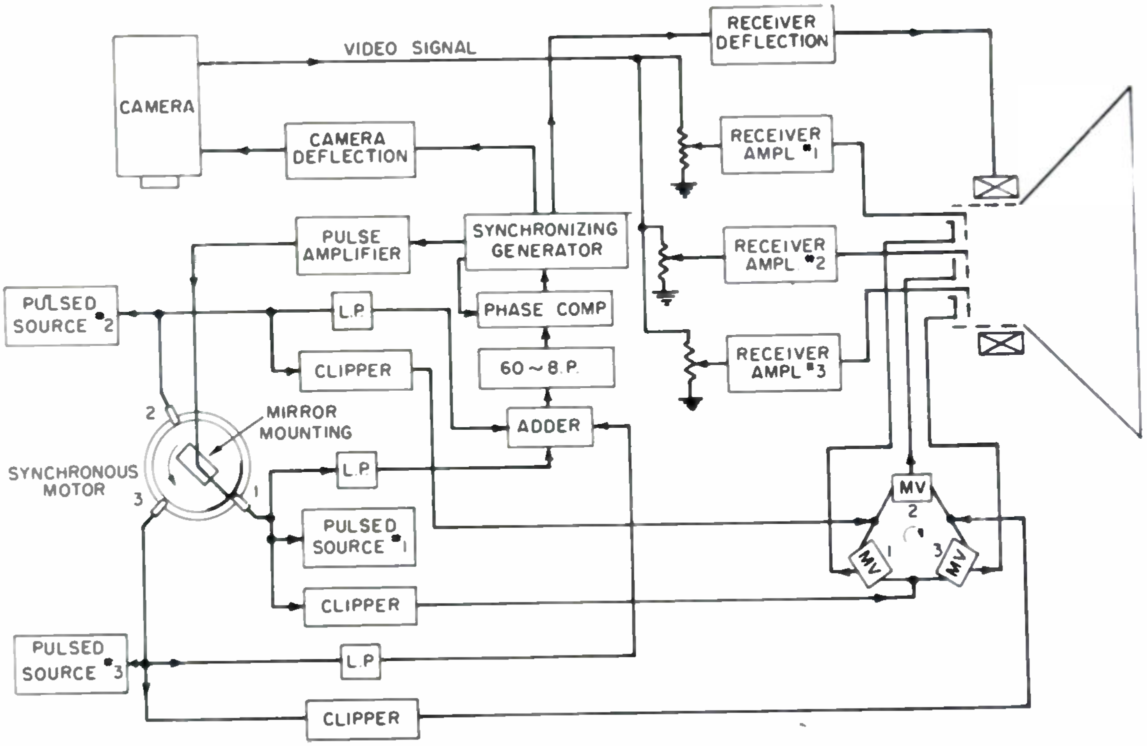

FIG. 6 — BLOCK DIAGRAM FOR COLOR TRANSLATING MICROSCOPE AT ROCKEFELLER INSTITUTE

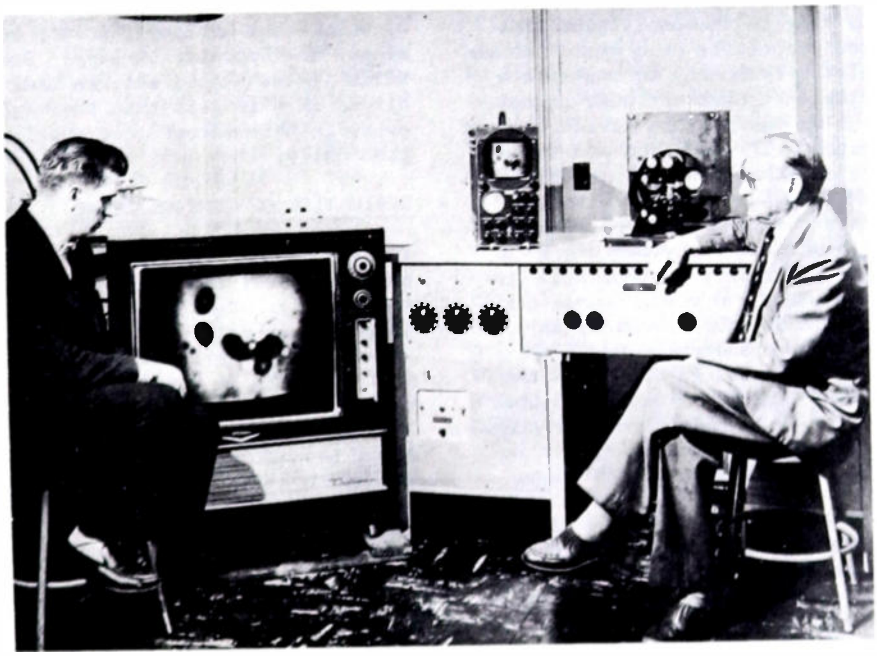

Fig. 7 The Complete Television Color Translating Microscope with Image Orthicon Camera.

AUTOMATIC FINE TUNING FOR TELEVISION RECEIVERS

by

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SUMMARY

A different approach to the problem of automatic frequency control of television receivers is presented. The system uses the amplitude of the intercarrier sound signal and is shown to closely approximate subjective manual tuning over a wide range of conditions. Frequency drift reduction factor and pull-in range limits are determined for the basic system. Two methods are shown that can be used to increase the pull-in range. The advantages of using this system in conjunction with a single detector color receiver are discussed.

GENERAL CONSIDERATIONS

Figure 1 shows a block diagram of the system. The dotted line shows an interconnection, to be explained later, which increases the pull-in range of the system. The 4.5 megacycle intercarrier signal at the second detector is amplified in the video amplifier and further amplified in a sound amplifier. The amplitude of the 4.5 megacycle signal is used to control the reactance device. The action of such a system is to maintain a constant level of 4.5 megacycle signal at the reactance device by varying the oscillator frequency. Examine the properties of a system which controls the oscillator frequency to maintain a constant level of the intercarrier sound signal. The A.G.C. system holds the picture level constant at the second detector. If the video and sound amplifier gains are constant, the 4.5 megacycle sound level at the second detector is also constant. The result is that the beats between the high frequency components of the video signal and the sound signal are indirectly controlled. In general this is the same way a viewer tunes a receiver manually, the oscillator frequency is raised to increase picture sharpness until the benefit of increased sharpness is offset by the appearance of too much "sparkle" in the picture. Furthermore, this control is effective over transmission, antenna and receiver passband tilts and operates dynamically for the case of "airplane flutter".

The method of tuning also provides considerable control over the 920 Kc beat interference between the sound and chroma signal of color broadcasts. Quantitative data showing this effect will be given later.

CONTROL SYSTEM ANALYSIS

Figure 2 shows the information necessary to understand the open and closed loop behavior of automatic frequency control systems.

The "discriminator" curve is designated the "μ" curve where μ is equal to the operating slope of the "μ" curve. The reactance curve is called the "β" curve and β is equal to the negative reciprocal of the slope of the "β" curve. The intersection of these two curves determines the closed loop frequency. The dotted "β" curve represents a shift in the open loop frequency to f_2, and the closed loop frequency becomes $f_1 + \Delta f$. Simple trigonometry shows that the frequency drift reduction factor is $\frac{1}{1-\mu\beta}$.

It is possible to have a large drift reduction factor and yet have a large change in correction voltage existing in the closed loop. This is shown for a high gain system by $V = F/-\beta$. The β required to maintain a given level of control signal in the presence of frequency drift follows directly from the formula.

The other important aspects determinable from these curves are the stability of crossover points and the pull-in and hold-in range. It can easily be shown that crossover points are stable if the slopes are opposite in sign. Stability also results if the slopes are the same sign if the magnitude of μ is less than the magnitude of β. However, in this latter case the loop gain is less than unity.

DERIVATION OF CONTROL CURVE

In order to understand the proposed A.F.C. system quantitatively the μ curve must be derived. The intercarrier sound amplitude versus oscillator tuning is largely determined by the A.G.C. system and the I.F. characteristic. Figure 3a shows the I.F. curve to be used. The curve is similar to conventional receivers except that the sound trap is deeper and is tuned somewhat lower in frequency. Also, the outband "pop-up" on the sound side has greater attenuation. The purpose of the I.F. passband shaping is to provide a desirable curve of 4.5 megacycle amplitude versus oscillator frequency.

The exact details of the A.G.C. system used affects the 4.5 megacycle amplitude curve; here it will be assumed to be a "peak" type operating on low frequency video. The 4.5 megacycle amplitude can be determined for the two conditions of most interest.

1. When the picture carrier is the strong signal at the 2nd detector.

2. When the sound carrier is the strong signal at the 2nd detector.

Condition 1 is the case for normal tuning and in this case the 4.5 megacycle beat is proportional to the sound carrier level at the 2nd detector. The level of 4.5 megacycle sound with the sound in the trap is taken as reference. The procedure of computing the sound level below picture level is straightforward and Figure 3b shows this level plotted as a solid line versus oscillator frequency. The exact peak of the curve is not particularly important. It has an effect on pull-in time and hold-in range, but in practice the "peak" can be limited to advantage.

Condition 2 is the case when the picture carrier is in the region of the adjacent channel sound trap. This part of the curve is important because the level of the 4.5 mc signal, when the oscillator is tuned to put the picture carrier in the adjacent sound trap, limits the pull-in range of the basic system.

Figure 4 shows the μ curve plotted on a linear scale. The absolute level of the μ curve is easily obtained from the second detector level and cascade amplification of the video and sound amplifier. The closed loop operating level given by the intersection of the μ and β curves is 10 V rms of the 4.5 megacycle signal. The curve has a slope of approximately 100 V/Mc at the operating point. The slope is even greater for higher frequencies, which is in the preferred direction because it raises the gain of the system when the oscillator frequency increases. This controls the sound in the picture which is what normally limits the high frequency end of the oscillator drift. The low drift will generally be limited by the level of 4.5 megacycles amplitude dropping below the level required for proper operation of the F.M. sound detector.

As shown in Figure 4, if the oscillator drifted 1 megacycle lower in frequency the locked up frequency change would be 100 K.C. If the oscillator drifted 1 megacycle higher in frequency the locked up frequency change would be 50 K.C. However, for the high drift case, if the loop were opened and closed again the closed loop frequency may be improper because there are now two stable lock-up conditions. One is correct and the other one is several megacycles too high in frequency. If the adjacent channel sound trap were deeper, the pull-in range of the system would be limited to approximately 1.5 megacycles. Also, the β curve gain is limited if maximum pull-in range is required.

INCREASING PULL-IN RANGE

The pull-in range of the basic system can be extended by two methods. Broadly speaking, one method involves recognizing when the sound carrier enters the I.F. passband and the other when the picture carrier enters the I.F. passband.

The first method uses the D.C. signal developed at the second detector by the action of the "peak" acting A.G.C. system. The peak level of the detected signal is held constant so that on a picture modulated signal the average value is less than peak depending upon the pciture content. For typical pictures the average value of the D.C. is approximately one-half the peak value. On an all black picture the D.C. is about three-fourths peak. The sound carrier having no amplitude modulation has the average D.C. equal to the peak value. This change of D.C. at the second detector versus oscillator frequency can be used to extend the pull-in range.

One method of using this signal is to add it to the 4.5 megacycle amplitude curve as shown in Figure 5. The low level portion of the second detector voltage is suppressed. It is desirable to suppress the low level contribution because it is a function of picture content and would vary the oscillator frequency somewhat if allowed to be effective. The composite μ curve has several megacycles more pull-in range than the basic system. For a strong signal the pull-in range is extended even more. This μ curve requires off-setting the oscillator frequency and in the absence of a signal the oscillator frequency is higher than for correct tuning. It is possible to have a system using the 4.5 megacycle amplitude with a non-offset oscillator frequency.

The principle involved in getting a non-offset system is to obtain a signal that exists when the oscillator is tuned correctly or lower in frequency but disappears when the oscillator is tuned higher than correct frequency. One signal of this type is the amplitude of the stripped sync. The advantage of stripped sync. over the composite video signal is the unvarying level with program material. The sync. signal disappears at low oscillator frequencies because the signal level eventually falls below the A.G.C. threshold. On the high frequency side the sound carrier becomes the A.G.C. controlled signal at the second detector and the sync. level decreases. The sync. signal can be used by rectifying the signal to give the opposite polarity from the 4.5 megacycle signal amplitude. Adding these signals gives the composite μ curve shown in Figure 6.

The oscillator frequency can be offset low to center the pull-in range and this would be advantageous on weak signals since the oscillator would be nominally nearer correct for fringe tuning.

This system has several megacycles more pull-in range than the basic system. If more pull-in range is desired this system can be further combined with the second detector D.C. signal to give an extremely wide pull-in range system.

In addition to wide pull-in range, the systems described provide control over picture and sound by the effective A.G.C. action provided on the intercarrier sound signal.

920 KC BEAT CONTROL

On color broadcasts the beat between the sound and chroma signals at the receiver second detector influences receiver tuning. For this 920 Kc beat to be invisible at normal viewing distances, the combined sound and chroma attenuation below picture level must be 30 db or more at the second detector. Attenuation in excess of the 30 db reference level will be called "beat reserve".

With the A.F.C. holding the sound level constant and the keyed A.G.C. holding the picture level constant, the chroma relative to picture level determines the 920 Kc beat variation.

The chroma relative to the picture level varies with oscillator tuning so the 920 Kc beat amplitude must be figured knowing the transmission changes in sound and chroma level and the I.F. curve. Two procedures are shown. One involves measuring the average slope at the picture, sound and chroma regions and calculating the 920 Kc attenuations. The other involves measuring the attenuation levels point by point and computing the 920 Kc attenuation. The two methods show close agreement over nominal variations and the first method is much faster. Figure 7 shows the I.F. curve and slope equations for two different transmission changes.

The variation of 920 Kc attenuation with sound level changes is shown in Case I in terms of initial attenuation and the I.F. slopes. Case II shows the 920 Kc attenuation for a linear passband tilt. These are simple linear equations and are sufficiently accurate for most purposes.

The beat reserve versus sound change plotted in Figure 8 shows that with an increased level of sound carrier the beat reserve increases. This is due to the A.F.C. tuning action which maintains the sound level at the second detector constant. The oscillator frequency decreases for increasing sound carrier level and raises the picture level on the I.F. passband and lowers the chroma and sound levels. The equation shows that a 17 db downward sound change gives 0 db beat reserve. The actual reserve, taken point for point from the I.F. curve, is the dotted curve and shows a 15 db downward sound change gives 0 db of reserve.

In the case of a linear tilt the chroma level changes .8 db for every 1.0 db of sound level change, Figure 9 shows that in this case the upward tilt causes a decrease in beat reserve. The actual reserve taken point by point correlates closely with the reserve calculated from the I.F. passband slopes.

If the chroma level changes .4 db for every 1 db of sound level change, the beat reserve would remain almost constant with tilt.

The typical black and white television set design does not have nearly as much beat reserve under tilts as shown here, unless tuned to seriously degrade the picture detail. Double detector color sets must be tuned very accurately to approach the performance shown here and of course do not automatically adjust for changing situations.

With the I.F. passband shown it would be possible to build a single detector color set and compensate the I.F. chroma slope at video. With A.F.C. this looks attractive, without A.F.C. the criticalness of tuning would be a problem.

OSCILLATOR TUNING VERSUS SOUND CHANGE

Figure 10 shows the oscillator frequency change for a change of sound level. Essentially this is a measure of the combined picture and sound I.F. slopes. The data plotted assumes a high B gain and shows that a 10 db sound tilt causes a 100 Kc change in oscillator frequency.

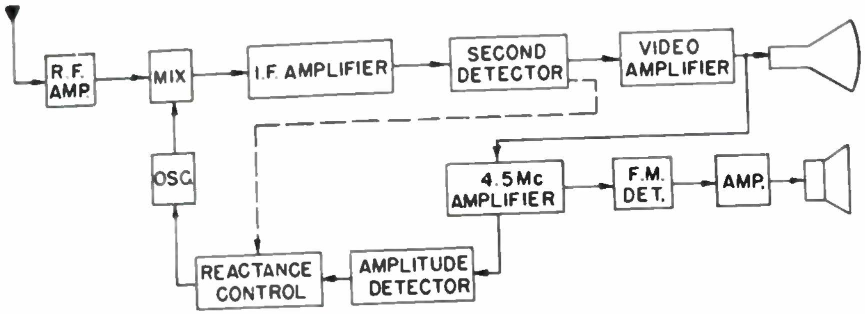
If the sound change is due to a transmission tilt, the frequency change is in the direction to provide video compensation. The picture carrier moves down on the I.F. passband to give high frequency boost for downward tilts of sound and provides high frequency attenuation for upward tilts,

CONCLUSION

An A.F.C. system particularly adapted to Television Receivers has been shown. The operation of the system provides effective A.G.C. action on the intercarrier sound level. This controls the beats between the picture harmonics and the sound carrier and offers a tuning condition closely allied with subjective manual tuning. This tuning is automatically maintained in the presence of sound level changes in the transmission or in the receiver.

Two methods of increasing the pull-in range were considered and signals exemplifying each were derived. These wide range systems are particularly suited to UHF television receivers because of the increased oscillator drift.

On color broadcasts the 920 Kc beat is indirectly controlled and severe transmission tilts can be compensated for. A single detector color receiver with this type of A.F.C. is attractive both from an economic and operational standpoint.



INTERCARRIER SOUND AUTOMATIC FREQUENCY CONTROL SYSTEM

Fig. 1

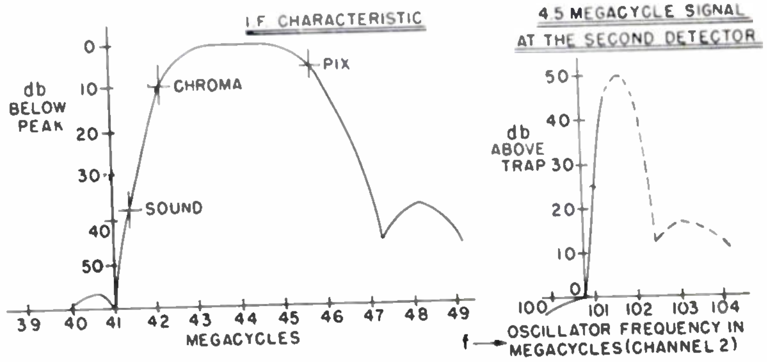
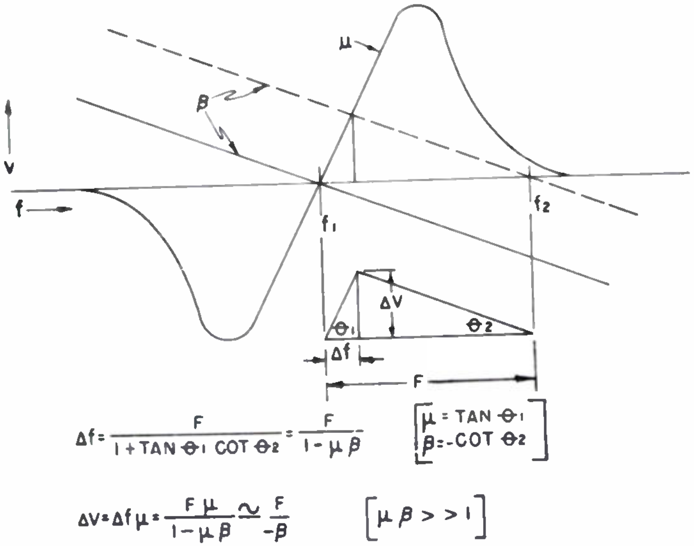


Fig. 3



BASIC THEORY

Fig. 2

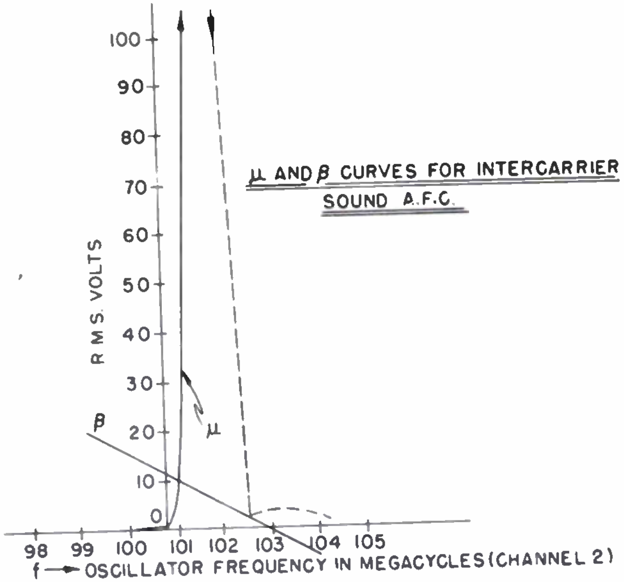


Fig. 4

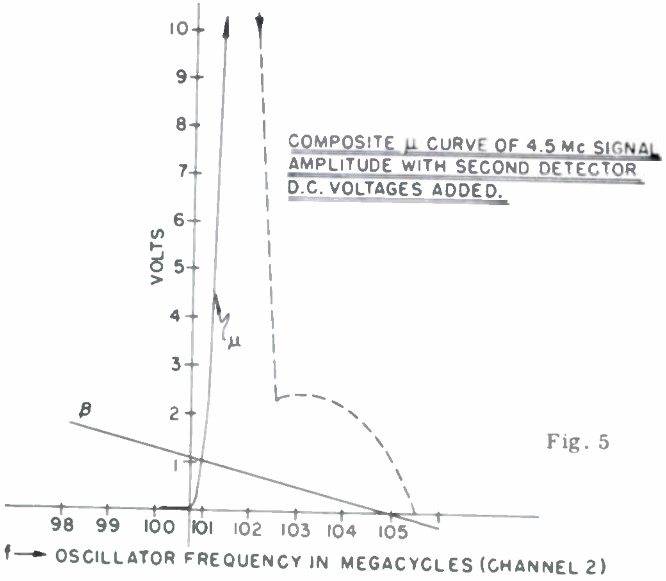


Fig. 5

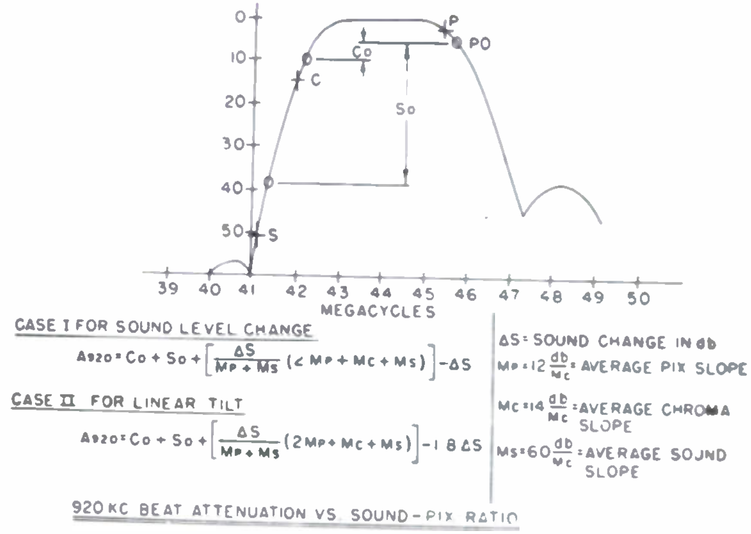


Fig. 7

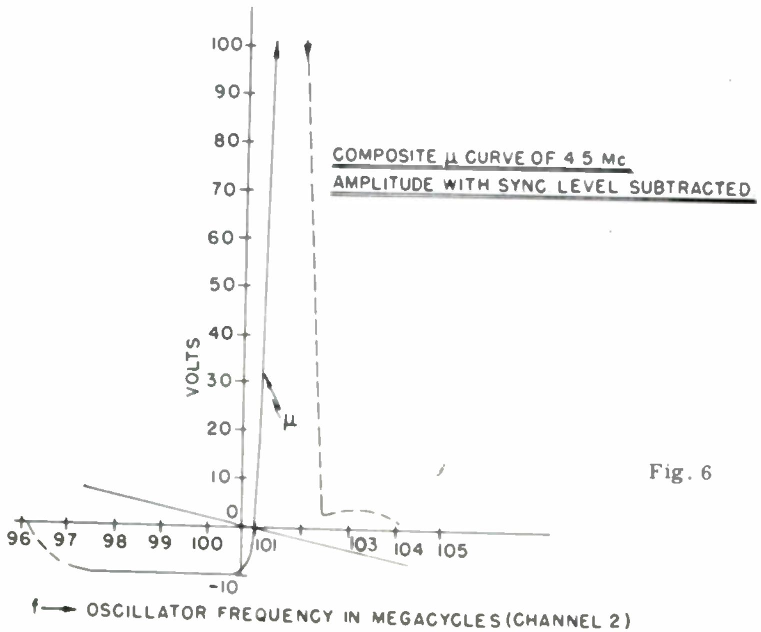


Fig. 6

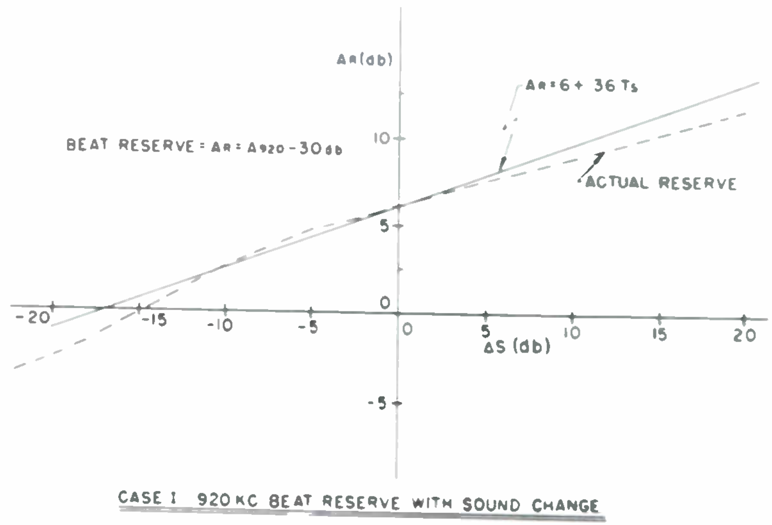


Fig. 8

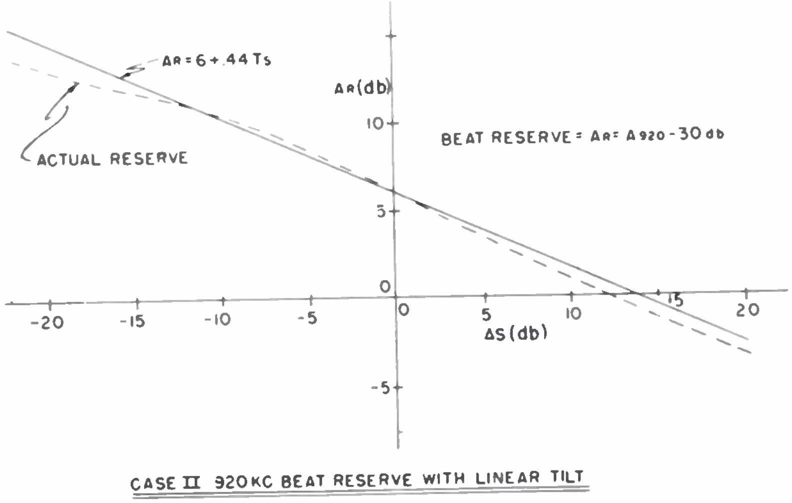

AR = 6 + .44 Ts
ACTUAL RESERVE
AR(db)
BEAT RESERVE = AR = A920 - 30 db
ΔS (db)
CASE II 920 KC BEAT RESERVE WITH LINEAR TILT

Fig. 9

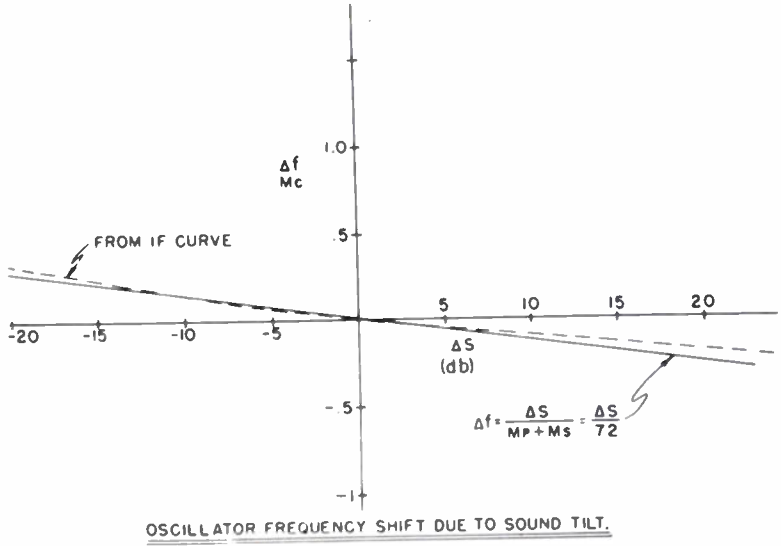

Δf
Mc
FROM IF CURVE
ΔS
(db)
Δf = ΔS / (MP + MS) = ΔS / 72
OSCILLATOR FREQUENCY SHIFT DUE TO SOUND TILT.

Fig. 10